Deadbeat Flux Vector Control as a One Single Control Law Operating in the Linear, Overmodulation, and Six-Step Regions With Time-Optimal Torque Control

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ABSTRACT This article proposes an enhanced version of the deadbeat flux vector controller (DBFC) as a one single control law that can operate in the entire torque–speed plane. The operation at any feasible modulation index can be accomplished by adequate determination of the flux trajectories at the different operating regions (e.g., PWM, overmodulation (I and II), and six-step). Continuous and seamless transition between the four operating regions is guaranteed, where the modulation index changes linearly with speed between PWM and six-step (without abrupt change in torque or acoustic problems). With the proposed strategy, undesirable torque dynamics, stability problems, and increased computational efforts associated with using multiple control laws are avoided. The transient performance of DBFC at the maximum voltage limit is analyzed in detail in the flux plane. A time-optimal torque control algorithm is developed to achieve the fastest possible torque dynamics and to considerably reduce the settling time, without the use of a voltage margin. The torque can be controlled with high accuracy and high robustness to machine parameter variations. With DBFC, no tradeoff between good steady state six-step behavior and good transient performance is needed due to the decoupling of switching and calculation frequencies. The proposed DBFC controller offers valuable features, and it is simple to implement. Simulation and experimental results are provided to validate the proposed control algorithm, which is implemented on an automotive microcontroller with a high-power/high-performance automotive traction machine.

INDEX TERMS Direct torque and flux control, flux trajectory, flux weakening, overmodulation (OVM), permanent magnet synchronous machine, six-step operation, time-optimal control, traction applications.

NOMENCLATURE

\( V, i, \psi \) Voltage, current, flux linkage.
\( L, R \) Inductance, resistance.
\( \psi_{pm} \) Permanent magnet flux linkage.
\( P \) Number of poles.
\( V_{dc} \) Dc-link voltage.
\( T_s \) Sampling period.
\( T_{em} \) Electromagnetic torque.
\( \omega_r \) Electrical angular speed of the rotor.
\( \omega_{rm} \) Mechanical angular speed of the rotor.
\( \psi_R \) Circumradius of the flux hexagon.
\( f_s \) Switching frequency.
\( f_1 \) Fundamental electrical frequency.
\( N_p \) Pulse number.
\( V_1 \) Peak magnitude of the fundamental phase voltage.
\( MI \) Voltage modulation index.
\( R_\phi \) Average flux magnitude normalized to \( \psi_R \).
\( f_{dq} \) Synchronous reference frame.

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Stationary reference frame. 
\( f_{\alpha\beta} \)  
Angle referred to the \( \alpha \)-axis. 
\( \theta \)  
Angle referred to the \( \delta \)-axis. 
\( \delta \)  
Operating point \( (x \in \mathbb{N}) \). 
DBFC  
Deadbeat flux vector control. 
OVM  
Overmodulation. 
PWM  
Pulsewidth modulation. 
TOC  
Time-optimal control. 
MTPA/F  
Maximum torque per ampere/flux. 
IPMSM  
Interior permanent magnet synchronous machine. 
(*)  
Estimated parameter. 
(*)  
Commanded variable.

I. INTRODUCTION
Flux weakening control is a prerequisite for automotive traction drive applications due to the limited dc-link voltage [1]. Six-step is a modulation strategy that increases the power density of the inverter by delivering the maximum possible fundamental voltage \( V_{1\text{-max}} = 2V_{dc}/\pi \) without any modification of the hardware setup [2]. This leads to an enhanced torque capability of the machine, an increased operating range of the constant-torque and constant-apparent-power regions, and a reduction of the current needed for flux weakening [3]. Moreover, the switching frequency \( f_1 \) becomes synchronized with the fundamental electrical frequency \( f_1 \), leading to reduced switching losses [4]. The six-step operation is achieved at the cost of increasing the torque ripple and the current harmonics in the dc link [1].

Fig. 1(a) shows the machine’s capability curve in the different operating regions [i.e., linear, overmodulation (OVM), and six-step]. The rated torque is produced by operating at the maximum torque per ampere (MTPA) flux while staying at the inverter’s current limit. The OVM region extends the constant torque region and avoids discontinuities in the modulation index. Fig. 1(b) and (c) shows the zoomed views of Fig. 1(a), at the base speed, for continuous and discontinuous changes in the modulation index. The discontinuous change in the modulation index leads to undesirable dynamics and may cause stability problems. Moreover, the abrupt increase in the modulation index (switching to six-step in the OVM region) increases the flux. The deviation from the MTPA flux reduces the torque and power production due to the current limitation, as seen in Fig. 1(d). Therefore, a transition region, from pulsedwidth modulation (PWM) to six-step, is necessary for maximum power and maximum efficiency operation, and for stability issues.

For optimum operation, the flux must have a constant average magnitude \( (|\psi_{\alpha\beta}|_{avg} = \psi_{\text{MTPA}}) \) in the constant torque region (Linear and OVM regions). The six-step operation takes place at the maximum voltage limit (end of OVM), where the flux weakening operation is active.

A. STATE-OF-THE-ART CONTROLLERS OPERATING IN THE OVM REGION INCLUDING SIX-STEP
Several control strategies have been presented in the literature for traction drive applications. These strategies differ by the range of the modulation index they cover. For a better description of the different modulation regions, the amplitude modulation index (MI) is defined in (1) as the ratio of the peak magnitude of the fundamental phase voltage \( V_1 \) to the base voltage \( V_{base} \), and indicates the voltage utilization level of the inverter. Different definitions of \( V_{base} \) are shown in Table 1. In this research, \( V_{base} \) is set to \( V_{dc}/\sqrt{3} \). Accordingly, the modulation index changes in the range of \( \text{MI} = [0, \frac{2\sqrt{3}}{\pi}] \)

\[
\text{MI} = \frac{V_1}{V_{base}} = \frac{V_1}{V_{dc}/\sqrt{3}}.
\]  

Fig. 2(a) and (c) shows the locus of all the voltage vectors in the linear voltage region (also referred to as the PWM region) and six-step, respectively. The OVM region [see Fig. 2(b)] is the set of operating points that lie between the linear region and six-step. In order to cover all the operating regions in Fig. 1, the control law should be able to operate at any modulation index between \( \text{MI} = 0 \) and \( \text{MI} = 1.103 \), where any voltage vector within the hexagonal voltage limit can be produced [see Fig. 2(d)]. To achieve a six-step operation (and other synchronous pulse patterns) with high dynamic performance, the controller should be able to influence the switching and pulse patterns actively. The switching and calculation/sampling frequency must be decoupled.

The current vector control (CVC) [5] and the deadbeat-direct torque and flux control (DB-DTFC) [6] were originally designed to operate in the linear voltage region (MI = [0, 1]). Other control strategies were designed for six-step operation (MI = 1.103), such as the voltage angle torque control.

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(VATC) [7], the direct self-control (DSC) [8], and the dead-beat flux vector control (DBFC) in [9]. Several possibilities exist to operate in all the regions of Fig. 1. Among them are the following.

1) **Option 1:** Employing a controller exclusively designed for the linear voltage region, and change the control law to cover the OVM and/or six-step, according to the operating conditions.

2) **Option 2:** Adapting an existing controller, originally operating in the PWM mode, to operate in the OVM and/or six-step, and vice versa.

A change in the control law and pulse patterns can cause undesirable torque dynamics (and possibly acoustic noise) and stability problems [7]. Moreover, using multiple control laws complicates the implementation and consumes additional computational and memory loads in the microcontroller. Selected state-of-the-art implementations are summarized as follows.

1) **VOLTAGE ANGLE TORQUE CONTROL (VATC)**

The voltage angle ($\theta_v$) is the only degree of freedom available in the six-step mode [see Fig. 2(c)] and can be employed for torque regulation [7], [10], [11], [12], [13], [14], [15]. The voltage angle can be controlled with a PI regulator [7], [10] or with a model-based design [11], [12]. Most of the model-based designs use the steady-state machine model for voltage angle calculation. The PI-regulator and the steady-state machine model methods inherently offer limited dynamic performance. A deeper analysis was made in [13] and [14] by using the transient motor model to improve the dynamic performance and achieve higher bandwidth for torque control. However, only the fundamental value of the voltage is considered in the analysis. Time-optimal voltage control was presented in [15], applying VATC. The results show excellent dynamic performance. However, the VATC methods are usually applied in six-step, and a change in the control law is needed to cover the different operating regions.

2) **CURRENT VECTOR CONTROL (CVC)**

Closed-loop current vector controllers were also employed to achieve six-step operation [16], [17], [18], [19], [20], [21], [22]. The PI-regulator is initially designed to control the fundamental current vector. The inherent low-order harmonics in the feedback current during OVM (including six-step) deteriorate the performance of the PI-regulator in steady state and transient operation and make six-step operation cumbersome [16]. Several attempts were made to provide instantaneous current control. In [17], a low-pass filter was used in the feedback current. However, this reduces the controller’s bandwidth and thus, the dynamic stiffness. In [18] and [19], the current setpoints were adjusted by a voltage error signal using a low-pass filter. However, a six-step operation could not be achieved in steady state.

Kwon et al. [21] presented a current vector controller that can operate in the entire torque–speed range and with enhanced dynamic performance. Nevertheless, the low-order current harmonics in the feedback path are not considered. Reference [21] does not address the integrator windup issue and the reliability of the controller operation in the OVM region [23].

Brosch et al. [22] presented a model predictive current vector controller that can operate in the linear, OVM, and six-step regions. A harmonic reference generator is used to control the instantaneous current vector by estimating the current harmonics and adjusting the reference currents accordingly. Brosch et al. [22] used a parameter-dependent open-loop solution for torque regulation, which relies on the accuracy of the machine model. A steady-state error in torque (mainly due to the error in the $d$-axis current) can be observed. The main concern of this implementation is the high computational complexity and memory burden needed.

3) **DIRECT SELF-CONTROL (DSC)**

DSC is a six-step controller known for its simplicity and robustness against voltage disturbances [8], [24]. DSC was originally designed for high-power applications (e.g., railway traction applications). It directly controls the instantaneous values of stator flux and torque by selecting the optimum inverter switching functions ($S_a$, $S_b$, and $S_c$) with the help of hysteresis controllers. DSC has a very high control bandwidth (closed loop) and, thus, high disturbance rejection capability. Moreover, folding the corners of the hexagonal flux trajectory enables 18-corner 3-pulse synchronous switching [24]. The main drawback of DSC is the unreliability in the low-speed
region (e.g., 30% of the base speed), without additional modifications, due to the hexagonal flux trajectory [24], [25]. The need for multiple control laws increases the sophistication of the control system and can lead to undesirable torque dynamics and stability problems. Moreover, the digital implementation of the hysteresis flux controller is challenging since it requires very fast calculation tasks.

4) STATOR FLUX-BASED SPACE-VECTOR MODULATION

Tripathi et al. [26] presented a stator flux-based space-vector modulation that can operate at any feasible modulation index. It is often referred to as a form of direct torque control space-vector modulation (DTC-SVM) [25]. The flux at time instant \( k + 1 \) is estimated with the simple linear prediction and the flux error vector can track its command by applying the corresponding volt-sec reference. The on-times of the active voltage vectors of the inverter at the corresponding hexagon sector are calculated by the space vector modulation method.

In this implementation, the reference signal has always a circular trajectory as opposed to the real signal due to the inherent voltage manipulation in the OVM region. This inconsistency in the flux control due to the voltage manipulation can deteriorate the torque control ability in the OVM region and may lead to instability problems. In addition, this method has a complex implementation as discussed in [25], [27], and [28].

5) DEADBEAT FLUX VECTOR CONTROL (DBFC)

As opposed to the DTC-SVM method in [26], commanding a feasible flux during six-step operation can avoid unnecessary complications, and can enhance the controller performance. The deadbeat flux vector controller (DBFC) is developed in [9] as a hexagonal-flux-trajectory-based six-step controller. DBFC controls the instantaneous value of the stator flux vector (magnitude and angle) in closed loop. By following a predefined hexagonal trajectory, DBFC can achieve six-step operation with high-bandwidth torque control capability, even if the ratio of calculation frequency to electrical frequency is not an integer.

The DBFC controller in [9], [29], [30], and [31] is an efficient control law that is designed to achieve reliable six-step operation. However, there are some shortcomings in the state-of-the-art implementation, given as follows:

1) DBFC is exclusively designed for six-step operation. The research in [9] uses DB-DTFC in the linear voltage region and changes the control law to DBFC in flux weakening. Here, the OVM region is not covered.

2) The torque controller in [9] uses a PI regulator to determine the flux angle setpoint. This method has a very good steady-state accuracy. However, a limited PI regulator bandwidth must be used, at the cost of worsening the dynamic performance, due to the high torque harmonics in six-step. An attempt to improve the torque controller was later presented in [29] and [30]. Here, the torque is controlled based on its average value during each 60 electrical degrees. The averaged torque is calculated with the help of an average torque observer, and used with a parameter-dependent inverse machine model to calculate the flux angle command. This solution complicates the implementation and adds undesirable computational burdens.

3) For infeasible torque commands, the change in torque is restricted by the hexagonal voltage limit. Without employing a decent methodology to guide the voltage vector (or flux vector) in transient operation, the flux follows a random trajectory in the transient mode, leading to increased settling time. Petit et al. [29] proposed a flux hexagon size adjustment during transient operation to improve the transient performance. The results show an improvement in dynamic performance. However, further development can be made to achieve time-optimal torque control during transient operation.

B. ARTICLE STRUCTURE

This article extends the research presented in [32] with deeper insights into the transition between the different operating regions, the various flux trajectories, and the torque control methods. Moreover, a time-optimal torque controller is developed for DBFC and analyzed in detail to guarantee an excellent dynamic performance. The current harmonics are analyzed in the six-step mode. The proposed control algorithm is implemented on an automotive microcontroller with a high-power/high-performance automotive traction machine (IPMSM). The rest of this article is organized as follows.

1) Section II reviews the specific characteristics of the different operating regions of the inverter.
2) Section III presents an overview of the state-of-the-art DBFC.
3) Section IV discusses the various flux trajectories in the different operating regions.
4) Section V shows a comparison between different torque control methods for DBFC.
5) Section VI presents a detailed explanation of a simple methodology for a smooth transition between the different operating regions.
6) Section VII analyses the DBFC operation in transient mode and designs a time-optimal torque control methodology.
7) Section VIII shows the block diagram of the proposed DBFC controller.
8) Section IX presents the experimental results.
9) Section X discusses the key conclusions.

II. OPERATING REGIONS OF THE INVERTER

The inverter can synthesize a fundamental voltage that varies linearly with the reference voltage for a certain range, known as the linear modulation region. This range can produce a fundamental voltage vector with a magnitude up to \( V_1 = V_{dc}/\sqrt{3} \) (neglecting the voltage drop of the inverter) by applying specific modulation strategies, such as the space vector pulsewidth modulation (SVPWM). The fundamental phase voltage can be further increased by entering the nonlinear
OVM region, in which the fundamental phase voltage changes nonlinearly with the reference voltage due to the hexagonal voltage limit. The six-step mode is the upper boundary of OVM \( V_l = 2V_{dc}/\pi \). In the linear region, the harmonics are located in the high-frequency range around the switching frequency and its integer multiples [33]. By going through the OVM region, the effective switching frequency decreases and the distortion in the phase voltage increases gradually due to the lower frequency harmonics.

Fig. 3 shows the relation between the desired and the obtained modulation indices in the linear and the OVM regions. The ratio of MI/MI* is equal to unity in the linear voltage region (the inverter has a unity gain). In the OVM region, MI changes nonlinearly with MI* and saturates at 1.103 (six-step). This change depends on the OVM strategy used [34]. An adequate modulation strategy is needed to compensate for the voltage nonlinearity in the OVM region and guarantee a linear transition to six-step.

In fact, the OVM region can be divided into two subregions (OVM I and OVM II) based on the way the reference voltage vector is adjusted. In OVM I (MI \( \in [1, 1.05] \)), the magnitude of the voltage vector is adjusted and then saturated by the hexagonal voltage limit while keeping the phase unchanged. To further increase the modulation index in OVM II (MI \( \in [1.05, 1.103] \)), the magnitude and the angle of the voltage reference must be adjusted in each hexagon sector. Different OVM strategies exist to achieve this goal [2], [35]. These strategies are essential to determine the quality of the transition between PWM and six-step and the reliability of the controller.

III. OVERVIEW OF THE STATE-OF-THE-ART DBFC

The block diagram of DBFC [9] is shown in Fig. 4. The reference voltage vector is calculated from the deadbeat flux regulator \( [\psi^*_{a\beta}(k) = \psi_{a\beta}(k + 1)] \) according to (2) using the flux vector command \( (F^*_{a\beta}) \) and the estimated flux \( \hat{\psi}_{a\beta} \) in the stationary reference frame \( f_{a\beta} \). The flux angle is employed for torque control. For a given flux angle, the flux magnitude command is obtained from a predefined hexagonal trajectory. The flux hexagon sides are calculated from the available dc-link voltage, the electrical angular speed, and the stator resistance. A voltage selection block (red) is used to guarantee an operation at the corners of the voltage hexagon (six-step). Here, the small voltage angle errors are ignored \( (\Delta \theta_e < \pi/6 \text{ rad}) \). This helps in achieving an inherent disturbance rejection with DBFC for small voltage deviations.

The temperature effects are considered in the estimated permanent magnet flux linkage and the estimated stator resistance needed to calculate the flux and voltage setpoints. A closed-loop discrete-time Gopinath flux linkage observer is used in this implementation [36]. It consists of two models. The current model is based on the machine’s magnetic model and is dominant in the low-speed region. The voltage model is based on the stator voltage equation and is robust to changes in \( L_d, L_q, \text{ and } \psi_{pm} \). The machine’s terminal voltage is estimated using the voltage setpoint and the inverter model. The sensitivity of the voltage model to voltage disturbances (e.g., resistance variation and inaccurate inverter model) is less significant at high speeds where the back electromotive force dominates the terminal voltage.

\[
v^*_{a\beta}(k) = \frac{\psi^*_{a\beta}(k) - \hat{\psi}_{a\beta}(k)}{T_s} + R_s i_{a\beta}(k).
\]  
(2)

IV. FLUX TRAJECTORIES IN THE DIFFERENT OPERATING REGIONS

The DBFC controller was originally designed to operate in six-step by following a hexagonal flux trajectory. Since DBFC controls the instantaneous value of the stator flux, it is theoretically possible for the flux vector to follow any feasible flux trajectory using DBFC.

By neglecting the stator resistance in (2), the stator voltage can be approximated to the rate of change of the flux as given by (3). Therefore, the commanded volt-sec vector guides the flux vector and determines the flux trajectory. Fig. 5 is a graphical representation of (3) in the volt-sec plane (stationary reference frame). Here, the scaled voltage vectors (by \( T_s \)) and flux vectors can be plotted together in one plane. Accordingly, the different voltage regions in Fig. 2 can be mapped into the volt-sec plane to determine the corresponding flux trajectories.

\[
v^*_{a\beta}(k)T_s = \Delta \psi_{a\beta}(k).
\]  
(3)

The steady-state flux trajectory changes according to the operating regions in Fig. 3. In the linear voltage region (PWM), the flux magnitude is constant, and the flux vector
follows a circular path. On the other hand, the flux trajectory has a hexagonal shape in six-step. In the transition region from PWM to six-step, the flux trajectory is either circular or a combination of a circle and a hexagon, as will be demonstrated later. Accordingly, the first step in designing a flux-trajectory-based DTC controller, which can operate in the entire torque–speed range, is to define the flux trajectories in the different operating regions. After determining the different flux trajectories, the DBFC control law can be extended to cover the linear and OVM regions.

A. FLUX TRAJECTORY IN THE LINEAR VOLTAGE REGION

Below the base speed, the flux magnitude may have a constant value since the machine operates within the voltage limit. For a given torque command, the flux magnitude command is selected based on the MTPA/loss to minimize the copper/total losses in the machine/drive system. Fig. 6 depicts the flux setpoint as a function of the desired torque, and its cubic function approximation.

$$y = ax^3 + bx^2 + cx + d$$

where $x$ and $y$ are the torque and the flux setpoints, respectively. The effect of rotor temperature on the MTPA flux can be considered by simply adapting the term $d$. Therefore, the optimum flux setpoint in the linear voltage region can be determined without any complicated calculations. The rotor temperature can be estimated, or measured with a wireless battery-free sensor.

B. FLUX TRAJECTORY IN THE SIX-STEP MODE

The determination of the flux hexagon size is essential for a proper six-step operation. Fig. 7 depicts the volt-sec hexagon and the flux hexagon in six-step. Here, the volt-sec vector has the magnitude of $\frac{2\sqrt{3}}{3}V_{dc}T_s$. According to the rotor electrical angular speed ($\omega_r$) and the available dc-link voltage ($V_{dc}$), the length of a hexagon side can be calculated as given by (4) and [9]

$$\psi_{R_{Rs=0}} = \frac{2\pi}{\omega_r} \left( \frac{2}{3} V_{dc} T_s \right) = \frac{2\pi V_{dc}}{9 \omega_r}.$$  

(4)

In fact, the flux trajectory in six-step is affected by the stator resistance in (2). Due to the resistive voltage drop vector ($R_s i_{\alpha\beta}$), the real flux trajectory is tilted, the flux magnitude decreases (in motoring mode), and the hexagon sides are not straight lines. These effects are discussed in more detail in [8], [9], and [37]. To simplify the implementation, the flux trajectory is modeled as a regular hexagon with straight sides. This assumption was found to be valid in simulation and experiments. The real hexagonal flux trajectory (yellow) is depicted in Fig. 8 and its size can be determined by (5).

$$\psi_{R_{Rs=0}} = \frac{2\pi}{\omega_r} \left( \frac{2}{3} V_{dc} T_s \right) = \frac{2\pi V_{dc}}{9 \omega_r}.$$  

(5)
C. FLUX TRAJECTORIES IN THE OVM REGION

1) FLUX TRAJECTORY IN OVM I

From the voltage point of view, the lower boundary of OVM I is reached when the circular voltage trajectory touches the voltage hexagon sides. The upper boundary is reached when the voltage trajectory is superposed with the hexagon’s boundary. These two boundaries are shown in Fig. 9 for steady-state operation. At the lower boundary [see Fig. 9(a)], the flux is a regular circle with a radius \( r_{\text{max}} \approx \frac{V_{\text{dc}}}{\sqrt{3} \omega y} \). The voltage vector follows a hexagonal trajectory at the upper boundary, as shown in Fig. 9(b). In this example, and by looking at (3), it can be concluded that the flux at this operating point follows an elliptic trajectory with \( r_{\beta}/r_{\alpha} \approx 0.997 \), where \( r_{\alpha} \) is the semimajor axis and \( r_{\beta} \) is the semiminor axis of the ellipse. The reason for the elliptic flux is that \( u_{\text{dc}-\text{max}} > u_{\beta-\text{max}} \). For any modulation index between \( MI = 1 \) and \( MI = 1.05 \), \( u_{\alpha-\text{max}} \) will be closer to \( u_{\beta-\text{max}} \) and the ratio \( r_{\beta}/r_{\alpha} \) will be closer to unity. The flux trajectory in OVM I can be approximated to a regular circle since even for the worst-case scenario, the error between \( r_{\alpha} \) and \( r_{\beta} \) is just 0.3%. This error might be even smaller than a possible error in the estimated resistive voltage drop \( (|\dot{R}_{Ia}\beta T_{s}|) \). Making this assumption simplifies the implementation of the control algorithm since only one flux trajectory is needed for the linear voltage region and OVM I. The operating region characterized by \( MI = [0, 1.05] \) is defined as the linear flux region.

2) FLUX TRAJECTORY IN OVM II

The flux trajectories in the linear voltage region \( (MI = [0, 1]) \), in OVM I \( (MI = [1, 1.05]) \), and in six-step \( (MI = 1.103) \) are defined in the previous sections. To cover the entire torque–speed plane and guarantee a continuous change in the modulation index, a flux trajectory in OVM II \( (MI = [1.05, 1.103]) \) must be identified. Since the flux trajectories in the linear flux region and six-step are circular and hexagonal, respectively, the flux trajectory in the OVM II is a combination of a circle and a hexagon (hybrid trajectory). Here, the flux trajectory can be changed gradually from a circle to a hexagon and the controller becomes able to produce any modulation index in the range \( MI = [1.05, 1.103] \).

Gaona et al. [37] proposed a simple way to achieve a hybrid flux trajectory in OVM II. The well-known minimum voltage phase error method can increase the modulation index from 1 to 1.05 by keeping the voltage phase unchanged and saturating the voltage magnitude to the hexagonal boundary when the voltage vector is outside the hexagon. With this method, the voltage trajectory can be gradually changed from a circle \( (MI = 1) \) to a continuous hexagon \( (MI = 1.05) \). The minimum flux phase error method follows the same concept as the minimum voltage phase error method, but it is applied to the flux vector instead of the voltage vector. Fig. 10 shows an example of the minimum flux phase error method. In contrast to the circular flux trajectory (yellow) in Fig. 10(a), which is feasible at all angles, the circular flux trajectory (red) in Fig. 10(c) is saturated by the flux hexagon (blue). In Fig. 10(b), and by applying the minimum flux phase error, the feasible flux trajectory is a combination of a circle and a hexagon. Consequently, the flux trajectory can be continuously changed from circular to hexagonal, and the modulation index can be simply increased from 1.05 to 1.103.

D. FLUX TRAJECTORIES FOR SYNCHRONOUS PWM

For high-power drives operating at high speeds, the pulse number \( (N_p) \) becomes low enough and the synchronous PWM is preferred over other PWM methods to minimize the total harmonic distortion of the stator current [24], [38]. The pulse number is defined as the ratio of the switching frequency to the fundamental frequency. For increasing stator frequency, pulse patterns with \( N_p = 9, 7, 5, 3, 1 \) can be used. A pulse number of \( N_p = 1 \) is equivalent to six-step operation. For a given pulse number, there exists a unique pattern known as the optimal pulse pattern (OPP) which produces a specific modulation index while minimizing the total harmonic distortion of the stator current.

By applying synchronous optimal modulation, the flux trajectory changes according to \( N_p \) (see Fig. 5). The regular hexagonal flux trajectory corresponds to \( N_p = 1 \) and has the highest possible average flux (for a given speed) and the lowest possible number of switching events per electrical revolution. However, the hexagonal flux trajectory inherently creates higher low-order harmonics compared to an operation in the linear region, OVM region, and OPP with \( N_p > 1 \). By folding the hexagon corners, as depicted in Fig. 11, a synchronous PWM is possible with \( N_p = 3 \). This trajectory is known as the 18-corner flux trajectory [24]. The flux trajectory for \( N_p = 5 \) can be found in [38].

For \( N_p = 3 \), the distance from the center of the flux plane to a folded corner \( (\psi_{\text{p-f}}) \) is variable and defined by (6). The variable \( K_f \) is an additional degree of freedom that can be used...
to achieve a certain optimization goal (e.g., elimination of the sixth harmonic in the dc-link current) [24].

DBFC can track any feasible flux trajectory. Therefore, the synchronous optimal PWM with variable $N_p$ can be achieved with DBFC by controlling predefined flux trajectories, such as the 18-corner trajectory, instead of the precalculated pulse patterns as in [38]. The optimal flux trajectories are simple to implement with DBFC. However, testing and analyzing these trajectories is outside the scope of this research.

$$K_f = \frac{\psi_{R-f}}{\psi_R}$$  \hspace{1cm} (6)

V. TORQUE CONTROL METHODS WITH DBFC

A flux trajectory is the locus of all possible flux-vector commands needed to achieve a certain modulation index. The flux command ($\psi_{a\beta}^* = |\psi_{a\beta}| e^{j \delta_{a\beta}^*}$) is defined by a flux angle $\delta_{a\beta}^*$ and a flux magnitude $|\psi_{a\beta}|$. The flux angle is an additional degree of freedom employed for torque control. The flux magnitude is obtained from the intersection of a predefined flux trajectory and a line passing through the origin with $\delta_{a\beta}$. After identifying the flux vector command, the voltage command can be calculated from (2).

The objective of this section is to design a reliable torque controller to properly determine the desired flux angle ($\delta_{a\beta}^*$) based on the torque command. The torque for IPMSMs can be expressed as a function of the flux magnitude ($|\psi_{a\beta}|$), the flux angle ($\delta_{dq}$), and the machine parameters ($L_d$, $L_q$, and $\psi_{pm}$) in the synchronous reference frame ($f_{dq}$), as given by (7).

$$T_{em} = \frac{3P}{4} \left( \frac{\psi_{pm} \sin(\delta_{dq})}{L_d} \cdot |\psi_{a\beta}| + \frac{L_d - L_q}{L_d L_q} \cdot \frac{\sin(2\delta_{dq})}{2} \cdot |\psi_{a\beta}|^2 \right).$$  \hspace{1cm} (7)

The torque is plotted in Fig. 12 as a function of $\delta_{dq}$ for a variable flux magnitude. Three possibilities to control the torque are discussed in this research as follows:

A. OPEN-LOOP SOLUTION

For given torque and flux magnitude commands, the flux angle can be calculated numerically by solving the inverse of (7). A 2-D lookup table (LUT) can be calculated offline to cover all the feasible operating points as depicted in Fig. 13(a). The saturation effects, obtained experimentally or through simulation by the finite-element analysis method, are considered in the LUT to improve its accuracy.
Fig. 13 shows the LUT data in the motoring mode for the desired operating range. The flux angle $\delta_{dq}^*$ is saturated to the MTPF-angle ($\delta_{\text{MTPF}}$) for any torque command higher than the MTPF-torque ($T_{\text{em}} - MTPF$). It should be noted that in addition to the global maximum and minimum points (MTPF points in the motoring and generating modes), the torque has a local maximum point and a local minimum point for each flux if $|\psi_{dq}| \geq \psi_{pm} L_d/(L_d - L_q)$ (see Fig. 12). To prevent the torque from going in the opposite direction, the flux angle is bounded to the region that has a monotonic change in torque. This explains why the flux angle in Fig. 14 is nonzero for high flux magnitudes ($T_{\text{em}} = 0 \text{ N.m}$).

The advantage of the LUT method is its simple implementation and the fact that the LUT output ($\delta_{dq}^*$) will be immediately updated for any change in the input signals ($T_{\text{em}}$, $|\psi_{dq}|$). However, the LUT relies on the accuracy of the machine parameters estimation, which can still change with aging, temperature, manufacturing tolerances, etc. [1].

**B. CLOSED-LOOP SOLUTION**

The flux angle setpoint can be obtained by employing a PI-regulator as the closed-loop torque controller [9], as shown in Fig. 13(b). With the help of the robust flux linkage observer in [39], the torque can be estimated with high accuracy and can be used as a feedback signal (average value). This method has a very good steady-state accuracy and high robustness against machine parameter variations. However, closing the torque loop with the estimated torque is challenging due to the high harmonics in torque during OVM and especially in six-step. This can deteriorate the performance of DBFC. Accordingly, a very limited PI regulator bandwidth must be used, which affects the dynamic performance of the torque controller and increases the settling time. The torque command (input signal) must be saturated to the MTPF limit (physical limit) to avoid a wind-up of the integrator.

**C. HYBRID SOLUTION**

The command tracking bandwidth of the solution in Fig. 13(b) can be improved as depicted in Fig. 13(c). Here, an initial guess (feedforward) of the flux angle setpoint is calculated with the help of a LUT ($\delta_{dq}^* - \text{LUT}$). The PI-regulator (feedback), with limited bandwidth, is used to adjust the flux angle command ($\delta_{dq}^* = \delta_{dq}^* - \text{LUT} + \delta_{dq}^* - \text{PI}$). The PI regulator reacts only when the LUT output is inaccurate. This approach has high steady-state accuracy, high dynamic performance, and is robust against machine parameter variations. The controller’s bandwidth should be increased if the PI regulator is used alone, or if the LUT data are wrong. In these cases, a tradeoff between the desired dynamic response and the stability of the controller must be made.

Table 2 summarizes the features of the three methods. It should be noted that the size of the LUT can be considerably reduced with the hybrid solution to reduce memory consumption. The interpolation errors, which are related to the table’s resolution, can be compensated with the PI regulator.

**D. SIMULATION RESULTS**

Fig. 15 shows the simulation results of the three methods for a step command in torque ($T_{\text{em}} = 0.85 \text{ p.u.}$), the speed is 0.29 p.u. In Fig. 15(a), the steady-state error in torque for the three methods is negligible. However, the PI regulator solution (red) has a limited bandwidth which worsens the dynamic performance. A negative torque is produced with the PI solution for small flux angles (see Fig. 12). For the second scenario, the estimated permanent magnet flux is intentionally increased by 10% ($\hat{\psi}_{pm} = 1.1 \psi_{pm}$). Here, the LUT solution (blue) produces 16% less torque since the calculated flux angle ($\delta_{dq}^* - \text{LUT}$) decreases. The PI solution has no steady-state error due to the robust voltage model of the flux observer. Thanks to the hybrid solution (yellow), the torque has an

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**TABLE 2 Summary of the Different Torque Control Methods**

| Torque controller | Type     | Dynamic response | Robustness to the parameter variations |
|-------------------|----------|------------------|----------------------------------------|
| LUT               | Open-loop| ✓                | ×                                      |
| PI                | Closed-loop|                | ✓                                      |
| Hybrid            | Closed-loop|                | ✓                                      |
The upper boundary of $\psi_R$ increases and reaches $0.866$ at $\omega = (8) |\psi|$, is obtained in six-step. In six-step, the flux hexagon is $\omega r = r VOLUME 3, 2022 = 0$ at the borders $\omega r = r \psi$. Simulation results showing the torque step responses ($T_m^* = 0 \rightarrow 0.85$ p.u.) in six-step for (a), (c) $\psi_{pm} = \psi_{pm}$ and (b), (d) $\psi_{pm} = 1.1 \psi_{pm}$ using the three torque control methods in Fig. 13. (c), (d) estimated flux angles. The speed is $0.29$ p.u. $T_s = 100$ $\mu$s. $K_r = 0.0001$ rad/N m. $K_\psi = 0.1$ rad/N m/s.

excellent transient response with negligible steady-state error even for wrong machine parameters.

VI. TRANSITION BETWEEN THE DIFFERENT OPERATING REGIONS

A. DETERMINATION OF THE FEASIBLE FLUX TRAJECTORIES

After defining the different flux trajectories which guarantee an operation at any feasible modulation index, and finding the flux angle for torque control, the objective now is to establish a continuous, seamless, and automatic transition between the four operating regions (linear, OVM I, OVM II, and six-step) by incorporating all the flux trajectories into one simple implementation. The variable $R_\psi$, given by (8), is the ratio of the average flux magnitude to $\psi_R$. The values of $R_\psi$ at the borders of the four regions are shown in Fig. 16.

$$R_\psi = \frac{|\psi_{avg}|}{\psi_R}. \quad (8)$$

For a given dc-link voltage and operating speed, the flux hexagon, calculated in (5), defines the maximum limit of any flux vector. Therefore, any flux trajectory which partially or entirely lies outside the flux hexagon is infeasible. For an increasing speed, the flux hexagon shrinks to the center of the flux plane since $\psi_R$ is inversely proportional to the speed. Any feasible flux trajectory must lie within the shrinking hexagonal limit.

A constant average flux magnitude is desired for a given torque command in the constant torque region (below the base speed) to assure MTPA operation. Fig. 17(a) and (b) shows an illustration of the different operating regions for a constant torque command and an increasing rotor speed. To operate in six-step in the flux weakening region and maintain MTPA operation below the base speed, the flux command must be determined as follows.

1) $PWM$ and $OVM I (\omega_1 < \omega_2)$: The upper boundary of OVM I is reached when the flux hexagon touches the MTPA flux circle $[\psi_{avg} = 0.85 \psi_{pm}$, see Fig. 17(d)]. Therefore, the MTPA flux trajectory is feasible for any speed below $\omega_2$, and the MTPA operation can be guaranteed without any adjustments. From the flux aspect, these two regions together can be regarded as a linear flux region (see Fig. 16).

2) $OVM II (\omega_2 < \omega_3 < \omega_4)$: The shrinking hexagon cuts the MTPA-circle. The circular trajectory is partially saturated by the hexagon, as shown in Fig. 17(c). Here, the average flux in the machine is nonlinearly changing with respect to the flux command (red-dashed circle), and a constant average flux is not possible in OVM II without further adjustments.

3) $Six-step (\omega_4 \geq \omega_4)$: In six-step, the flux hexagon is entirely located inside the circular flux (gray area) and the flux vector can only follow the shrinking hexagon [see Fig. 17(f) and (g)]. The machine must enter the flux weakening mode to guarantee an operation at the voltage limit. The maximum possible $R_\psi$ is obtained in the six-step mode ($R_\psi = 0.909$).

In conclusion, the feasible flux trajectory [shown in yellow in Fig. 17(c)–(g)] depends on the relative size of the shrinking hexagon and the circle and can be simply determined, for a given flux angle (e.g., $\delta_{avg}$), according to (9) and Fig. 18. Equation (9) is very simple to implement and is valid in all the operating regions (linear, OVM I, OVM II, and six-step).

$$|\psi_{avg}| = \min\{|OC|, |OH|\}. \quad (9)$$

B. ACHIEVING A LINEAR CHANGE IN THE MODULATION INDEX IN OVM II

The magnitude of the flux command is set to $|\psi_{avg}| = \psi_{MTPA}$ in the linear flux region. Here, the obtained flux is equal to the flux command, as shown in Fig. 19(a). For an increasing speed, the ratio $|\psi_{avg}|/\psi_R$ increases and reaches $0.866$ at the end of the linear flux region (lower boundary of OVM II). From this point onward, the flux command is partially saturated by the shrinking hexagon where the obtained flux is nonlinearly changing with the flux command [see Fig. 19(a)]. The six-step is reached when $|\psi_{avg}|/\psi_R = 1$.

To guarantee a linear change in the modulation index in OVM II, an adjustment of the radius of the circle must be made according to the LUT data in Fig. 19(b). The LUT data are the inverse of the nonlinear part in Fig. 19(a). With the help of the LUT, the nonlinearity between the flux command and the actual average flux can be compensated, as seen in Fig. 19(c). Accordingly, the widths of OVM I and OVM II regions become almost equal, and a constant average flux (unity gain) is guaranteed in the constant torque region with a simple adjustment of the flux command in OVM II.
FIGURE 17. (a) MTPA flux and $\psi_R$ for an increasing rotor speed and constant torque command. (b)–(f) Feasible flux trajectories (yellow) at different rotor speeds. The MTPA flux circle (red) has a constant radius while the flux hexagon (blue) is shrinking.

FIGURE 18. Calculation of the feasible flux command (applies to the linear, OVM I, OVM II, and six-step regions). $H$ is a point on the hexagon and $C$ is a point on the circle. (a) $\|OC\| > \|OH\|$. (b) $\|OC\| < \|OH\|$.

Given that $\psi_R$ in (5) is inversely proportional to the speed, and with the help of (9) and the LUT in Fig. 19(b), a smooth and continuous transition between the operating regions is possible (in the entire torque–speed range). It should be noted that for a changing torque command at a constant speed, the radius of the circle in Fig. 17 will change while the flux hexagon is fixed, leading to a smooth transition between the four regions of operation. It is worth mentioning that the average flux information, which is necessary to calculate the correct flux angle setpoint ($\delta_{dq}^* - \text{LUT}$) in Fig. 13, can be directly estimated from Fig. 19 for the six-step mode.

FIGURE 20. (f)–(u) depicts the flux trajectory at 16 different time instants between $t = 0.4$ ms and $t = 4.6$ ms. The commanded torque is infeasible and multiple steps are needed to reach the desired torque in steady state. A detailed description is given as follows.

1) $t = 0.4$ ms, Fig. 20(f): The inverter operates in the linear voltage region and the flux tracks the circular-MTPA flux trajectory.
2) $t = 0.5$ ms, Fig. 20(g): A step change in torque is applied to the torque controller, leading to an increase in the desired flux angle ($\delta_{dq}^*$). The flux vector command (yellow dot) is determined from the knowledge of $\psi_R$ and $\delta_{dq}^*$.
3) $t = 0.6$ ms, Fig. 20(h): The infeasible voltage command is saturated by the hexagonal voltage limit. The main constraint in achieving a deadbeat control of torque and flux is the limited dc-link voltage. Due to the voltage limit, several steps might be needed to move the flux vector from one point to another in the flux plane.

A dynamic OVM is usually employed to use more voltage and achieve a faster torque response. In six-step, the voltage magnitude is already at its maximum in steady state. Since the torque during transient operation cannot be decoupled from the flux magnitude in flux weakening ($MI = 1.103$), and due to the hexagonal flux nature, it is especially necessary to understand and analyze the dynamic performance at the maximum voltage limit.

A. UNOPTIMIZED FLUX TRAJECTORY DURING TRANSIENT OPERATION

The DBFC transient performance is analyzed in Fig. 20. A step command in torque ($T_{em}^* = 0 \rightarrow 0.51$ p.u.) is sent to the torque controller at $t = 0.5$ ms. The speed is 0.47 p.u. Fig. 20(f)–(u) depicts the flux trajectory at 16 different time instants between $t = 0.4$ ms and $t = 4.6$ ms. The commanded torque is infeasible and multiple steps are needed to reach the desired torque in steady state. A detailed description is given as follows.

1) $t = 0.4$ ms, Fig. 20(f): The inverter operates in the linear voltage region and the flux tracks the circular-MTPA flux trajectory.
2) $t = 0.5$ ms, Fig. 20(g): A step change in torque is applied to the torque controller, leading to an increase in the desired flux angle ($\delta_{dq}^*$). The flux vector command (yellow dot) is determined from the knowledge of $\psi_R$ and $\delta_{dq}^*$.
3) $t = 0.6$ ms, Fig. 20(h): The infeasible voltage command is saturated by the hexagonal voltage limit. The
FIGURE 19. (a) Obtained average flux magnitude compared to the commanded flux (normalized to $\psi_R$) in the different operating regions. (b) LUT data needed to achieve a constant average flux magnitude in OVM II. (c) Linear relationship between the commanded flux and the obtained average flux is achieved with the help of the LUT in (b).

FIGURE 20. Simulation results showing the (a) torque, (b) flux magnitude, (c), (d) flux and current trajectories in the synchronous reference frame ($f_{dq}$), (e) flux angle, and (f)–(u) flux trajectory in the stationary reference frame ($f_{\alpha\beta}$) during transient operation for a step command in torque ($T^*_{em} = 0 \rightarrow 0.51 \text{ p.u.}$). The speed is $0.47 \text{ p.u.}$. $T_s = 100 \mu\text{s}$.

Voltage vector is adjusted to the closest voltage-hexagon corner. The flux angle starts increasing leading to positive torque production [see Fig. 20(e)]. The $d$-axis rotates in the anticlockwise direction by $\omega_r T_s$, during every $T_s$ seconds, and the flux vector command should rotate more and more in the same direction to keep $\delta_{dq}$ constant.

4) $t = 0.7 \rightarrow 1.4 \text{ ms}$, Fig. 20(i)–(m): The flux angle $\delta_{dq}$ keeps on increasing yielding to more torque. However, due to the unmodeled and uncompensated rotation of the $d$-axis, the flux has a random trajectory in the transient state.

5) $t = 1.6 \rightarrow 2.4 \text{ ms}$, Fig. 20(n)–(q): Due to the random flux trajectory and the unoptimized voltage vector, the flux vector goes outside the flux hexagon (flux strengthening), then inside it (flux weakening), see Fig. 20(b). The deviation from the optimum hexagonal flux trajectory during this time period is undesirable and causes a residual flux error that needs multiple steps to be compensated. Moreover, the current magnitude will have oscillations as shown in Fig. 20(d). The flux magnitude inherently decreases during the transient mode to provide a shorter path to the flux vector to lead the $d$-axis ($\delta_{dq} > 0$) and increase the torque in the machine.
Transient flux trajectory (yellow) from an initial steady-state \( \text{(10)} \), it is found that the flux is different due to the remarkable change in the settling time. The transient flux trajectory is not optimal due to \([\text{Fig. 20(f)}–(u)]\).

The random flux trajectory in the previous section can cause unexpected under- and overshoots in torque, flux magnitude, and current. This behavior can be completely avoided if the rotation of the \( d \)-axis is considered in the initial flux setpoint at the moment when the torque command is updated. Hence, the flux path during transient operation is guided by the hexagonal volt-sec limit only. Forcing the flux to follow a straight-line trajectory between two steady states minimizes the settling time needed to reach a final steady state and achieves time-optimal torque control with high accuracy.

**B. TIME-OPTIMAL TORQUE CONTROL WITH DBFC DURING TRANSIENT OPERATION**

If the torque and flux commands are not feasible in one single step, the deadbeat control law converges to a finite-settling-step control law, which can be optimized to minimize the settling time (time-optimal control).

The random flux trajectory in the previous section can cause unexpected under- and overshoots in torque, flux magnitude, and current. This behavior can be completely avoided if the rotation of the \( d \)-axis is considered in the initial flux setpoint at the moment when the torque command is updated. Hence, the flux path during transient operation is guided by the hexagonal volt-sec limit only. Forcing the flux to follow a straight-line trajectory between two steady states minimizes the settling time needed to reach a final steady state and achieves time-optimal torque control with DBFC.

Fig. 21 depicts an optimal flux trajectory during transient operation. The transient flux trajectory is defined by an initial steady-state \( A \) [see Fig. 21(a)] and a final steady-state \( B \) [see Fig. 21(b)]. The magnitude of the transient flux trajectory \((|\overrightarrow{AB}|)\) is directly related to the number of steps needed to reach a final steady state (due to the limited dc-link voltage). Therefore, a transient flux trajectory having the shape of a straight line ensures the shortest path between the two steady states in the volt-sec plane. For a step increase in the torque command, the flux angle will increase from \( \Delta \theta_{d} \) to \( \delta_{d} \). However, the change in the angular position of the flux vector \((\Delta \theta_{AB})\) is higher than the quantity \( \delta_{d} \) due to the rotation of the \( d \)-axis. The angular change \( \Delta \theta_{AB} \) can be determined in (10) [see Fig. 21(c)].

\[
\Delta \theta_{AB} = \delta_{d} - \delta_{d} + \Delta \theta_{d} \tag{10}
\]

It can be seen from Fig. 21 that there exists a unique number of steps \((n_{opt})\) to achieve the desired torque command for time-optimal operation at the given operating conditions. For an arbitrary number of steps \( n \), the synchronous reference frame rotates by an angle \( \Delta \theta_{d} = n \omega_{T} T_{s} \). A time-optimal control can be achieved if, and only if, the point \( B \) (with \( \Delta \theta_{AB} = \delta_{d} - \delta_{d} + n_{opt} \omega_{T} T_{s} \)) is reached at the time instant \( T_{B} = T_{s} + n_{opt} T_{s} \).

To demonstrate this statement, the transient performance of DBFC is analyzed at ten different values of \( n (n = 1:1:10) \). Fig. 22 shows a detailed illustration of the transient operation for a step increase in torque command \((T_{em} = 0.20 \rightarrow 0.51 \text{ p.u.})\), the speed is 0.47 p.u. Only two conditions are shown: the unoptimized trajectory, and \( n = 7 \). In contrast to Fig. 20, where there is a transition from circular to hexagonal flux trajectory, the analysis here focuses on the voltage-limited operation (six-step) with a hexagonal trajectory in the initial and final steady states \((A \text{ and } B)\). The time scale in Fig. 22 is different due to the remarkable change in the settling time. A description of Fig. 22 is presented as follows.

1) **Fig. 22(1):** The transient flux trajectory is not optimized. The flux angle setpoint is determined from the torque controller in Fig. 13(c). The PI regulator is used to eliminate the steady-state error in torque, and it is deactivated during the transient time. The flux vector command \( \psi_{d}^{*} \) rotates by \( \omega_{T} T_{s} \) during every \( T_{s} \) seconds due to the rotating \( d \)-axis. The real flux vector tries to follow its command. However, achieving deadbeat torque control is not possible because of the voltage limitation. Thus, the magnitude of the volt-sec vector is saturated to its maximum \((\frac{2}{3}V_{dc}T_{s})\) in the transient state. To stay within the hexagonal volt-sec limit, the angle of the volt-sec vector is adjusted to the closest hexagon corner. This angle changes randomly due to the rotating flux setpoint \( \psi_{d}^{*} \) [similar to Fig. 20(f)–(u)]. The flux magnitude inherently decreases to increase the flux angle \( \delta_{d} \). The torque increases slowly with an unpredicted change. The settling time can be approximated to \( t_{setting} = 35 T_{s} = 3.5 \text{ ms} \). In this scenario, it takes \( 35 T_{s} \times f_{1} = 1.4 \) electrical cycles to reach the steady-state operation at point \( B \).

2) **Fig. 22(2):** For \( n = 7 \), it is found that the flux is capable of reaching its setpoint after seven time-steps. Therefore, the final steady state is reached with...
minimum settling time \( t_{\text{settling}} = n_{\text{opt}} T_s = 7 T_s = 0.7 \text{ ms} \) where the time-optimal control is successful. To achieve a straight-line trajectory, the flux angle setpoint is kept constant during the transient time. A 0.31 p.u. increase in the average torque is achieved in seven steps while operating in six-step (voltage limit). An undershoot in torque takes place due to the transient flux trajectory. The reduced flux magnitude leads to an undershoot in the \( d \)-axis current. An overshoot in the stator current magnitude \(|i_{dq}|\) is observed (17% of the steady-state current). The flux magnitude during transient is symmetrical around a vertical line since the transient flux path is a straight line. To produce a straight line, any voltage vector within the voltage hexagon can be used (PWM), and not only the six fundamental vectors. In fact, this path is not an ideal line due to the resistive voltage drop in the machine. The transient flux trajectory is a shortcut that helps the flux vector to move forward in the anticlockwise direction with an angular speed higher than \( \omega_r \) \((\omega_r - \omega_{\text{TOC}} > \omega_r)\).

By applying TOC, the settling time in this example can be decreased by 80%. It should be noted that the DBFC performance is stable in transient mode, even with an unoptimized transient trajectory. The undershoot in torque (and overshoot in current) in Fig. 22(b2) can be observed at some transient operating conditions. This is attributed to the straight-line trajectory where the angular difference between the flux vector and the \( d \)-axis decreases in the first time-steps and then increases rapidly, as depicted in Fig. 22(d2). This behavior is inherent to the time-optimal torque controllers in general, and not only for DBFC. If this behavior is not acceptable (based on the application), a simple solution can avoid this issue by limiting the rate of change of the torque command. In this case, the controller will react less aggressively, at the cost of slightly increasing the settling time.

It should be noted that the TOC tests shown in this research are for increasing step commands in torque. Decreasing the torque with very fast transient operation is inherently possible without the need for time-optimal torque control. For decreasing torque, the flux vector rotates in the opposite direction of the \( d \)-axis in the transient mode. Thus, the torque angle (angle between the flux vector and the \( d \)-axis) can be reduced very quickly. The developed time-optimal torque controller can still be used for this condition, but it does not offer considerable improvement to the settling time as in the increasing torque case.

**C. Determining the Optimal Number of Steps for TOC**

To complete the design of TOC for DBFC, it is also desirable to calculate the optimal number of time steps \( n_{\text{opt}} \) for any transient torque conditions. Thus, the TOC can be achieved naturally, and without having to manually set (or test) \( n_{\text{test}} \) as in the previous section. The optimal number of steps can be obtained by solving the system of equations in (11).

The variable \( n_1 \) is the ratio of the change in the rotor position during the transient time \((\Delta \theta_{\text{rot}})\) to \( \omega_r T_s \). On the other hand, \( n_2 \) is the ratio of \(|A\overline{B}|\) to the approximated average magnitude of the applied volt-sec vectors \((\frac{1}{2} V_{dc} T_s h)\), and it is the number of time steps needed to move the flux vector with a straight line from \( A \) to \( B \), considering the voltage limit. The variable \( h \) is a gain used to approximate the transient volt-sec vectors to a set of vectors with constant magnitude \((h < 1)\). The variable \( h \) is determined empirically and was found to be close to 0.95.

The condition of the TOC is that \( n_1 \approx n_2 \). For given operating conditions and initial steady-state flux (point \( A \)), fulfilling the condition in (11) results in a unique final steady-state flux (point \( B \)). An angle “\( \delta \)” is defined as an angular shift with respect to the \( d \)-axis, whereas an angle “\( \theta \)” is an angular shift...
The system of equations in (11) can be rearranged into one equation and solved for $\theta_B$. However, finding its analytical solution is cumbersome and complicates the implementation. The optimal number is, however, calculated numerically with iterations as shown in Fig. 23 (yellow).

There is a small chance that $n_1 = n_2$, due to the discrete-time nature of the control law, which operates at a constant sampling frequency. The length of the transient-straight-line trajectory in the volt-sec plane is generally not an integer multiple of the volt-sec command, where the following equation is not valid for most of the transient operating conditions:

$$|\Delta n| = \frac{|\Delta \theta|}{\Delta \theta} n_{\text{opt}}$$

An integer number of optimal time steps that minimizes the error $|\Delta n| = |n_1 - n_2|$ is selected. A variable $n_1$ is set to 1 in the first iteration, where the $d$-axis rotation for TOC is assumed to be $n_1 \omega_r T_s = 1 \omega_r T_s$. The angle $\theta_B$ is calculated in the next step, based on the $n_1$ value from (11) (see Fig. 21). The magnitude $|\overline{AB}|$, and eventually $n_2$ are obtained from $\theta_B$. The error between the two variables is stored in $\Delta n(i) = |n_1(i) - n_2|$. The value of $n_1$ is increased by “+1.” The same process is repeated until the maximum number of iterations is reached (ten iterations in this example). While activated, the TOC-algorithm does not block the execution of the DBFC controller. The optimal number of time steps ($n_{\text{opt}}$) is equal to the variable $n_1$ that minimizes the error vector $\Delta n$.

In the next step (gray), the angle, and magnitude of the vector $\overline{OB}$ are properly defined (polar coordinates of point B). The flux setpoint ($\psi^{*}_{d-B}$) is fixed during the transient time ($t_{\text{transient}} = n_{\text{opt}} T_s$). The DBFC controller uses this information to calculate the voltage command to achieve TOC. The voltage command is calculated and updated according to the flux setpoint, considering the resistive voltage drop, which slightly changes the direction of the voltage command. Since the torque command is continuously updated in automotive drive applications, the TOC algorithm is only activated if the change in the torque command (or the flux angle command) is higher than a predefined threshold as given by (13). In this research, the threshold is set to 10% of the rated torque. Consequently, the TOC algorithm is deactivated in steady-state operation or for small torque transients, which saves unnecessary computational loads.

$$|T_{\text{em}}^{*}(k) - T_{\text{em}}^{*}(k-1)| > \Delta T_{\text{em}}^{*}\text{--threshold}.$$  (13)

**VIII. PROPOSED DBFC CONTROL SCHEME**

Fig. 24 depicts the proposed DBFC control scheme for IPMSMs, which works in the four operating regions and within the voltage and current limits. The closed-loop torque controller (green) calculates the desired flux angle ($\delta_{d-B}^*$) from the torque and the flux information. The hexagonal flux limit (purple), defined by $\psi_R$, is calculated in (5). The MTPA flux (tan) is calculated from the torque command. $\psi_{\text{MTPA}}$ and $\psi_R$ are used to determine the feasible flux trajectory (orange). According to the operating point ($\psi_{\text{MTPA}}/\psi_R$), the desired flux trajectory can be circular, hexagonal, or a combination of a circle and a hexagon.

After determining the flux vector command, the voltage reference (red) is calculated from (2). A voltage selection block is needed to guarantee six-step operation and ensure correct tracking of the flux at the corners of the flux hexagon (more details can be found in [9]). As opposed to the voltage hexagon in Fig. 4 (red), any feasible voltage vector within the hexagonal voltage boundary can be produced in the current implementation, covering the entire voltage hexagon [see Fig. 2(d)]. The closed-loop discrete-time Gopinath flux linkage observer in [39] is used to estimate the state variable at the next sampling time instant. A current observer is not necessary for this implementation, therefore, saving some computational burdens.
The DBFC controller operates inherently within the voltage limit due to its feasible flux trajectory. The stator current is restricted by the inverter (or the machine) thermal limit. In DBFC, the torque and flux are closed-loop controlled parameters. Additional monitoring is required to ensure operation within the current limit. A PI current limiter (yellow) is applied to enforce feasible torque commands with respect to maximum current, thus providing thermal protection for the machine and the inverter. A PI regulator is used for this purpose.

IX. EXPERIMENTAL RESULTS

The control algorithm in Fig. 24 is tested on an industrial automotive testbench with an automotive microcontroller. The IPMSM machine under test is a high-power/high-performance automotive traction machine. The torque-controlled test machine is mechanically coupled with a speed-controlled load machine. The base current and base torque are equivalent to the rated current and rated torque, respectively. The base speed is equal to the maximum speed. The sampling frequency is 10 kHz. The switching frequency in the linear region is 5 kHz.
A. TEST 1: EXPERIMENTAL EVALUATION OF CONTINUOUS TRANSITION FROM THE LINEAR REGION TO SIX-STEP

In this test, the torque command is set to $T_m^* = 0.68$ p.u. The speed is increased from 0.29 to 0.59 p.u. to cover the transition region. The control algorithm is capable of smoothly transitioning between the different operating regions as illustrated in Fig. 25. A constant torque is produced in the linear (yellow), OVM I (green), and OVM II (blue) regions with a constant average flux magnitude (between ① and ⑦). The flux ripples increase in OVM II and reach their maximum at ⑦ where the flux weakening region is entered while operating in six-step (gray).

Due to the constant average flux below the base speed, the modulation index, calculated from the measured line voltages [see Fig. 25(e)], increases linearly and continuously from 0.88 to 1.103, confirming the seamless and smooth transition...
between the different operating regions. The widths of OVM I and OVM II are almost equal since the modulation index is linearly changing with the speed.

More current is needed to produce a constant torque in flux weakening. At a certain speed (around 0.42 p.u. \( t = 17.6 \) s), the current vector leaves the constant torque hyperbolas and moves along the maximum torque per flux (MTPF) line, leading to a reduced torque and current magnitude. A similar measurement covering the entire speed range can be found in [32].

The measured line voltages and phase currents at the eight different operating points (see \( 1^\circ \)–\( 8^\circ \)) are shown in Fig. 26. The three line voltages can be used to obtain the phase voltages needed to calculate the modulation index. The rms value of the voltage increases by moving from \( 1^\circ \) to \( 7^\circ \). The standard six-step voltage waveform is obtained in \( 7^\circ \) and \( 8^\circ \). The oscillations seen in Fig. 26 are mainly attributed to the dc-link capacitance value.

The current waveform depends on the voltage command and the operating point (e.g., torque, speed, etc.). In six-step, the current harmonics decrease at higher speeds (see \( 7^\circ \) and \( 8^\circ \)). This is mainly attributed to the frequency-dependent inductive reactance, which serves as a low-pass filter for the stator current waveform.

As shown in Fig. 26, the voltage increases gradually and reaches its maximum in six-step. The pulse number \( N_p = f_{\text{switching}} / f_{\text{electrical}} \) decreases during the transition and becomes unity in six-step, where the switching frequency is equal to the fundamental electrical frequency. Similar to the observation in Fig. 26, the current, and consequently the torque, ripples decrease in \( 8^\circ \) compared with \( 7^\circ \). The current ripples in Figs. 26 and 27 look slightly different due to the higher sample rate in the oscilloscope (2 MS/s). Due to the discrete-time nature of the control law (noninteger sampling-to-fundamental frequency ratios), tracking the
FIGURE 28. Experimental results showing the (a) measured phase current and (b) the corresponding harmonic spectrum, for different load conditions at a constant speed (0.82 p.u.).

flux hexagon’s corners is only possible with a single PWM insertion as discussed in Section VIII. If needed, two fundamental voltage vectors are used in one calculation period. The vector addition of their equivalent volt-sec quantities is shown as light-yellow crosses (in the voltage plane) in Fig. 27. Nevertheless, the inverter operates at the real maximum voltage limit as depicted in the voltage measurement.

B. TEST 2: EXPERIMENTAL EVALUATION OF THE CURRENT HARMONICS IN SIX-STEP

It was shown in Section IX-A that the current harmonics depend on the operating frequency. In fact, these harmonics also rely on load conditions. In this test, the operating speed is constant (0.82 p.u.) and the torque is increased from 0.042 to 0.255 p.u., as illustrated in Fig. 28. The phase currents are measured with a LeCroy 4 channel oscilloscope (sample rate is 2 MS/s). It can be concluded that the main current harmonics (fifth and seventh) decrease with the increasing torque, despite the same harmonic content in the phase voltage at all the operating points.

A high-bandwidth CVC is initially designed to control the fundamental current vector. If not properly compensated, the inherent low-order harmonics in the feedback current during six-step deteriorate the performance of the PI-regulator in steady state and transient operation and make six-step operation cumbersome. Due to the instantaneous flux vector control with DBFC, and the use of the Gopinath flux linkage observer, the estimation and compensation of the current harmonics is not necessary, which reduces the complexity of the implementation.

C. TEST 3: EXPERIMENTAL EVALUATION OF THE TORQUE CONTROLLER IN STEADY STATE

Fig. 29 evaluates the torque accuracy for an increasing torque command ($T_{em}^* = 0 \rightarrow 0.68$ p.u.) at a constant speed (0.41 p.u.). The hybrid torque controller in Fig. 13(c) is employed in this test.

The torque-LUT data are not precise in this example for the low-torque commands (between 1° and 6°). In the zoomed-in view of Fig. 29(a), at 1° and 2°, the LUT immediately updates its output according to the torque command. The PI regulator ($K_p = 0.001 \text{ rad/N \cdot m}$, $K_i = 0.01 \text{ rad/N \cdot m/s}$), with a low bandwidth, reacts to slowly reduce the steady-state error in torque. The controller bandwidth can be slightly increased to reduce the settling time. The LUT and PI regulator outputs ($\delta_{LUT}^*$ and $\delta_{PI}^*$) are given in Fig. 29(b).

The flux magnitude and the modulation index increase with the torque command in the linear and OVM regions (between 1° and 5°) while staying at MTPA. At 3°, the ratio $\psi_{MTPA}/\psi_R$ becomes high enough to enter OVM I. Here, the switching frequency starts decreasing and the modulation index is close to unity, as shown in Fig. 29(e) and (f), respectively. The OVM II region is entered at 4°. The flux and torque ripples increase due to the inherent nature of the flux trajectory. The six-step operation is automatically realized at 5° where the MTPA flux circle is entirely located outside the flux hexagon (see Fig. 17). The switching frequency becomes equal to the fundamental electrical frequency and the switching losses are minimized. The flux hexagon shrinks slightly between 5° and 7° since the resistive voltage drop is proportional to the increasing current.

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From (6) onward, the torque-LUT data are more precise, and accurate torque control can be guaranteed with minor interventions from the PI regulator. The torque is controlled with excellent accuracy and stable operation in the four regions. In this test, the stator current is within its maximum limit. A similar measurement covering the operation at the current limit and the MTPF limit can be found in [32].

D. TEST 4: EXPERIMENTAL EVALUATION OF THE TIME-OPTIMAL TORQUE CONTROL

1) CHANGING $n_{opt}$ MANUALLY

The TOC algorithm is tested experimentally by repeating the same test in Fig. 22. The sampling/calculation period is set to 125 $\mu$s to test the performance of the controller at a higher value of $T_s$. The calculation of $n_{opt}$ in Fig. 23, is switched OFF for this test. The optimal number of time steps is determined empirically, by changing the $n_{test}$ value from 1 to 10 and looking at the transient flux trajectory. A description of the results is presented as follows.

1) Fig. 30(1): The transient flux trajectory is not optimized ($n_{test} = 0$). The torque increases slowly with an unpredictable change. The settling time can be approximated to $t_{setting} = 30 T_s = 3.75$ ms (time difference between the initial steady-state A and the final steady-state B).

2) Fig. 30(2): $n_{test}$ is set to 5. The flux vector leaves point A at $t_1$ and tries to reach a final steady state within five steps, while achieving an average torque increase of 0.31 p.u. However, five time-steps are not enough to
reach the final target and achieve TOC. The settling time in this scenario decreases to $t_{\text{settling}} = 14 T_s = 1.75$ ms.

3) Fig. 30(3): For $n_{\text{test}} = 6$, the flux is capable of reaching its setpoint after six time-steps ($n_{\text{test}} = n_{\text{opt}}$). The final steady state is reached with minimum settling time ($t_{\text{settling}} = n_{\text{opt}} T_s = 6 T_s = 0.75$ ms) where the time-optimal control is successful. The main difference between the simulation (see Fig. 22) and the experiment (see Fig. 30) is the increased calculation period in the experimental test. The circumradius of the volt-sec hexagon is proportional to $T_s$ (and $V_{dc}$). Accordingly, the flux vector can be changed with a higher rate at higher $T_s$ values. Therefore, the TOC control is achieved here ($T_s = 125 \mu s$) with one step less than in the simulation test ($T_s = 100 \mu s$), but the settling time remains similar.

4) Fig. 30(4): The optimal number of steps is assumed to be seven ($n_{\text{test}} = 7$). This number is higher than the real $n_{\text{opt}}$, leading to higher torque production in transient. The average torque decreases to reach 0.51 p.u., but more steps are needed to correct the undesirable error in the initial flux-setpoint. The settling time is $t_{\text{settling}} = 11 T_s = 1.375$ ms.

In conclusion, the experimental results agree with the simulation results. By applying TOC, the settling time in this test can be decreased by approximately 80%. DBFC performance is stable in transient-torque operation, even for the unoptimized trajectory or for $n_{\text{test}} \neq n_{\text{opt}}$.

2) AUTOMATIC CALCULATION OF $n_{\text{opt}}$

In Fig. 31, $n_{\text{opt}}$ is automatically calculated by the TOC algorithm. The transient response is analyzed at three different operating conditions as follows.

1) Fig. 31(1) and (2): $T_{\text{em}}^* = 0 \rightarrow 0.20$ p.u. To test the dynamic transition from the circular-MTPA flux to the hexagonal flux in six-step, a step change in torque is sent to the torque controller.

For the unoptimized transient trajectory, several steps are needed to leave the MTPA circle and reach a final steady state on the flux hexagon. At the operating conditions ($\Delta T_{\text{em}}^*$, speed, $V_{dc}$, $T_s$, etc.), $n_{\text{opt}}$ is between 5 and 6. In this case, the floor of $n_{\text{opt}}$ is taken ($n_{\text{opt}} = 5$) in order to stay within the flux hexagon ($A \rightarrow D$). An additional corrective step is needed, aiming at the next flux setpoint ($C_2$, $\theta_{C_2} = \theta_{C_1} + \omega_r T_s$). In other words, the infeasible straight-line flux trajectory can be divided into two segments ($AD$ and $DE$) to compensate for the discrete-time resolution where (12) is not valid. If point $C_2$ is not reached due to the voltage limit, the flux trajectory follows another straight line ($EB$), which is parallel to the closest flux hexagon side. However, most of the desired torque is already produced at $E$, in six time-steps. The implementation of this specific methodology is very simple and costs negligible computational load. It should be noted that aiming at $C_1$ instead of $C_2$ in the 6th step is possible, but not favorable since $C_1$
would slightly decrease the flux angle and the produced torque.

2) Fig. 31(3) and (4): $T_{em}^* = 0 \rightarrow 0.43$ p.u., and the speed is 0.71 p.u. At low sampling-to-fundamental frequency ratios (e.g., high speeds), the area of the volt-sec hexagon becomes closer to that of the shrinking flux hexagon. Thus, any error arisen from the noninteger ratio in (12) must be considered. The machine operates in flux weakening (six-step), even for null torque. In total, ten time-steps ($10 T_s$) are needed to reach a final steady-state ($B$) for the unoptimized trajectory. Here, $n_{opt}$ is not an integer. The time-optimal control can be achieved in $4T_s$ with one additional corrective step.

3) Fig. 31(5) and (6): $T_{em}^* = 0 \rightarrow T_{MTPF} = 0.27$ p.u., the speed is 1 p.u. The flux hexagon’s area is approximately 2.5 times bigger than that of the volt-sec hexagon. The instantaneous value of the torque reaches the reference value in just four steps for the unoptimized trajectory. This short period of time is mainly attributed to the size of the flux hexagon. Here, the flux hexagon is small enough that the flux vector can change its position on the hexagon very quickly. However, six more time steps are needed to compensate for the residual error in the flux. Moreover, the flux vector goes outside the flux hexagon. With TOC, the desired torque is produced in almost three steps, one additional corrective step is needed. The settling time decreases by 60% in this test.
X. CONCLUSION

The results of this research can be summarized as follows:

1) A DBFC is developed as a single control law that can operate in the entire torque–speed range (linear region, OVM I, OVM II, and six-step) by accurately tracking the flux trajectories in the different operating regions. By gradually changing the flux trajectory from circular to hexagonal, a smooth and continuous transition between the four regions is ensured where the inverter can operate at any desired modulation index between $MI = 0$ and $MI_{\text{six-step}} = 1.103$. Undesirable torque dynamics, stability problems, and high computational burdens associated with the use of multiple control laws are avoided.

2) The transient performance of DBFC at the voltage limit is analyzed in detail. If the torque and flux commands are not feasible in one single step, the deadbeat control law converges to a finite-setting-step control law, which can be optimized to minimize the settling time. A time-optimal torque control algorithm is developed to achieve the fastest possible torque dynamics with a simple implementation. It calculates the number of optimal time steps needed to achieve to desired torque with a straight-line flux trajectory in the stationary reference frame during transient operation. In steady state, the torque can be controlled with high accuracy and high robustness to parameter variations.

3) According to the abovementioned points, DBFC offers two significant advantages, simultaneously: operating at any feasible modulation index (also maximum MI) with smooth transition and ensuring excellent dynamic performance. The proposed DBFC control law, with TOC, is simple to implement and requires limited computational burden. It is tested on an automotive microcontroller with an automotive traction machine.

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