Enhanced DBCC for high-speed permanent magnet synchronous motor drives

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Abstract: High bandwidth and accuracy of the current control loop are fundamental requisites when a fast torque response is required or for facilitating the reduction of torque ripple in high-performance drives, especially at high speed. One of the most suitable control methods to achieve these goals is dead beat current control (DBCC). Many types of DBCCs have been proposed and implemented in the literature. This study proposes a DBCC incorporating two new functionalities. One is a two steps current prediction to improve prediction accuracy when current measurements are taken place before each sampling period; and particularly to reduce the overshoot during transients when mean value is used as current feedback. The second is a novel compensation method for the rotor movement to eliminate offset errors which occur at high speed. Moreover, the dynamic and steady-state performance of the proposed DBCC is assessed in simulations. On the basis of the simulation tests, the control parameters are tuned for experiments and the performance of the proposed functionalities are verified. Finally, the advantage of DBCC, compared with a classical dq proportional–integral current regulator, is verified in experiments.

1 Introduction

Dead beat current control (DBCC) is categorised as belonging to the predictive control family. It is one of the possible solutions to achieve high-bandwidth and high-accuracy current control loop has been successfully applied for many industrial fields where high performance is required. For examples, for grid generation system in [1]; for multilevel converters in [2]; and for permanent magnet synchronous motors (PMSMs) drives [3–8]. It has been introduced for the first time for the control of a pulse-width modulation (PWM) inverter used in an uninterruptible power supply (UPS) [9].

PMSMs are widely used in industry applications and different works have been proposed recently in literature [10–14]. In PMSMs, high-frequency electromagnetic torque ripple appears due to the distorted stator flux linkage, variable magnetic reluctance at the stator slots, and imperfect mechanical alignment. Therefore, high bandwidth and high accuracy in the current control loop (such as DBCC loop) are the fundamental requisites for facilitating the reduction of torque ripple or when a fast torque response is required in high-performance PMSM drives. For example, the basic structure of DBCC for PMSM drives, which is embedded for the compensation of torque harmonics, is proposed and validated with simulation and experimental results in [3, 4]. Wipasuraromont et al. in [7] propose a DBCC scheme to achieve fast dynamic response.

Alongside the variety of DBCC schemes proposed for power electronics and drives, comparative studies against classical methods and the different types of DBCC have also been presented in [15–17]. It is recognised in scientific literature that a current control loop with DBCC has potential to have higher bandwidth compared with a current loop using traditional proportional–integral (PI) current regulators. However, some of these studies claim that, given its model-based characteristic, deterioration of DBCC performance and eventual instability are possible due to mismatch in model parameter values, un-modelled delays, dead-time effects, and other errors in the model.

Many solutions have been proposed in the literature in order to solve this problem. Zhou and Liu [18] propose a fast PI controller based on dead beat algorithm for active power filters. Disturbance observers have been applied to the DBCC for an UPS application to reduce control sensitivity for model uncertainties, parameter mismatch, and noise on sensed variables [19]. For a PMSM drive application, the DBCC combined with classical PI current regulators has been proposed to reduce the current errors that arise due to model mismatch and the non-ideal behaviour of the inverter during steady-state operation in [7]. For a three-phase PWM voltage-source inverter, the DBCC with an adaptive self-tuning load model has been proposed to reduce model mismatch in [20]. A current observer with an adaptive internal model is instead proposed in [21] to compensate system uncertainties of DBCC. A novel neural network-based estimation unit has been proposed to estimate, in real time, the grid impedance and voltage vector simultaneously in [22].

Though the problem of DBCC has been claimed and many solutions have been proposed, quantitative assessments for its dynamic and steady-state performance on the pre-mentioned uncertainties are not sufficient in existing literature.

DBCC can ideally force the control error to zero in one sampling interval after a correct voltage has been applied to the motor. Since this ‘correct voltage’ needs to be calculated before being applied, the traditional DBCC implemented in [8] can achieve the current reference two sampling periods (one period for calculation and waiting to be applied at the beginning of the next period) after a new current reference has been applied in the controller. In such implementation, the current measurements are designed to be taken at the beginning of each sampling interval. It may be worth redesigning the traditional DBCC assuming measurements are taken place before the beginning of each sampling interval considering the following three reasons: (i) feedback signals cannot be acquired instantaneously, but sampling and conversion times are necessary before a new measurement is made available. To be sure that new measurements are available for the controller
before the beginning of the next sampling period, the acquisitions need to be started in advance. In some cases, the amount of time the acquisitions are started in advance is much greater than the time strictly necessary because it could be more convenient to synchronise the acquisition to particular instants within the PWM pattern to minimise the acquired noise. (ii) As the speed of PMSM increases, the current measured at the middle of the sampling interval can be more and more different from the value measured at the beginning due to rotor movement. Sampling in advance can possibly make the measurements more close to the mean current; therefore, the controller can work to bring the mean current to the demand. (iii) Particularly, in case the current is oversampled for reducing noise, and mean current value is calculated as feedback, overshoots in current response may occur using the traditional DBCC, which can be reduced by properly setting the advanced sampling time.

What is more, the rotor movement is also responsible for an error between the voltage demand and real voltage seen by the motor, consequently steady-state error in current response increases as speed increases. Therefore, the rotor movement effect needs to be compensated.

Hence, this paper first proposes a DBCC with two new functionalities: one is a two steps current prediction (Section 2.1) to improve the accuracy of the current prediction when measurements are taken before the beginning of a period; and particularly in case of the mean current over a period is used as the feedback as in this paper, to cancel the false current error during transients. The other one is a novel compensation method for the motor movement prediction (Section 2.3) to eliminate offset errors which occur at high speed. Second, this paper reveals the availability of samples at the beginning of the 4th interval, $T_{cs}$, should be chosen long enough to cover analogue/digital conversion and transmission times. Full calculations for the DBCC are executed in each sampling period $T_s$. Therefore, the calculation time $T_{calc}$ needs to be smaller than $T_s$. Essentially the required calculations are performed in three steps as follows.

### 2 Proposed dead beat current control

This section demonstrates the proposed DBCC with two steps current prediction and rotor movement compensation.

The voltage equations of PMSM in a dq reference frame synchronous with the rotor are as follows:

\[
v_d = R_d i_d + L_d \frac{di_d}{dt} - \omega_L L_d i_q \\
v_q = R_q i_q + L_d \frac{di_q}{dt} + \omega_L L_d i_d + \omega_L \psi_m
\]

where $v_{dq}$ are the stator dq-axis voltages, $i_{dq}$ are the stator dq-axis currents, $\omega_L$ is the rotor electrical angular speed. Stator inductances $L_{dq}$, stator resistance $R_s$, and permanent magnet flux linkage $\psi_m$ are assumed to be independent from stator currents $i_{dq}$ and rotor angle $\theta$.

The inverter output is updated according to the $dq$ stator voltage references $v_{dq}^{ref}$ only at the beginning of every sampling period $T_s$.

### 2.1 Current prediction

The calculated $(k+1)$th $dq$ voltage references will be effectively applied at $t_{k+1}$, so the initial $dq$ currents used for calculation should be those at $t_{k+1}$, and not the ones measured at $(t_{k+1}-T_{cs})$. This to prevent any overshoot or inaccuracy in the response during transients, since the $dq$ currents at $t_{k+1}$ can be different from those measured at $t = t_{k+1} - T_{cs}$ as a consequence of the reference voltages applied during the $(k-1)$th and $k$th sampling periods. Since the $(k+1)$th $dq$ currents are in the future with respect to the $k$th sampling period (during which calculations are being performed), they must be predicted using the motor model. Considering that the $(k-1)$th and $k$th dq voltage references are applied in the time interval $[t_{k-1}, t_{k}]$, the predictions must be performed in two steps. The first current prediction is to predict the $dq$ currents at $t_{k}$ using the $(k-1)$th $dq$ reference voltages and the measured $dq$ currents at $t_{k-1}-T_{cs}$. The second current prediction is to predict the $dq$ currents at $t_{k+1}$ using the 4th $dq$ voltage references and the previously predicted $dq$ currents at $t_{k}$.

For the first current prediction, the stator voltage references $v_{dq}^{ref}(k-1)$ are used instead of the real stator voltages $v_{dq}$, since voltage sensors are not usually present in a drive for cost reasons. Please note that $v_{dq}^{ref}(k-1)$ are available during the 4th sampling period since calculated during the $(k-2)$th sampling period. The $v_{dq}^{ref}(k-1)$ and the rotor electrical angular speed $\omega_L(k)$ are assumed constant in one sampling period. Here, $\omega_L(k)$ can be also assumed constant even in several sampling periods, since the mechanical system time constant is much larger than $T_s$. Furthermore, under the previous hypotheses, (1) and (2) can be rewritten as

\[
v_d^{ref}(k-1) = R_d i_d + L_d \frac{di_d}{dt} - \omega_L(k) L_d i_q \\
v_q^{ref}(k-1) = R_q i_q + L_d \frac{di_q}{dt} + \omega_L(k) L_d i_d + \omega_L(k) \psi_m
\]

Both sides of (3) and (4) are integrated from $t_{k-1}-T_{cs}$ to $t_{k}$.

\[
\int_{t_{k-1}}^{t_{k}} v_d^{ref}(k-1) dt = \int_{t_{k-1}}^{t_{k}} \left[ R_d i_d + L_d \frac{di_d}{dt} - \omega_L(k) L_d i_q \right] dt
\]

\[
\int_{t_{k-1}}^{t_{k}} v_q^{ref}(k-1) dt = \int_{t_{k-1}}^{t_{k}} \left[ R_q i_q + L_d \frac{di_q}{dt} + \omega_L(k) L_d i_d + \omega_L(k) \psi_m \right] dt
\]

The following equations are obtained from (5) and (6) assuming that current profiles are linear. This hypothesis is necessary for avoiding a closed-form solution of the differential equations (5) and (6). Though the hypothesis will not be true once the motor starts rotating and will even cause steady-state errors in current responses, this steady-state error can be cancelled by the rotor movement compensation method proposed in Section 2.3. The hypotheses may also be wrong if $T_{calc}$ is not sufficiently small (ten times) compared with the electrical time constant of the motor. In the case study of this paper, $T_{i}$ is 52 times smaller than

![Fig. 1: Timing sequence of DBCC](image-url)
the electrical time constant

\[ T_{cs}v^\text{ref}_d(k - 1) = \left( \frac{R_{cs}L_d}{2} + L_d \right) i_d(t_k) + \left( \frac{R_{cs}L_d}{2} - L_d \right) i_d(t_k - T_{cs}) \]

\[ - \frac{\omega_e(k)L_dT_{cs}}{2} \left[ i_d(t_k) + i_d(t_k - T_{cs}) \right] \] (7)

\[ T_{cs}v^\text{ref}_q(k - 1) = \left( \frac{R_{cs}L_d}{2} + L_d \right) i_q(t_k) + \left( \frac{R_{cs}L_d}{2} - L_d \right) i_q(t_k - T_{cs}) \]

\[ + \frac{\omega_e(k)L_dT_{cs}}{2} \left[ i_q(t_k) + i_q(t_k - T_{cs}) \right] + \omega_e(k) \psi_m T_{cs} \] (8)

The 4th dq currents are detected at \( t_k - T_{cs} \). Therefore, the following assumptions for the measured currents are used to execute the first current prediction

\[ i_d^\text{mea}(k) = i_d(t_k - T_{cs}) \] (9)

\[ i_q^\text{mea}(k) = i_q(t_k - T_{cs}) \] (10)

Therefore, the estimated dq currents \( ^{\text{est}}v_d(k) \) of the first current prediction are derived from (7) to (10) (see (11 and 12))

\[ a_1 = \frac{L_dL_q^2}{2}, \quad a_2 = \frac{R_{cs}L_d}{2} + L_d, \quad a_3 = \frac{R_{cs}L_d}{2} + L_q, \quad a_4 = \frac{R_{cs}L_d}{2} - L_q \]

In the experimental tests of this paper, the currents are sampled at a rate of 15 MHz on the field programmable gate array (FPGA), and the mean current value over one sampling period is fed back to the controller so that it can bring the mean currents to its reference in steady state. As a consequence, to avoid the false current error during transients, it is necessary to use the first prediction to predict the real current at the end of each sampling interval from the mean value. In such case, the mean current can be assumed to be the same as the current measured at the middle of the sampling interval \( i_d(t_k - T_{cs}/2) \) since a linear profile of current is assumed. Hence, only for taking into account the particular way the current are measured in the experimental system, \( T_{cs} = T_s/2 \) in this paper.

The second current prediction predicts the dq currents at \( t_{k+1} \) using the 4th voltage reference and the dq current at \( t_k \) obtained from the first current prediction. For the second current prediction, the stator voltage references \( ^{\text{ref}}v_d(k) \) are used instead of the real stator voltages \( v_d(k) \). Again, the \( ^{\text{est}}v_d(k) \) are available during the 4th sampling period since calculated during the \((k-1)^{\text{th}}\) sampling period and the \( ^{\text{ref}}v_d(k) \) and the rotor electrical angular speed \( \omega_e(k) \) are assumed constant during the sampling period

\[ v^\text{ref}_d(k) = R_i j_d + L_d \frac{d i_d}{dt} - \omega_e(k)L_i j_q \] (13)

\[ v^\text{ref}_q(k) = R_i j_q + L_q \frac{d i_q}{dt} + \omega_e(k)L_i j_d + \omega_e(k)\psi_m \] (14)

Both sides of (13) and (14) are integrated from \( t_k \) to \( t_{k+1} \)

\[ \int_{t_k}^{t_{k+1}} v^\text{ref}_d(k)dt = \int_{t_k}^{t_{k+1}} \left\{ R_i j_d + L_d \frac{d i_d}{dt} - \omega_e(k)L_i j_q \right\}dt \] (15)

\[ \int_{t_k}^{t_{k+1}} v^\text{ref}_q(k)dt = \int_{t_k}^{t_{k+1}} \left\{ R_i j_q + L_q \frac{d i_q}{dt} + \omega_e(k)L_i j_d + \omega_e(k)\psi_m \right\}dt \] (16)

The following equations are obtained from (15) and (16)

\[ T_{cs}v^\text{ref}_d(k) = \left( \frac{R_{cs}L_d}{2} + L_d \right) i_d(t_{k+1}) + \left( \frac{R_{cs}L_d}{2} - L_d \right) i_d(t_k) \]

\[ - \frac{\omega_e(k)L_dT_{cs}}{2} \left[ i_d(t_{k+1}) + i_d(t_k) \right] \] (17)

\[ T_{cs}v^\text{ref}_q(k) = \left( \frac{R_{cs}L_d}{2} + L_d \right) i_q(t_{k+1}) + \left( \frac{R_{cs}L_d}{2} - L_d \right) i_q(t_k) \]

\[ + \frac{\omega_e(k)L_dT_{cs}}{2} \left[ i_q(t_{k+1}) + i_q(t_k) \right] + \omega_e(k)\psi_m T_{cs} \] (18)

Once again the following relations are considered

\[ i_d(t_k) = \hat{v}^\text{est}_d(k) \] (19)

\[ i_q(t_k) = \hat{v}^\text{est}_q(k) \] (20)

Therefore, the estimated dq currents of the second current prediction are derived from (17) to (20) (see (21)) and (equation (22) at the bottom of the next page)

\[ b_1 = \frac{L_dL_q^2}{2}, \quad b_2 = \frac{R_{cs}L_d}{2} + L_d, \quad b_3 = \frac{R_{cs}L_d}{2} + L_q, \quad b_4 = \frac{R_{cs}L_d}{2} - L_q \]

\[ \hat{v}^\text{est}_d(k) = \left[ a_1T_{cs}v^\text{ref}_d(k - 1) + \frac{(L_dT_{cs}^2/2)\omega_e(k)v^\text{ref}_d(k - 1)}{a_2\omega_e(k)^2 + a_3} \right] \frac{1}{a_1\omega_e(k)^2 + a_2a_3} \]

\[ - \frac{\omega_e(k)a_1 + a_2a_3}{a_1\omega_e(k)^2 + a_2a_3} \] (11)

\[ \hat{v}^\text{est}_q(k) = \left[ (L_dT_{cs}^2/2)\omega_e(k)v^\text{ref}_q(k - 1) - a_2T_{cs}v^\text{ref}_q(k - 1) \right] \]

\[ + \frac{L_d^2T_{cs}\omega_e(k)v^\text{est}_e(k) + \omega_e(k)^2a_1 + a_2a_3}{a_1\omega_e(k)^2 + a_2a_3} \] (12)

\[ \hat{v}^\text{est}_d(k + 1) = \] (21)

\[ \begin{bmatrix} b_1T_{cs}v^\text{ref}_d(k) + \frac{L_dT_{cs}^2}{2} - \omega_e(k)v^\text{ref}_d(k) \\ b_1\omega_e(k)^2 + b_2b_3 \end{bmatrix} \]

\[ - \frac{b_1\omega_e(k)^2 + b_2b_3}{b_1\omega_e(k)^2 + b_2b_3} \]

\[ \left[ \frac{L_dT_{cs}^2/2}2 \omega_e(k)v^\text{ref}_q(k) \right] \]

\[ \frac{b_1\omega_e(k)^2 + b_2b_3}{b_1\omega_e(k)^2 + b_2b_3} \]
2.2 Voltage references with current predictions

The \((k+1)\)th \(dq\) stator voltage references \(v_{dq}^{ref}(k+1)\), applied in the period \(t_{k+1}\) to \(t_{k+2}\), are calculated from the predicted \(dq\) currents at \(t_{k+1}\), \(i_{dq}^{pre}(k+1)\), and \(dq\) current references at \(t_k\), \(i_{dq}^{com}(k)\), as in the following. Again, the current profiles are assumed to be linear in order to simplify the differential terms

\[
v_{dq}^{ref}(k+1) = R_c i_{dq}^{ref}(k) + \frac{L_d}{T_s} \left[ i_{dq}^{ref}(k) - i_{dq}^{pre}(k+1) \right]
- L_q \alpha(k) i_{dq}^{ref}(k) \tag{23}
\]

\[
v_{dq}^{ref}(k+1) = R_c i_{dq}^{ref}(k) + \frac{L_d}{T_s} \left[ i_{dq}^{ref}(k) - i_{dq}^{pre}(k+1) \right]
+ L_q \alpha(k) i_{dq}^{ref}(k) + \psi_m \omega(k) \tag{24}
\]

In this case, the \(k\)th \(dq\) current references will effectively set the \(dq\) currents at \(t_{k+2}\), and consequently the digital control introduces a delay equal to two sampling periods. The only way to eliminate this delay is to predict the \((k+2)\)th \(dq\) current references starting from the \(k\)th ones which are known only when the calculations are performed. However, such a prediction is typically done by interpolation and is effective only in case of periodic references. When the \((k+2)\)th \(dq\) current references, \(i_{dq}^{ref}(k+2)\), are available, the \(v_{dq}^{ref}(k+1)\) can be calculated from the \(i_{dq}^{ref}(k+1)\) and the \(i_{dq}^{pre}(k+2)\) as described by the following equations

\[
v_{dq}^{ref}(k+1) = R_c i_{dq}^{ref}(k+2) + \frac{L_d}{T_s} \left[ i_{dq}^{ref}(k+2) - i_{dq}^{pre}(k+1) \right]
- L_q \alpha(k) i_{dq}^{ref}(k+2) \tag{25}
\]

\[
v_{dq}^{ref}(k+1) = R_c i_{dq}^{ref}(k+2) + \frac{L_d}{T_s} \left[ i_{dq}^{ref}(k+2) - i_{dq}^{pre}(k+1) \right]
+ L_q \alpha(k) i_{dq}^{ref}(k+2) + \psi_m \omega(k) \tag{26}
\]

For emphasising the intrinsic delay of the DBCC, no prediction for the reference currents is performed in this paper.

2.3 Rotor movement compensation

The \(v_{dq}^{ref}(k+1)\) are maintained for the whole \((k+1)\)th sampling period; however, during that period the rotor moves and the real \(dq\) voltages \(v_{dq}^{real}(t)\) applied to it are different from the \(v_{dq}^{ref}(k+1)\). Especially at high speed, the rotor movement in a sampling period is not negligible. Consequently, the non-constant real voltage \(v_{dq}^{real}(t)\) is responsible for the non-linear current profile, and the difference between the real average voltages \(v_{dq}^{avg}(k+1)\) applied to the rotor during the \((k+1)\)th sampling period and the reference voltages \(v_{dq}^{ref}(k+1)\) is responsible for a steady-state error/offset between the reference currents \(i_{dq}^{ref}(k+2)\) and the actual currents at \(t_{k+2}\). To avoid this issue, the novel technique proposed in this paper is to apply the compensated \(dq\) voltage references \(v_{dq}^{com}(k+1)\) at \(t_{k+1}\) so that the average \(dq\) voltages \(v_{dq}^{avg}(k+1)\) effectively applied to the motor during the \((k+1)\)th sampling period are equal to the \(v_{dq}^{com}(k+1)\). The relationships between the instantaneous \(v_{dq}^{real}(t)\) and the \(v_{dq}^{com}(k+1)\) are described by the following equations with reference to the \(dq\) reference frame taking into account the rotor movement in Fig. 2

\[
v_{dq}^{real}(t) = \cos(\theta(t) - \theta(t_{k+1})) v_{dq}^{com}(k+1) + \sin(\theta(t) - \theta(t_{k+1})) v_{dq}^{com}(k+1) \tag{27}
\]

\[
v_{dq}^{real}(t) = - \sin(\theta(t) - \theta(t_{k+1})) v_{dq}^{com}(k+1) + \cos(\theta(t) - \theta(t_{k+1})) v_{dq}^{com}(k+1) \tag{28}
\]

The term \((\theta(t) - \theta(t_{k+1}))\) represents the difference between the instantaneous position and the initial position at the beginning of each sampling period. Therefore, \((\theta(t) - \theta(t_{k+1}))\) can be replaced by \((\omega(k) t)\). Both sides of (27) and (28) are integrated from 0 to \(T_s\) to obtain

\[
T_s v_{dq}^{avg}(k+1) = \int_0^{T_s} v_{dq}^{real}(t) \, dt
= \int_0^{T_s} \cos(\omega(k) t) v_{dq}^{com}(k+1) \, dt + \int_0^{T_s} \sin(\omega(k) t) v_{dq}^{com}(k+1) \, dt \tag{29}
\]

\[
T_s v_{dq}^{avg}(k+1) = \int_0^{T_s} v_{dq}^{real}(t) \, dt
= - \int_0^{T_s} \sin(\omega(k) t) v_{dq}^{com}(k+1) \, dt + \int_0^{T_s} \cos(\omega(k) t) v_{dq}^{com}(k+1) \, dt \tag{30}
\]

The average voltages \(v_{dq}^{avg}(k+1)\), compensated voltages \(v_{dq}^{com}(k+1)\), and the rotor angular speed \(\omega(k)\) are assumed constant for the integration interval. By imposing that the applied \(v_{dq}^{com}(k+1)\) equals to \(v_{dq}^{real}(k+1)\), the following equations are obtained from

\[
\frac{\psi_m}{L_q} \omega(k) v_{dq}^{com}(k) - b_2 T_s v_{dq}^{com}(k) - b_1 \frac{T_s^2}{2} \omega(k) v_{dq}^{com}(k) - b_1 T_s \omega(k) v_{dq}^{com}(k)
\]

\[
\frac{\psi_m}{L_q} \omega(k) v_{dq}^{com}(k) - b_2 T_s v_{dq}^{com}(k) - b_1 \frac{T_s^2}{2} \omega(k) v_{dq}^{com}(k) - b_1 T_s \omega(k) v_{dq}^{com}(k)
\]

\[
\frac{\psi_m}{L_q} \omega(k) v_{dq}^{com}(k) - b_2 T_s v_{dq}^{com}(k) - b_1 \frac{T_s^2}{2} \omega(k) v_{dq}^{com}(k) - b_1 T_s \omega(k) v_{dq}^{com}(k)
\]

\[
\frac{\psi_m}{L_q} \omega(k) v_{dq}^{com}(k) - b_2 T_s v_{dq}^{com}(k) - b_1 \frac{T_s^2}{2} \omega(k) v_{dq}^{com}(k) - b_1 T_s \omega(k) v_{dq}^{com}(k)
\]

\[
\frac{\psi_m}{L_q} \omega(k) v_{dq}^{com}(k) - b_2 T_s v_{dq}^{com}(k) - b_1 \frac{T_s^2}{2} \omega(k) v_{dq}^{com}(k) - b_1 T_s \omega(k) v_{dq}^{com}(k)
\]
Therefore, the $v^\text{com} q(k + 1)$ are derived from (31) and (32) (see (33 and 34)).

A scheme of the implemented control can be seen in Fig. 3.

3 Proposed control strategy assessment with simulation and experimental tests

This performance evaluation is carried out with respect to three main issues: (i) the influence of detuned parameters and inverter dead time on the performance of DBCC, (ii) the effectiveness of the proposed DBCC with two steps ahead predictions and rotor movement compensation, and (iii) the advantage of DBCC over classical PI regulators. This section is therefore divided into three parts, each one presenting the results relative to the previous three issues.

Simulation tests are carried out using MATLAB Simulink. The simulation model includes the PMSM, the controller, and the three-phase inverter which is modeled considering a dead-time-related voltage error of which the polarity varies according to the polarity of the phase currents. The experimental test rig is set up as in Fig. 3 where a DBCC and a PI controller are both implemented for comparison. The current control schemes can be switched over in the software easily. The speed control loop is used to keep the rotor speed constant or generate the high-frequency sinusoidal signal in the $i^\text{ref}$ for test reasons. A ‘Triphase’ evaluation system, composed by an inverter and a real-time control platform, is used in the experimental tests. The program in the FPGA inside the Triphase Real-time Target can be compiled directly using MATLAB Simulink, and the Triphase inverter is controlled by the Real-time Target to drive a Control Technique PMSM (115UMC300) with parameters as shown in Table 1. A Siemens induction motor (1LA9113) is used as load.

3.1 Performance assessment

Regarding the desired performance for reducing torque ripple or achieving fast torque dynamic in PMSMs, it would be ideal if the DBCC could track a wide-frequency range of sinusoidal references with no attenuation, smallest possible phase shift and no offset in the average value. In this section, the Bode diagram is mapped in simulation to verify bandwidth, amplitude, and phase response of DBCC, whereas the steady-state error in step response is mapped to verify the offset. In summary, the following results in Sections 3.1.1 and 3.1.2 show that the dynamic performance (amplitude and delay) is affected mainly by detuned inductances, whereas the steady-state performance by the detuned magnetic flux.

3.1.1 Bode diagram: For mapping Bode diagrams as shown in Fig. 4, the reference current $i^\text{ref}$ is set to be a sinusoidal signal with bias of 8.34 A (the rated current), the motor speed $N_s$ is set to be a constant, and the amplification or attenuation in the magnitude of the response as well as the phase shift are calculated. This procedure has been repeated iteratively for current references of different frequencies and with different motor speed, different motor parameter detuned. For example, $L^\text{est}/L_d = 0.5$ in Fig. 4b means that $L^\text{est}$ (estimated in the control) is 50% of $L_d$ (real value in the motor). It is also to be noted that the results of DBCC in Fig. 4e are confirmed by the experimental results (in Section 3.3) shown in Fig. 9u for 10,000 rad/s, and Fig. 10ac for 5000 rad/s.

Regarding bandwidth and amplitude response, the high-bandwidth characteristic of DBCC is reliably maintained for varying speed $N_s$ (Fig. 4e) and for parameter mismatch of the magnetic flux $F_{\phi}$ (Fig. 4d). Also, since $i^\text{ref} = 0$ A, the control performance is reliable with detuned $L_d$ (Fig. 4b). A mismatch of $R_{\phi}$ (Fig. 4a) not necessarily reduces the bandwidth, but may result in an increase (when $R^\text{est} > R_{\phi}$) or a decrease (when $R^\text{est} < R_{\phi}$) in the amplitude of $i_q$. In the case of detuned $L_q$ (Fig. 4c), the bandwidth of DBCC can be significantly reduced when $L^\text{est}_q > L_q$; whereas the amplification introduced at high frequencies may bring challenges for the stability of the $i_q$ control loop. It can thus be suggested, during the design of DBCC needs, to consider a smaller $dq$-axis inductances for stability reason, but not too small to avoid sacrificing the bandwidth.

Before discussing the phase response, it would be necessary to define the smallest possible phase shift for DBCC. Theoretically,
the reference current can be achieved only 2\(T_f\) after the reference change has been detected by the control or even longer (an example is given in Figs. 7c and d in Section 3.2.1) depending on the demanded current change and the available DC bus voltage. As a result, when operating below the voltage limitation, the phase shift of a signal is proportional to its frequency and equals to the frequency multiplied by \(2T_f\). This calculation is supported by almost all the phase plots in Fig. 4 apart from the case of the detuned \(L_q\) (Fig. 4c). Considering the worst case when \(L_q^\text{est} = 0.5L_q\) in Fig. 4c, the phase lagging for 5000 rad/s is 85.7° (=1.496 rad) which is 28.4° more than the theoretical value (57.3°). However, if we convert the phase delay into a delay in time (i.e. 1.496/5000 = 299 µs), it can be seen that the delay is <1.5 times the theoretical value (27°). It is interesting to note that the delay introduced by DBCC is affected mainly by detuned inductances; however, no significant influence is noted when the detuned inductance is within a reasonable range (50–150% of the real value).

### 3.1.2 Steady-state error plot:

For mapping the steady-state error plots as shown in Fig. 5, both the reference current \(i_{d}^{\text{ref}}\) and motor speed \(N_c\) are set to be constant and the error at steady state in the response current \(i_d\) is calculated using (35) and (36).

Again, the procedure is repeated for different settings of reference current value, motor speed, detuned motor parameter, and inverter dead time. Fig. 6 shows the influence of switching devices dead time on the steady-state errors in current responses using DBCC. The sampling frequency of DBCC is fixed at 10 kHz while the dead time varies from 0 to 10 μs.

\[
\text{steady-state error for } i_d = \frac{i_d - \bar{i}_d^{\text{est}}}{\sqrt{(\bar{i}_d^{\text{est}})}^2 + \bar{i}^2} \times 100 \tag{35}
\]

\[
\text{steady-state error for } i_q = \frac{i_q - \bar{i}_q^{\text{est}}}{\sqrt{(\bar{i}_q^{\text{est}})}^2 + \bar{i}^2} \times 100 \tag{36}
\]

Considering that practically the machine parameters are tuned before actual operation, the mismatch in parameters is likely to be within ±20%; also for surface-mounted PMSM, the \(d\)-axis reference current \(i_d^{\text{ref}}\) is normally controlled to be zero without field weakening. One important finding from Figs. 5 and 6 is that the steady-state offset of DBCC is the most sensitive to the magnetic flux \(F_m\) and the dead time of inverter. Though the d-q-plane offset can be compensated by many existing compensation methods [23], considering the results shown in Fig. 4e it may also be possible to compensate the remaining steady-state errors at a certain speed (non-zero) by tuning the estimated \(F_m\) in the control since the performance of the DBCC will not be influenced by \(F_m\).

Another interesting finding is that the results in Figs. 5 and 6 may give a clue for offline tuning of the machine parameters. For example, through increasing the estimated \(F_m\), \(L_d\), and \(R\) or decreasing the estimated \(L_q\) in the controller, the \(d\)-axis current \(i_d\) can be increased; similarly, the \(q\)-axis current \(i_q\) can be increased by increasing the estimated \(F_m\), \(L_q\), and \(R\) or decreasing the estimated \(L_d\) in the controller. As the dead time increases, the steady-state error positively increases in \(i_d\) and negatively increases in \(i_q\).

### 3.2 Effectiveness of the proposed DBCC

Turning now to the experimental tests of the proposed DBCC with two steps current prediction and rotor movement compensation, the machine parameters are tuned based on the findings in Section 3.1.2 and the results are as shown in Table 1. It is also to be noted that the insulated gate bipolar transistor voltage drops and dead time of inverter is compensated by using a look-up table as demonstrated in [23].

#### 3.2.1 Two steps prediction and operating at physical limit:

The current responses of DBCC to rectangular current references below and above voltage limitations are shown in Figs. 7b and c. In both cases, the rectangular current references are generated by the \(i_d\) current loop. Moreover, the current responses of the same rectangular reference using the proposed DBCC (with two steps current prediction) and the traditional DBCC (with only one step prediction) are compared in Fig. 7b to show the necessity of having two steps current prediction instead of one.

It can be seen from Fig. 7a that with properly tuned parameters and properly compensated inverter non-linearities, the proposed DBCC can achieve zero steady-state error in the response.

By comparing the measured current (green) and the first prediction current (red) in Fig. 7b, it can be seen that the first prediction current reaches its demand after \(2T_f\) and the measurement current, which is a mean current, reaches the demand after \(3T_f\). By comparing the measured current (green) for the proposed DBCC and the measured current (light blue) for the traditional DBCC (the first prediction is removed), it can be seen that an overshoot of 18.6% occurs without the first prediction. These results confirm the necessity of the first prediction as discussed in Section 2.1. The first prediction works to predict the real instantaneous current at 4.003s so that the controller can have a fairly accurate judgement of whether the demand has been achieved or not.

Fig. 7c shows that a longer settling time (as discussed in Section 3.1.1) of \(5T_f\) is required for DBCC to achieve a current step of 11 A with the peak value of the three-phase voltage limited at 450 V.

#### 3.2.2 Rotor movement compensation:

Figs. 8a and b show the effects of the rotor movement compensation in simulation, which starts at 0.01 s. The voltage references maintain a maximum two sampling time delays including a calculation step. Therefore, it is shown that the offsets start to decrease after two sampling periods. Similarly, the experimental results in Figs. 8c and d confirm the effectiveness of the proposed rotor movement compensation. The steady errors are calculated by (35) and (36), where mean values are used for \(i_d\) and \(i_q\) in the equations. By adding the proposed rotor position compensation, at 1000 rpm, the steady-state error in \(i_d\) reduces from 132 to 122%, whereas the error in \(i_q\) reduces slightly from −19 to −17%. Meanwhile, at 2930 rpm, the steady-state error in \(i_d\) reduces significantly from 81 to 27%, whereas the error in \(i_q\) reduces by half from −37 to −18%. It is obvious from both the simulation and the experiment that the rotor position compensation works for all speeds, but is much more effective at high speed as discussed in Section 2.3. Unfortunately, due to the parameter variations in motor and imperfect inverter compensation, the rotor position compensation method cannot remove all steady-state errors in the experiments.

### 3.3 Comparison between DBCC and PI

Furthermore, the performance of DBCC is compared with that of a conventional \(dq\) PI current regulator with a decoupling circuit. The bandwidth of the \(dq\) PI control loop used for comparison is designed to be around 900 Hz with classical single-input–

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**Table 1** Motor parameters

| Name of parameter | Value | Name of parameter | Value |
|-------------------|-------|-------------------|-------|
| Rated power       | 2.54 kW | \(d\)-axis stator inductances | 4.5 mH |
| Rated speed       | 3000 min\(^{-1}\) | \(q\)-axis stator inductances | 7.4 mH |
| Rated torque      | 9.4 Nm | \(L_d\) | 14 Ω |
| Rated voltage     | 400 V | \(R\) | 0.237 Wb |
| Rated current     | 5.9 A | \(I_s\) | 0.007 kgm\(^2\) |
| Number of pole pairs | 3 | – | – |

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Fig. 4 Simulation results of Bode diagrams of the q-axis current control loop ($i_{ref}^q = 0$ A, bias of $i_{ref}^q = 8.34$ A, $T_s = 100$ μs)

- $a$ $R_{est} = 0.5 \approx 1.5$ ($R_s$, $N_r = 3000$ rpm)
- $b$ $L_{est}^d = 0.5 \approx 1.5$ ($L_d$, $N_r = 3000$ rpm)
- $c$ $L_{est}^q = 0.5 \approx 1.5$ ($L_q$, $N_r = 3000$ rpm)
- $d$ $F_{est}^m = 0.5 \approx 1.5$ ($F_m$, $N_r = 3000$ rpm)

- $N_r = 10$, 3000, 6000 rpm
single-output control design methods. The no load test results for the current responses to a very high-frequency current reference (10,000 rad/s) are shown in Fig. 9. The no load and full load test results for the current responses to a high-frequency current reference (5000 rad/s) are shown in Fig. 10. The full load operation is tested using a commercial drive (ABB ACS800-11). The ASEA Brown Boveri ABB drive is operated under a speed control mode. The rotor speed is kept at 2500 rpm and the $i_{ref}^q$ is set at the rated value of 8.34 A. No tests are performed for frequencies higher than 10,000 rad/s since it is unlikely to have speed ripple at such high frequencies due to the low-pass filtering effect of the mechanical system.

For DBCC, it can be seen from Fig. 9a that the first current prediction is able to track the reference after two samples which is the smallest possible delay as discussed in Section 3.1.1. The phase shift is $115^\circ$ ($=2$ rad) which is exactly 10,000 rad/s multiplied by $2T_s$ ($=0.0002$ s). This last result confirms the phase plot in Fig. 4e. Additionally, the attenuation is about $-0.55$ dB. This is reasonably close to the simulated result for 10,000 rad/s since it is unlikely to have speed ripple at such high frequencies due to the low-pass filtering effect of the mechanical system.

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(−0.8 dB) as shown in Fig. 4e. Furthermore, the attenuation (−0.25 dB) and phase shift (57°) for 5000 rad/s in Fig. 4e are also supported by the experimental results as shown in Figs. 10a and c, where the phase lag of the first prediction current equals to 5000 rad/s multiplied by 2T_s (=0.0002 s), thus 57°, and the attenuation is about −0.61 dB.

When using a traditional PI regulator, as can be seen from Fig. 9b and Fig. 10b, d, the phase lag for 10,000 rad/s is about 200° (=3.5 rad), which can be also represented by a delay time of 350 μs (=3.5 rad divided by 10,000 rad/s) and the phase lag for 5000 rad/s is about 133° (=2.3 rad), which is equivalent to a delay time of 460 μs. Moreover, the signal is attenuated of −9.7 dB (out of bandwidth) at 10,000 rad/s and of −1.4 dB at 5000 rad/s. Though the bandwidth and response can be improved by a better designed PI, the fact that the delay of the PI control loop varies with frequency cannot be changed.

It is worth noting that, in the no load test, the PMSM is driven under speed control mode, whereas in the full load test, it is under current or torque control mode. Practically, depending on specific applications, PMSM can also be operating under position control mode. Cascaded outer loops are added when under speed or position control modes. These experimental tests confirm the same behaviour of the inner current loop under different control modes. When choosing between DBCC and PI, it is worth considering if only inner loop is used (i.e. in torque control mode), or outer loops are used (i.e. in speed and position control mode) since the

$$\text{Fig. 7 } \text{Experimental results of rectangular current responses of DBCC}$$

a With the proposed DBCC, $i_d^\text{ref}$ is 0.5 Hz (the inverter compensation is activated around 2 s)
b Comparison between the proposed DBCC and the traditional DBCC (without the first prediction), $i_d^\text{ref}$ is 0.5 Hz
c With the proposed DBCC and inverter compensation, $i_d^\text{ref}$ steps from 0 to 11 A (above voltage limitation)
d Module of reference voltage under condition e

$$\text{Fig. 8 } \text{Steady-state current errors due to rotor movement (i}_{q}^\text{ref} = 5 A )$$

a Simulation results $i_d^\text{ref} = 5 A$, $N_r = 100$ rpm
b Simulation results $i_d^\text{ref} = 5 A$, $N_r = 5000$ rpm
c Experimental results $i_d^\text{ref} = 0 A$, $N_r = 1000$ rpm, the rotor position compensation is activated at 0.005 s
d Experimental results $i_d^\text{ref} = 0 A$, $N_r = 2930$ rpm, the rotor position compensation is activated at 0.005 s

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steady-state error of current response can matter more in the torque control, and delay of the current loop may be more important when outer loop are present.

Analysing the results for DBCC and PI, considering the current references is likely to be a signal containing more than one frequency under speed or position control mode and that the operating point is likely to be within physical limits, DBCC can be a better choice than PI for the inner loop due to its capability of producing the same delay time for all frequencies. This can benefit the design of the outer loop since the delay of a dead beat current loop can be easily compensated than that of a PI current loop.

4 Conclusions

This paper has presented an improved high-performance DBCC including two steps current predictions and a novel rotor movement compensation method for PMSM motor drives and has provided a detailed performance assessment under different operating conditions supported by simulation and experimental results. The two steps current prediction is necessary to improve the accuracy of the current prediction during transients. The rotor movement compensation is particular useful at high speed for reducing steady-state errors in the current response due to the rotor position changing during each sampling period. Bode diagrams of the DBCC controlled system show that the bandwidth and phase response characteristic of DBCC are reasonably maintained even in the case of parameter detuning. Steady-state analysis of the DBCC is useful in the design phase in order to indicate the current responses obtained with mismatched parameters, so to facilitate control tuning. The comparison between DBCC and a traditional PI regulator in dq reference frame with decoupling circuit in terms of control dynamics and steady-state characteristics shows the main convenience of using DBCC, besides faster dynamics, is to have the fixed delay time for different frequencies. Therefore, the delay introduced by the DBCC loop can be compensated from the outer loop easily.

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