Abstract—In medium-voltage drives the switching frequency is limited to a few hundred Hz, necessitating control and modulation schemes capable of producing low levels of current and torque distortions at low switching frequencies. Model Predictive Direct Current Control (MPDCC) has emerged as a promising scheme for medium-voltage induction-motor drives. By forecasting the trajectory of the stator currents over a timespan known as the prediction horizon, MPDCC regulates the stator currents within a set of hysteresis bounds while minimising the inverter switching frequency. Despite the recent surge in popularity of predictive control, such schemes in the field of power electronics and drives were proposed already in the early 1980’s, Forced Machine Current Control (FMCC) is an early predictive current control scheme which shares several similarities with MPDCC. However, a comprehensive review and comparison of FMCC with the modern MPDCC scheme has never been carried out. Through simulation, it is shown that the steady state performance of MPDCC and FMCC is similar when the prediction horizon of the former is limited. However, when the prediction horizon is extended, the performance of MPDCC is shown to be superior to FMCC; the horizon of which is inherently restricted.

Index Terms—Model predictive control, current control, medium-voltage drive

I. INTRODUCTION

The inverter-fed induction machine has been a staple of industry for several decades. As expectations regarding motor-drive performance have increased, the traditional control and modulation schemes, which have been applied to machine-side inverters, have been superseded by a number of alternative schemes. Predictive control techniques have recently been applied to motor drives [1] - [5]. The primary attraction of such schemes is their ability to reduce the average switching frequency, and therefore switching losses, of the inverter, while maintaining acceptable levels of harmonic distortion in the current and torque of the machine. In addition to motor drives, predictive control has been applied to active filters, power factor correction and grid-connected converters [6] - [8].

Model Predictive Control (MPC), which was developed in the process control industry in the 1970’s [9], has received significant attention from industry. Model Predictive Direct Torque Control (MPDTC), which emerged several years ago, is a variant of MPC and an extension of Direct Torque Control (DTC), which features an online-optimisation process in the control of machine torque [1], [10]. Model Predictive Direct Current Control (MPDCC) is a more recent variant of MPC which treats the machine’s stator currents as the variables to be controlled [3], [11].

Although in the field of power electronics and drives MPC has only recently become popular, such schemes were proposed already in the early 1980’s. In particular, a Forced Machine Current Control (FMCC) scheme for induction motor drives [12] - [15], first described in 1983 by Holtz and Stadtfeld for the control of two-level inverters, shares a number of significant similarities with MPDCC. MPC-based schemes have been extensively compared with carrier-based Pulse Width Modulation (PWM), Space Vector Modulation (SVM) and Optimised Pulse Patterns (OPP) [4]. However, a review and comparison of FMCC against a modern MPC scheme has never been carried out. Such a comparison is useful, as it gives a clear picture of the benefits of modern MPC schemes, relative to early predictive control techniques. The aim of this paper is to therefore describe and benchmark FMCC against the modern MPDCC scheme through simulation of a Medium-Voltage (MV) induction motor drive. The trade off between switching frequency and distortion is a fundamental principle to power converters and will form the basis of comparison. Comparison will be made at steady state, with the key indicators of performance being the inverter switching frequency and the harmonic distortion of the machine’s stator currents and torque. The schemes have been compared through a MATLAB-based drive-system simulation which consists of a Neutral Point Clamped (NPC) three-level Voltage Source Inverter (VSI) driving a squirrel-cage Induction Motor (IM). Since the aim of this comparison is to gauge the quality of the control schemes in as general a sense as possible, effects such as deadtime, measurement noise and controller delay have been neglected.

II. DRIVE SYSTEM

A. System Setup

As shown in Fig. 1, the drive system used in this paper utilises an inner and outer control loop. The outer speed and
flux regulators are PI controllers which regulate the stator current reference value based on the speed and rotor flux references. The outer loop operates in the rotating dq reference frame. The inner predictive loop makes switching decisions based on state feedback and the current reference provided by the outer controllers. It is the inner loop which relates to the predictive control schemes described in this paper.

B. Inverter Model

The typical setup for a three-level NPC inverter driving an IM is shown in Fig 2. Each phase leg is able to assume one of three states, which may be represented by the integer variables \( u_a, u_b, u_c \in \{-1, 0, 1\} \). With three voltage levels per phase and three phases, there are \( 3^3 = 27 \) possible switching states of the form \( u_{abc} = [u_a, u_b, u_c]^T \). Within those states, 19 distinct voltage vectors exist which the inverter is capable of producing. The voltage vectors can be represented by transforming the switching states from the three-phase \( abc \) system to the orthogonal \( \alpha \beta \) system. The corresponding voltage at the machine terminals is given by

\[
v_{\alpha\beta} = \frac{V_{dc}}{2} P u_{abc}
\]

where \( v_{\alpha\beta} = [v_a, v_b]^T \), \( V_{dc} \) is the DC-link voltage and \( P \) is the transformation matrix

\[
P = \frac{2}{3} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} \end{bmatrix}.
\]

In the inverter under consideration all switching transitions are allowed except for those which involve switching between the upper and lower rails. For example, a transition from \( u_{abc} = [1 \ 1 \ 1]^T \) to \( [0 \ 0 \ 1]^T \) is allowed, whereas a transition to \([1 \ 1 \ 1]^T \) is not.

C. Induction Machine Model

The IM is modeled in the \( \alpha \beta \) reference frame, with the mechanical load assumed to be constant. The state system variables are the \( \alpha \beta \) components of the stator currents and rotor fluxes, \( i_{s\alpha}, i_{s\beta}, \Psi_{r\alpha} \) and \( \Psi_{r\beta} \) respectively. The input vector is the three-phase switch position \( u_{abc} \). The model parameters are the angular velocity of the rotor, \( \omega_r \), the resistances of the stator and rotor \( r_s \) and \( r_r \) respectively, the reactances of the stator and rotor \( x_{ls} \) and \( x_{lr} \) respectively, the mutual reactance \( x_m \), the mechanical inertia of the load, \( J \), and the mechanical torque of the load, \( T_l \). With \( x = [i_{s\alpha}, i_{s\beta}, \Psi_{r\alpha}, \Psi_{r\beta}]^T \) and \( u = [u_a, u_b, u_c]^T \) we can define the continuous-time state equation of the system as [16]

\[
\frac{dx}{dt} = Ax + Bu
\]

with \( A \) being the state matrix

\[
A = \begin{bmatrix}
-\frac{1}{r_s} & 0 & \frac{k_s}{r_s} & \frac{k_s \omega_r}{r_s} \\
0 & -\frac{1}{r_r} & -\frac{k_s}{r_r} & \frac{k_s \omega_r}{r_r} \\
\frac{x_m}{r_s} & 0 & -\frac{1}{r_s} & -\omega_r \\
0 & \frac{x_m}{r_r} & \omega_r & -\frac{1}{r_r}
\end{bmatrix}
\]

and \( B \) the input matrix

\[
B = \begin{bmatrix}
\frac{1}{r_s r_{s*}} & 0 & 0 & 0 \\
0 & \frac{1}{r_r r_{r*}} & 0 & 0
\end{bmatrix}^T \frac{V_{dc}}{2} P
\]

with the electromagnetic torque, \( T_e \), given by

\[
T_e = k_r (i_{s\alpha} \Psi_{r\alpha} - i_{s\beta} \Psi_{r\beta})
\]

and the relationship between rotor speed and torque

\[
\frac{d\omega_r}{dt} = \frac{1}{J} (T_e - T_l).
\]

The deduced parameters used in the above equations are the rotor coupling factor \( k_r = \frac{x_m}{x_s} \), total leakage factor \( \sigma = 1 - \frac{x_m^2}{x_s x_r} \), leakage reactance \( x_{\sigma} = \sigma x_s \), where \( x_s = x_{ls} + x_m \) and \( x_r = x_{lr} + x_m \), and equivalent resistance \( r_{\sigma} = r_s + k_r^2 r_r \). The deduced time constants include the transient stator time constant \( r_{s*} = \frac{x_s}{r_s} \), and the rotor time constant \( r_{r*} = \frac{x_r}{r_r} \).

Equations (3) – (7) provide a complete description of the dynamic behaviour of the IM when non-idealities such as magnetic saturation, the skin effect and variations in the rotor resistance are ignored.

D. Internal Model of the Controller

In order for a predictive control scheme to be implemented, a discrete-time model of the drive is required to serve as an internal prediction model for the controller. The model’s purpose is to predict the trajectory of the stator currents and rotor fluxes over as many sampling intervals as are required. Due to the fact that the rotor time constant greatly exceeds the length of a prediction horizon, the rotor speed is assumed to be constant within the prediction horizon and is treated as a model parameter rather than an additional variable [3], [11]. From the continuous time state equation of (3) – (5), and with the output vector defined as \( \gamma = [i_{s\alpha}, i_{s\beta}, \Psi_{r\alpha}, \Psi_{r\beta}]^T \), the following discrete-time model of the drive can be derived

\[
x(k + 1) = (I + T_s A)x(k) + T_s B u(k)
\]

\[
y(k) = C x(k)
\]

where \( I \) is the 4x4 identity matrix, \( T_s \) is the sampling time of 25\( \mu s \) and \( C \) is the output matrix

\[
C = \begin{bmatrix}
1 & 0 & 0 & 0 \\
0 & 1 & 0 & 0
\end{bmatrix}.
\]
III. CONTROL SCHEMES

Both MPDCC and FMCC replace the inner current control loop and modulator of traditional Field Oriented Control (FOC) schemes with a single online-optimisation stage. In doing so, MPDCC and FMCC are able to address the control and modulation problems simultaneously by directly manipulating the switching state of the inverter. A boundary area is defined around the stator current reference $i_s^*$, with the objective of the controller being to keep the stator current $i_s$ inside the boundary area while minimising the switching frequency or switching losses of the inverter. The dimension of the boundary area determines the current ripple, and therefore Total Demand Distortion (TDD), of the stator current, and in doing so sets the trade-off between switching frequency (or losses) and current distortion.

A. Model Predictive Direct Current Control (MPDCC)

In MPDCC, symmetrical hysteresis bounds are defined around each of the $abc$ stator currents. $\delta$, denotes the difference between the upper (or lower) bound and the reference. The aim of the controller is subsequently to keep each of the stator currents within the bounds while minimising the switching frequency of the inverter.

It is important to clearly differentiate the switching horizon $N_s$ from the prediction horizon $N_p$. The switching horizon refers to the number of switching events within the horizon, while the prediction horizon refers to the number of time-steps the controller looks forward, starting from the current time-step $k$. The switching horizon is composed of the elements ‘$S$’ and ‘$E$’, which stand for switch and extend, respectively. During the extension steps, which occur between and after the switching events (with the switching state held constant), the trajectory of the outputs is extended until one or more bounds are violated. Extension may either be exact, utilising the internal model of the controller, or an approximation based on linear or quadratic extrapolation or interpolation [17]. By utilising extension steps, a short switching horizon including only one to three ‘$S$’ events may result in a prediction horizon of 30 to more than 100 steps.

By lengthening the switching horizon, the controller is able to look further into the future and make better informed switching decisions. One could therefore expect an improvement in performance, for example a reduced switching frequency for the same current distortion, by extending the switching horizon from ‘eSE’ to ‘eSESESE’. Note that the lower case ‘e’ refers to an optional extension event at the beginning of the horizon.

At each time-step $k$ the set of allowable switching sequences forward in time is determined for the switching horizon $N_s$ based on the current switching state $u(k-1)$. For each sequence, the trajectory of the output currents forward in time is predicted using the internal model of the controller. For each sequence, the output trajectory must remain a candidate over the entire prediction horizon $N_p$. A candidate sequence is one for which all of the output variables are either feasible, or pointing in the correct direction. An output variable is feasible if it lies within its hysteresis bounds. Pointing in the correct direction denotes the instance where an output variable lies outside the bounds, but moves closer to it at every time-step of the prediction horizon. For a given switching horizon, each candidate sequence $U^i(k) = [u^i(k), u^i(k+1), \ldots, u^i(k+N_p)]$ yields an associated cost which can be determined from

$$C^i = \frac{1}{N_p} \sum_{\ell=k}^{k+N_p-1} \|u^i(\ell) - u^i(\ell - 1)\|_1$$

(10)

for minimisation of switching frequency, or

$$C^i = \frac{E^i}{N_p}$$

(11)

for minimisation of switching losses, where $E^i$ is the total switching energy loss over the prediction horizon. A detailed description of the calculation of switching losses is given in [10]. The switching sequence $U^i(k)$ with the minimal cost is subsequently determined

$$i = \arg\min C^i$$

(12)

with the switching state $u(k) = u^i(k)$ applied. The horizon is subsequently shifted one step forward, with the process repeated at $k+1$. [3] and [10] provide further details on the MPDCC problem and control procedure.

In order to simplify the problem, MPDCC can be reformulated in the $\alpha\beta$-plane. By transforming the $abc$ hysteresis bounds to the $\alpha\beta$-plane, a hexagonal boundary area centred on the reference current $i_s^*$ results. The problem can be further transformed to the $dq$-reference plane, which results in a hexagonal-boundary area centred on the stationary reference current $i_s^{*_{dq}}$. Due to the synchronously-rotating nature of the $dq$-plane, the hexagonal boundary is not fixed in space. As shown in Fig. 3a, the boundary area rotates at the angular speed of the rotor flux, $\omega_s$, in an anti-clockwise direction.

Fig. 3a also illustrates an example output prediction for an arbitrary switching sequence over a switching horizon of ‘SESESE’. At time-step $k$ switching is necessitated due to imminent violation of the boundary area. Following extension, switching subsequently occurs at time-steps $k + N_{p1}$ and $k + N_{p2}$. The final extension leg of the horizon results in a total prediction horizon of length $N_p$ time-steps. Note that in Fig. 3a all trajectories are referred to the position of the boundary area at time-step $k$.

B. Forced Machine Current Control (FMCC)

Unlike MPDCC, where the bounds are defined around each of the phase currents separately, FMCC as described in [12], utilises instead a radial boundary area defined around the stator current reference on the $\alpha\beta$-plane. $\delta_r$ denotes the radius of the boundary area. Like MPDCC, FMCC aims to keep the output current regulated about the reference while minimising the switching frequency of the inverter. The problem can be simplified by transforming it to the $dq$-plane, which results in a fixed reference current $i_s^{*_{dq}}$ and boundary area as shown in Fig. 3b.

Unlike MPDCC, where the switching horizon is variable and can be made up of a variety of ‘$S$’ and ‘$E$’ elements, the switching horizon for FMCC is effectively limited to ‘SE’, thus restricting the length of the prediction horizon. The control procedure for FMCC is similar to that of MPDCC. At each time-step $k$ the stator current of the machine is sampled, and any intersection of the current trajectory with the boundary circle detected. When an intersection is detected (meaning the stator current vector lies outside the boundary area) the set of allowable switching states which can be applied to the inverter...
at time-step \( k \) is determined based on the current switching state \( u(k - 1) \). For each allowable switching state which can be applied at time-step \( k \), the trajectory of the output currents is predicted forward in time using a linear approximation technique as described in [12] - [14]. During extension, the switching state is held constant until another intersection of the boundary area occurs. Each candidate switching state \( u'(k) \) will yield a prediction horizon of length \( N_p \), where \( N_p \) is the number of time-steps from the switching time-step \( k \) to the next intersection of the boundary. Since FMCC minimises the switching frequency of the inverter, the cost associated with each switching state can be determined from (10) with the optimal state being that which minimises (12). Fig. 3b illustrates an example output prediction for an arbitrary switching state. At time-step \( k \) switching is necessitated due to the output current intersecting the boundary circle. The trajectory of the output current is predicted for the candidate switching state \( u(k) \), with extension of the current trajectory resulting in a prediction horizon of length \( N_p \).

In addition to the FMCC scheme outlined above, several variant FMCC schemes were proposed. In [14], Holtz and Stadtfeld proposed a method of optimisation by double prediction, where the controller preemptively selects a new switching state in order to avoid intersection of the boundary. This has the benefit of enforcing strict observance of the stator current boundary, as is the case for MPDCC with a horizon of the form 'eSE'. In [13] and [15], a variant of FMCC which utilises a rectangular boundary area around the stator current reference was proposed. This allows the torque and current distortion to be controlled with a large degree of independence, through variation of the height and width of the rectangle. The use of a rectangular boundary area parallels MPDTC, where the boundary area is defined such that the machine torque is directly controlled.

**IV. PERFORMANCE EVALUATION**

This section summarizes the performance of MPDCC and FMCC for simulations carried out using the drive system outlined in Section II. A 3.3 kV, 50 Hz, 2 MVA squirrel-cage IM has been used, a typical machine used in the MV drive industry. The NPC inverter has a total nominal DC-link voltage of 5.2 kV. ABB’s 35L4510 4.5 kV 4 kA Integrated Gate Commutated Thyristor (IGCT) is used for all switches. ABB’s 10H4520 fast recovery diode is used for all diodes. A summary of the machine and inverter parameters can be found in Table I. The p.u. system uses base values of \( V_B = \sqrt{2/3} V_{rat} = 2694 \) V, \( I_B = \sqrt{2} I_{rat} = 504 \) A, and \( f_B = f_{rat} = 50 \) Hz.

In order to gauge the performance of MPDCC and FMCC, three well known modulation schemes – carrier-based PWM, SVM and OPP – have been included for comparison, which have been used for benchmarking in previous papers on MPC in [3] and [4]. The carrier-based PWM and SVM modulation schemes have been studied extensively and are common in industry, and as such provide useful performance references for the predictive control schemes. OPPs, which are calculated off-line, minimize the current distortion for a given pulse number (switching frequency). This is done through optimisation of the switching angles for all possible operating points over a quarter of a fundamental period. The steady state performance of OPP provides a useful benchmark for predictive schemes, which aim to achieve optimal results through online optimisation.

Simulations were run at 60% speed and full torque at steady state. All simulations have been carried out under the assumption of a fixed neutral point. MPDCC simulations have been run with switching horizons of 'eSE', 'eSESE', and 'eSESESE', and with the cost function penalising the inverter switching frequency. For FMCC, all simulations have been run with a circular boundary area and single prediction. In addition, all extension steps for both MPDCC and FMCC

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**TABLE I: Rated values (left) and parameters (right) of the drive model used**

| Induction Motor |
|-----------------|-----------------|-----------------|
| Voltage         | 3300 V          | \( r_s \)       | 0.0108 pu       |
| Current         | 356 A           | \( r_r \)       | 0.0091 pu       |
| Real power      | 1.587 MW        | \( x_{ds} \)    | 0.1493 pu       |
| Apparent power  | 2.035 MVA       | \( x_{tr} \)    | 0.1104 pu       |
| Frequency       | 50 Hz           | \( x_m \)       | 2.3489 pu       |
| Rotational speed| 596 rpm         |                 |                 |

<table>
<thead>
<tr>
<th>Inverter</th>
</tr>
</thead>
<tbody>
<tr>
<td>DC-link voltage</td>
</tr>
<tr>
<td>( V_{dc} )</td>
</tr>
</tbody>
</table>
Fig. 4: Performance trade-off for MPDCC with switching horizons of ‘eSE’, ‘eSESE’ and ‘eSESESE’ and FMCC. Each plot shows the variation of a different performance metric as the boundary dimensions (δi for MPDCC, δr for FMCC) are varied from 5% to 20% p.u. at increments of 0.25%. The operating point is ωe = 0.6 p.u, Te = 1 p.u. At all points, each scheme was simulated over 20 fundamental periods.

Table II: Comparison of FMCC with MPDCC. The table shows the average percentage values of current distortion ITDD, torque distortion TTDD and switching frequency fs, relative to FMCC. Each value represents the average from across the range of bound widths.

<table>
<thead>
<tr>
<th>Control scheme</th>
<th>Switching horizon</th>
<th>fs [%]</th>
<th>ITDD [%]</th>
<th>TTDD [%]</th>
</tr>
</thead>
<tbody>
<tr>
<td>FMCC</td>
<td>-</td>
<td>100</td>
<td>100</td>
<td>100</td>
</tr>
<tr>
<td>MPDCC</td>
<td>eSE</td>
<td>96.2</td>
<td>97.8</td>
<td>100</td>
</tr>
<tr>
<td>MPDCC</td>
<td>eSESE</td>
<td>77.6</td>
<td>91.3</td>
<td>77.0</td>
</tr>
<tr>
<td>MPDCC</td>
<td>eSESESE</td>
<td>77.8</td>
<td>90.2</td>
<td>73.0</td>
</tr>
</tbody>
</table>
TABLE III: Comparison of FMCC with MPDCC, carrier-based PWM, SVM and OPP. The first comparison is at a current TDD of approximately 4%, while the second is at a switching frequency of about 180 Hz. The first section summarises the control settings, the second the absolute values summarising performance, and the third section the performance values relative to carrier-based PWM. \( f_c \) denotes the carrier frequency of PWM/SVM, while \( d \) denotes the pulse number for OPP.

<table>
<thead>
<tr>
<th>Control scheme</th>
<th>Control setting</th>
<th>Switching horizon</th>
<th>( f_{sw} ) [Hz]</th>
<th>( T_{TDD} ) [%]</th>
<th>( T_{TDD} ) [%]</th>
<th>( P_{sw} ) [kW]</th>
<th>( f_{sw} )</th>
<th>( T_{TDD} ) [%]</th>
<th>( T_{TDD} ) [%]</th>
<th>( P_{sw} ) [kW]</th>
</tr>
</thead>
<tbody>
<tr>
<td>PWM ( f_c = 780 ) Hz</td>
<td>-</td>
<td>-</td>
<td>400</td>
<td>3.97</td>
<td>1.34</td>
<td>9.40</td>
<td>100</td>
<td>100</td>
<td>100</td>
<td>100</td>
</tr>
<tr>
<td>SVM ( f_c = 570 ) Hz</td>
<td>-</td>
<td>-</td>
<td>300</td>
<td>3.92</td>
<td>1.59</td>
<td>6.88</td>
<td>74.1</td>
<td>98.7</td>
<td>119</td>
<td>73.2</td>
</tr>
<tr>
<td>FMCC ( \delta_c = 0.0525 )</td>
<td>-</td>
<td>-</td>
<td>326</td>
<td>4.05</td>
<td>3.17</td>
<td>8.47</td>
<td>80.5</td>
<td>102</td>
<td>237</td>
<td>90.1</td>
</tr>
<tr>
<td>MPDCC ( \delta_c = 0.0575 ) eSE</td>
<td>-</td>
<td>-</td>
<td>292</td>
<td>4.02</td>
<td>3.21</td>
<td>8.30</td>
<td>72.1</td>
<td>101</td>
<td>240</td>
<td>88.9</td>
</tr>
<tr>
<td>MPDCC ( \delta_c = 0.0600 ) eSE</td>
<td>-</td>
<td>-</td>
<td>250</td>
<td>3.94</td>
<td>2.70</td>
<td>6.82</td>
<td>61.7</td>
<td>99.2</td>
<td>201</td>
<td>72.6</td>
</tr>
<tr>
<td>MPDCC ( \delta_c = 0.0600 ) eSE</td>
<td>-</td>
<td>-</td>
<td>243</td>
<td>3.98</td>
<td>2.67</td>
<td>6.68</td>
<td>60.0</td>
<td>100</td>
<td>199</td>
<td>68.9</td>
</tr>
<tr>
<td>OPP ( d = 7 )</td>
<td>-</td>
<td>-</td>
<td>210</td>
<td>4.04</td>
<td>2.71</td>
<td>5.24</td>
<td>52.8</td>
<td>102</td>
<td>202</td>
<td>55.7</td>
</tr>
</tbody>
</table>

At a switching frequency of about 180 Hz, MPDCC with a short horizon yields a current TDD 10.4% lower and torque TDD 10.6% lower than FMCC. With a long horizon, these reductions are 32% and 44%, respectively. However, FMCC improves on the current TDD of PWM by 23.6% and achieves a similar level of current TDD to SVM. When compared to OPP, the predictive schemes all produce higher levels of current and torque TDD.

V. CONCLUSION

This paper has presented a review of FMCC and a comparison against the more recent MPDCC. Initially proposed for two-level inverters, FMCC has here been extended to a three-level topology with a model-based prediction technique utilised, in order to allow comparison with MPDCC. An internal control model for the drive system has been derived, and the conceptual similarity between FMCC and MPDCC summarised. FMCC has been shown to perform to a similar level as MPDCC when the horizon is short. However, as the switching horizon is lengthened, MPDCC performs to a higher level than FMCC. Despite this, it is important to note that at the time when FMCC was conceived in the early 1980s, the computational power required for long switching horizons was unavailable, and it remains a very good control scheme when computational resources are restricted. In addition, the performance of FMCC could be improved with the addition of double prediction. Both FMCC and MPDCC have been shown to perform to a higher level than carrier-based PWM in terms of current TDD and switching frequency. The torque TDD is in general poorer for the predictive control schemes than for both carrier-based PWM and SVM. If minimisation of torque TDD is a priority, then MPDTC or FMCC with a rectangular boundary area are better options, as the hysteresis bounds for such schemes are shaped to minimise the torque ripple.

REFERENCES


