Article

Three-Level NPC Inverter-Fed IM Drives under PTC, Minimizing the Involved Voltage Vectors and Balancing the DC Bus Capacitor Voltages

Wiem Zouari, Imen Nouira El Badsi, Bassem El Badsi and Ahmed Masmoudi *

Abstract: The paper presents a comparative study of the steady-state and transient behaviors of three-level neutral-point clamped (3L-NPC) inverter-fed induction motor (IM) drives under the control of three predictive torque control (PTC) schemes: the conventional one (C-PTC) and two new strategies involving selected stator voltage vectors (SVs), namely: (SV-PTC1) and (SV-PTC2). Compared to the C-PTC, the latter enable a reduction of the number of voltage vectors as well as the weighting factors. The introduced PTCs mainly differ by the cost function, which is more simple in the case of SV-PTC2. However, SV-PTC1 allows a systematic clamping of each stator phase to the DC bus voltage, at a low level of 60 degrees, and at a high level at 60 degrees per cycle, while such a clamping is arbitrary in the case of SV-PTC2. Simulations prove the higher performances of the introduced SV-PTCs over the C-PTC ones, in terms of the reduction of the current distortion and of the torque ripple. Simulation results were validated by the experiments.

Keywords: predictive torque control; three-level neutral-point clamped inverter; induction motor; reduction of the active voltage vectors; stator phase clamping; neutral point voltage balance

1. Introduction

The introduction of the direct torque control (DTC) induction motor (IM) drives in the mid-1980s [1,2] has enabled access to higher torque dynamics over the ones exhibited by field-oriented control (FOC) [3]. Moreover, the DTC has better robustness against motor parameter variations. Finally, DTC implementation schemes are simpler than those of FOC. In spite of their potentialities, DTC strategies suffer from high ripples affecting the torque and flux responses, unless controlled commutation frequency strategies are adopted. Nevertheless, this leads to complicated implementation schemes.

In recent years, model predictive control (MPC) strategies have been investigated for the control of different variable speed drives equipped with induction, permanent magnets, and switched reluctance machines [4–6]. They have been widely implemented in sustainable mobility applications, covering electric [7–10] and hybrid [11–14] road vehicles, as well as aerospace ones [15,16]. MPC strategies have also been selected for the control of sustainable energy systems, especially wind turbines [17–19].

The control of output features is achieved using the electric machine model in order to formulate the so-called cost function. However, a major problem of MPC schemes is the weighting factors involved in the cost functions; the drive performances are directly allied to their suitable online adjustments [20]. To date, there is a lack of theoretical support dedicated to the assessment of weighting factors. This makes their online adjustments time-consuming procedures, especially when the number of involved voltage vectors is high, such as in the case of three-level inverters. The latter requires a specific criterion while designing the cost functions of MPC strategies, which is the neutral point voltage balance [21–23].
Several MPC strategies have been implemented in three-level neutral point clamped (3L-NPC) inverter-fed IM drives. These strategies mainly differ by the variables involved in their cost functions, such as the electromagnetic torque, the stator flux, and the stator current. This represents a state-of-the-art topic, which has been reviewed by several teams. A literature review of recent works is reported hereunder.

MPC strategies whose cost functions mainly involve the electromagnetic torque and the stator flux, yielding the so-called predictive torque control (PTC), are first discussed. In [24], Habibullah et al. took advantage of the large number of voltage vectors generated by 3L-NPC inverters to improve the quality of the features of an IM drive under a conventional PTC, with a comparison with those yielded by FOC and DTC schemes. In order to reduce the required CPU time, the same team extended their work by considering the reduction of the involved voltage vectors from 27 to 14, considering the case of an encoder to measure the speed in [25], and the case of a speed sensorless in [26]. In both works, the neutral point voltage balance was taken into consideration in the cost function. In [27], Zhang et al. proposed an approach to design the weighting factors of the cost function of a PTC intended for the control of 3L-NPC inverter-fed IM drives. In order to reduce the CPU time required for the online tuning procedure, the weighting factors were adjusted using a fuzzy logic-based numerical approach. In [28], Osman et al. proposed an approach to reduce the number of voltage vectors from 27 to 17 in the prediction stage of a PTC strategy for 3L-NPC power inverter-fed IM drives, resulting in a 30% decrease in the execution time compared to the one required by the conventional PTC, with similar dynamic and steady-state performances. In [29], Bandy and Stumpf considered the PTC of 3L-NPC inverter-fed IM drives using separate cost functions for the torque and the stator flux. In order to reduce the CPU time required for the selection of the voltage vector, they proposed a hybrid sorting algorithm consisting of two sorting networks and a merging step.

Another class of MPC strategies, characterized by cost functions independent of the stator flux weighting factor, was also implemented in 3L-NPC inverter-fed IM drives. In [30], Xiao et al. proposed a predictive flux and neutral-point voltage control scheme for a 3L-NPC inverter-fed IM drive. The proposed strategy considered a cascaded structure of predictive control with the stator flux predicted and evaluated using 13 switching states in the maximum, and the neutral-point voltage predicted and minimized for two switching states, according to a predefined limit. In [31], Xiao et al. introduced another cascaded predictive control scheme involving three separate cost functions aimed at a sequential control of the stator flux, the neutral point voltage, and the inverter switching loss. In [32], Osman et al. developed a predictive flux control strategy based on the prediction of the reference stator flux vector. In the first stage, the evaluation of the six long voltage vectors was carried out. The optimal long vector enabled the definition of an optimal hexagon containing 11 medium and short voltage vectors, resulting in a reduction of admissible voltage vectors to 17 among the 27 permissible ones.

The third class of MPC strategies basically considered stator current dependent cost functions. In [33], Jun et al. proposed a predictive strategy that simultaneously controlled the output current and the neutral point voltage of the 3L-NPC inverter, by generating an output phase voltage resulting from the sum of a reference voltage and an offset one. The reference voltage was predicted by applying the voltage equation of the connected load. The offset voltage was the difference between the voltages of the upper and lower DC-link capacitors. In [34], Wang et al. developed a third-class MPC strategy with an emphasis on the enhancement of the robustness of the 3L-NPC inverter IM drive against measurement noises and IM parameter variations. A Ляпунов function was designed in order to prove the stability of the proposed MPC strategy. In [35], Begh et al. developed the concept and the analysis of a MPC strategy allied to the optimization of the control pulse patterns for medium-voltage 3L-NPC inverter-fed IM drives. The proposed scheme enabled high transient behavior of the drive along with ripple-free stator current waveforms. The same authors treated a real-time implementation of the introduced MPC strategy, which was carried out using the hardware-in-loop (HIL) environment [36].
In this paper, the performances of 3L-NPC inverter-fed IMs were investigated under the control of two new PTC strategies and were compared to those exhibited by the conventional one. The introduced PTC strategies have the merit to:

- Minimize the number of applied active voltage vectors to six.
- Reduce the commutation of the inverter power switches thanks to the clamping of the stator phase terminals to the DC bus voltage.
- Simplify the cost functions by the elimination of one or more weighting factors.

The comparison study was achieved considering simulations of the steady-state operation and of the transient behavior of a 3L-NPC inverter-fed IM. The simulation results were validated by experiments carried out using a developed test bench.

2. Induction Motor Modeling

The IM electric and magnetic equations involving variables expressed in the αβ-stationary frame are provided in [37]. Considering the same frame, the dynamic of the stator current vector \( I_s = I_{sαβ} \) is governed by the following equation:

\[
\tau_σ \frac{dI_s}{dt} + I_s = \frac{1}{r_σ} V_s + \frac{k_r}{r_τ} \left( \frac{1}{τ_r^2} I_2 - \omega_r J^r \right) Ψ_r
\]

where \( ω_r \) is the rotor angular frequency, \( V_s \) is the stator voltage vector, \( Ψ_r \) is the rotor flux vector, \( I_2 \) is the identity matrix of rank 2, \( J \) is an orthogonal matrix defined as:

\[
J = \begin{bmatrix}
0 & -1 \\
1 & 0
\end{bmatrix}
\]

and where:

\[
\begin{align*}
\tau_σ &= r_s + k_r^2 r_r \\
l_σ &= σ l_s
\end{align*}
\]

with \( r_s \) and \( r_r \) are the stator and rotor phase resistances, respectively, \( l_s \) and \( l_r \) are the stator and rotor phase self-inductances, respectively, \( M \) is the stator–rotor mutual inductance, \( k_r = \frac{M}{l_r} \) is the rotor coupling factor, and \( σ = 1 - \frac{M^2}{l_s l_r} \) is the total leakage factor.

The mechanical equation under the motor operation is expressed as follows:

\[
T_{em} - T_l = J \frac{dΩ_m}{dt} + f Ω_m
\]

where \( T_l, Ω_m, N_p, J, \) and \( f \) represent the load torque, the rotor speed, the number of pole pairs, the total inertia, and the viscous friction coefficient, respectively, and \( T_{em} \) is the electromagnetic torque with:

\[
T_{em} = \frac{3}{2} N_p Ω_m \{ Ψ_s^r I_s \}
\]

3. 3L-NPC Inverter Modeling

The circuit diagram of the three-phase 3L-NPC inverter, also called the B12 inverter, is presented in Figure 1. It is composed of three legs. Each one includes two pairs of series-connected insulated gate bipolar transistor (IGBT) power switches (upper and lower), noted \( K_{pi} \) where \( p \) refers to the stator a–b–c phases and \( i \) (integer with \( 1 ≤ i ≤ 4 \)) refers to the switch number, in parallel with four freewheel diodes. It also includes two clamping diodes per leg, enabling the clamp of the DC voltage at three desired levels per phase: \( \frac{V_c}{2}, 0, -\frac{V_c}{2} \). Referring to Figure 1, one can notice that the DC voltage \( V_{dc} \) is split into two equal levels using two series-connected capacitors, \( C_1 \) and \( C_2 \), with a middle point in between, the so-called “neutral point". In each leg, the state of the first and the third switches are
controlled complementary. For instance, if the switch \( K_{p1} \) is in the ON state (binary “1”), its complement \( K_{p3} \) is in the OFF state (binary “0”), and inversely. The same rule is applied for the pair of switches \( K_{p2} \) and \( K_{p4} \).

![Figure 1. Three-phase three-level NPC inverter circuitry.](image)

Let us call \( C_p \) the combination of the states of the power switches of a given leg, with: \( C_p = [K_{p1}, K_{p2}, K_{p3}, K_{p4}] \). Taking into account the complementarity in the states of the pair of switches and in order to obtain the three-level output voltages, only these three combinations are allowed:

- \( C_p = [0, 0, 1, 1] \) which can be simplified to \( C_p = -1 \).
- \( C_p = [0, 1, 1, 0] \) which can be simplified to \( C_p = 0 \).
- \( C_p = [1, 1, 0, 0] \) which can be reduced to \( C_p = 1 \).

Accounting for the possible combinations of the states of the power switches of the three legs, the 3L-NPC inverter can generate 27 voltage vectors, as shown in Figure 2.

![Figure 2. Voltage vectors generated by the 3L-NPC inverter in the (α-β) plane.](image)

Depending on their magnitude, the 27 voltage vectors could be classified into four categories [38]:

- \( \alpha \) and \( \beta \) are the coordinates of the \( \alpha-\beta \) plane.
- Sector 1 (0° ≤ \( \alpha < 60° \))
- Sector 2 (60° ≤ \( \alpha < 120° \))
- Sector 3 (120° ≤ \( \alpha < 180° \))
- Sector 4 (180° ≤ \( \alpha < 240° \))
- Sector 5 (240° ≤ \( \alpha < 300° \))
- Sector 6 (300° ≤ \( \alpha < 360° \))
Three zero voltage vectors with a magnitude of 0 V;
- Twelve small voltage vectors with a magnitude of \( \frac{1}{\sqrt{6}} V_{dc} \) (colored in red in Figure 2);
- Six medium voltage vectors with a magnitude of \( \frac{1}{\sqrt{2}} V_{dc} \) (colored in green in Figure 2);
- Six large voltage vectors with a magnitude of \( \sqrt{\frac{3}{2}} V_{dc} \) (colored in blue in Figure 2).

The increase in the number of vectors, compared to the two-level inverter ones, offers
further freedom degrees in the vector selection, thanks to which a reduction in the ripple
(within the different feature waveforms) is gained. The twenty-seven state combinations of
the three legs and the \( \alpha-\beta \) components of the resulting voltages with their magnitudes and
angles are summarized in Table 1. These vectors are obtained under the assumption that
the net DC bus voltage is symmetrically split across the two DC bus capacitors, C1 and C2
\( V_{C1} = V_{C2} = \frac{V_{dc}}{2} \).

### Table 1. Switch state combinations of the three legs, \( \alpha-\beta \) components of the resulting vectors with
their magnitudes and angles.

| \( V_i \) | \( S_a \) | \( S_b \) | \( S_c \) | \( V_{ai} \) | \( V_{bi} \) | \( \| V_i \| \) | \( \theta_i \) |
|---|---|---|---|---|---|---|---|
| \( V_0 \) | 0 | 0 | 0 | 0 | 0 | 0 | 0 |
| \( V_7 \) | 1 | 1 | 1 | 0 | 0 | 0 | 0 |
| \( V_{14} \) | -1 | -1 | -1 | 0 | 0 | 0 | 0 |
| \( V_1 \) | 1 | 0 | 0 | \( \frac{1}{\sqrt{6}} V_{dc} \) | 0 | \( \frac{1}{\sqrt{6}} V_{dc} \) | 0 |
| \( V_8 \) | 0 | 1 | -1 | \( \frac{1}{\sqrt{6}} V_{dc} \) | 0 | \( \frac{1}{\sqrt{6}} V_{dc} \) | 0 |
| \( V_2 \) | -1 | 1 | 0 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{\pi}{3} \) |
| \( V_9 \) | 0 | 0 | 1 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{2\pi}{3} \) |
| \( V_{10} \) | -1 | 0 | -1 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{2\pi}{3} \) |
| \( V_4 \) | 0 | 1 | -1 | \( \frac{1}{\sqrt{6}} V_{dc} \) | 0 | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{\pi}{2} \) |
| \( V_{11} \) | -1 | 0 | 0 | \( \frac{1}{\sqrt{6}} V_{dc} \) | 0 | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{\pi}{2} \) |
| \( V_5 \) | 0 | 0 | 1 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{-2\pi}{3} \) |
| \( V_{12} \) | -1 | 1 | 0 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{-2\pi}{3} \) |
| \( V_6 \) | 1 | 0 | 1 | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{-\pi}{3} \) |
| \( V_{13} \) | 0 | 1 | 0 | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{1}{\sqrt{6}} V_{dc} \) | \( \frac{-\pi}{3} \) |
| \( V_{21} \) | 1 | 0 | -1 | \( \sqrt{\frac{3}{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{\pi}{6} \) |
| \( V_{22} \) | 0 | 0 | 1 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{\pi}{6} \) |
| \( V_{23} \) | -1 | 0 | -1 | \( \sqrt{\frac{3}{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{5\pi}{6} \) |
| \( V_{24} \) | -1 | 1 | 0 | \( \sqrt{\frac{3}{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{-5\pi}{6} \) |
| \( V_{25} \) | 0 | 1 | -1 | 0 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{-\pi}{2} \) |
Table 1. Cont.

| \( V_i \) | \( S_a \) | \( S_b \) | \( S_c \) | \( V_{ai} \) | \( V_{bi} \) | \( \| V_i \| \) | \( \theta_i \) |
|----------|-----|-----|-----|---------|---------|----------|---------|
| \( V_{26} \) | 1-1 0 | \( \sqrt{\frac{2}{3}} V_{dc} \) | \( -\frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \(-\frac{\pi}{6}\) |
| \( V_{15} \) | 1-1-1 | \( \sqrt{\frac{2}{3}} V_{dc} \) | 0 | \( \sqrt{\frac{2}{3}} V_{dc} \) | 0 |
| \( V_{16} \) | 1 1-1 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{\pi}{6} \) |
| \( V_{17} \) | -1 1-1 | \( -\frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{\pi}{6} \) |
| \( V_{18} \) | -1 1 1 | \( -\sqrt{\frac{2}{3}} V_{dc} \) | 0 | \( \sqrt{\frac{2}{3}} V_{dc} \) | \( \pi \) |
| \( V_{19} \) | -1-1 1 | \( -\frac{1}{\sqrt{2}} V_{dc} \) | \( -\frac{1}{\sqrt{2}} V_{dc} \) | \( \frac{2\pi}{3} \) |
| \( V_{20} \) | 1-1 1 | \( \frac{1}{\sqrt{2}} V_{dc} \) | \( -\frac{1}{\sqrt{2}} V_{dc} \) | \( \sqrt{\frac{2}{3}} V_{dc} \) | \( -\frac{\pi}{6} \) |

Large voltage vectors

4. PTC Strategies Dedicated to 3L-NPC Inverter-Fed IM Drives

In this section, the bases of the conventional PTC strategy (C-PTC) are recalled. Then, two PTC strategies, namely SV-PTC1 and SV-PTC2, are introduced and their performances investigated and compared to the conventional PTC ones. These have the merit to minimize the number of the involved voltage vectors in the estimation, prediction, and actuation steps, thanks to a novel selection approach.

4.1. Conventional PTC Basis

The C-PTC implementation scheme is illustrated in Figure 3. It includes three major steps:

1. The estimation of the rotor and stator fluxes at time \( kT_s \) with \( T_s \) is the sampling period,
2. The prediction of the stator and rotor fluxes, the stator current, and the electromagnetic torque at time \( (k+1)T_s \). For the sake of a time delay compensation in a real-time implementation, a second prediction is required at time \( (k+2)T_s \),
3. The actuation following the minimization of the cost function.

![Figure 3. Implementation scheme of the C-PTC strategy.](image-url)
The prediction of the stator flux $\Psi_s$, the stator current $I_s$, and the electromagnetic torque $T_{em}$ involves twenty-four active voltage vectors $V_i$, and three zero ones $V_0, V_7, V_{14}$.

Regarding the cost function $g$, it represents a combination of objective sub-functions whose minimizations enable the determination of the optimal voltage vector $V_{opt}$ to be applied at $(k + 1)T_s$.

4.1.1. Rotor and Stator Flux Estimation

The rotor flux vector $\Psi_r$ can be expressed in the $dq$ frame linked to the rotor ($\omega_r = \omega_s - \omega_m = 0$), in terms of the stator current vector $I_s$, as follows:

$$\tau_r \frac{d\Psi_{rdq}}{dt} + \Psi_{rdq} = MI_{sdq} \quad (6)$$

Applying the bilinear transform [39], the discrete-time expression of Equation (6) is given by:

$$\Psi_{krdq} = k_1(I_{ksdq} + I_{k-1sdq}) + k_2\Psi_{k-1rdq} \quad (7)$$

where

$$\begin{cases} k_1 = \frac{MT_s}{2\tau_r + T_s} \\
 k_2 = \frac{2\tau_r - T_s}{2\tau_r + T_s} \end{cases} \quad (8)$$

The estimation of the rotor flux vector in the $\alpha\beta$-stationary frame ($\Psi_{r\alpha\beta}$) is carried out using Equation (6), then applying the $dq$–to–$\alpha\beta$ transform, as:

$$\hat{\Psi}_r = \begin{bmatrix} \cos \theta_m & -\sin \theta_m \\
 \sin \theta_m & \cos \theta_m \end{bmatrix} \hat{\Psi}_{rdq} \quad (9)$$

where $\theta_m$ is the rotor electrical angular position.

The estimation of the stator flux vector $\hat{\Psi}_s$ at the $k$th sampling period uses Equation (9), as:

$$\hat{\Psi}_s^k = k_r \hat{\Psi}_r + \sigma l_s I_s^k \quad (10)$$

4.1.2. Stator Flux, Stator Current, and Electromagnetic Torque Prediction

The prediction of the stator flux vector $\Psi_s^{k+1}$ at the $(k + 1)$th sampling period is based on the forward-Euler approximation of the stator voltage vector $V_s$, using the IM model given in [37], with:

$$\Psi_s^{k+1} = \Psi_s^k + T_s V_s^k - r_s T_s I_s^k \quad (11)$$

According to Equation (5), the prediction of the electromagnetic torque $T_{em}^{k+1}$ requires the prediction of the stator current vector $I_s^{k+1}$ at the $(k + 1)$th, which is achieved by applying the forward-Euler approximation to Equation (1), as:

$$I_s^{k+1} = k_3 I_s^k + \frac{T_s}{L_s} V_s^k + k_4 \Psi_r^k \quad (12)$$

where:

$$\begin{cases} k_3 = 1 - r_s T_s / J \\
 k_4 = \frac{k_r T_s}{J} (1 - \omega_r) \end{cases} \quad (13)$$
Thus, the prediction of the electromagnetic torque $T_{em}^{k+1}$ at the $(k+1)$th sampling period is achieved using Equations (5), (11), and (12), as:

$$T_{em}^{k+1} = \frac{3}{2} N_p \Omega m \{ \Psi_s^{k+1} \times I_s^{k+1} \}$$  \hspace{1cm} (14)

Real-time implementation constraints make it necessary the compensation for the delay during the prediction step. To do so, the variables at the $(k+1)$th sampling period are used as initial conditions for the prediction at the $(k+2)$th sampling period, such that:

$${\begin{array}{*{20}c}
\Psi_s^{k+2} \\
I_s^{k+2} \\
T_{em}^{k+2} \\
\end{array}} = {\begin{array}{*{20}c}
\Psi_s^{k+1} + T_s V_s^{k+1} - r_s I_s^{k+1} \\
k_3 I_s^{k+1} + \frac{T_s}{L_e} V_s^{k+1} + k_4 \Psi_r^{k+1} \\
\frac{3}{2} N_p \Omega m \{ \Psi_s^{k+2} \times I_s^{k+2} \} \\
\end{array}}$$  \hspace{1cm} (15)

4.1.3. Cost Function Optimization

In the case of the two-level inverter-fed IM drives, the cost function $g$ of the C-PTC strategy is basically built around two main errors: (i) the difference between the desired and the predicted electromagnetic torque, and (ii) the difference between the desired and the predicted stator flux. The latter is multiplied by a weighting factor $\lambda$. While in the case of the 3L-NPC VSI, two additional criteria have to be incorporated into the cost function in order to achieve better performance, such that:

- Balancing the neutral point voltage. The corresponding error and weighting factor are noted as $\Delta V_{c12}$ and $\lambda_{cv}$, respectively,
- Minimizing the number of switching transitions $n_{sw}$. The weighting factor is noted as $\lambda_s$.

Neutral point voltage balancing: the neutral point “o” voltage has to be as close as possible to zero. In other words, the voltages of the DC bus capacitors $C_1$ and $C_2$ have to be balanced, with the error $\Delta V_{c12}$ tending to zero, in an attempt to achieve a ripple-free flux and torque.

Referring to [24], $\Delta V_{c12}$ could be defined in the $(k+1)$th sampling period, as:

$$\Delta V_{c12}^{k+1} = V_{c1}^{k+1} - V_{c2}^{k+1} = \Delta V_{c12}^k - \frac{T_s}{C} \Delta I_{c12}^k$$  \hspace{1cm} (16)

where:

$${\begin{array}{*{20}c}
\Delta V_{c12}^k \\
\Delta I_{c12}^{k+1} \\
\end{array}} = {\begin{array}{*{20}c}
V_{c1}^k - V_{c2}^k \\
I_{c1}^{k+1} - I_{c2}^{k+1} \\
\end{array}}$$  \hspace{1cm} (17)

The DC bus capacitor voltages are expressed at the $(k+1)$th sampling period, as [24]:

$${\begin{array}{*{20}c}
V_{c1}^{k+1} - V_{c1}^k \\
V_{c2}^{k+1} - V_{c2}^k \\
\end{array}} = {\begin{array}{*{20}c}
\frac{T_s}{C} I_{c1}^k \\
\frac{T_s}{C} I_{c2}^k \\
\end{array}}$$  \hspace{1cm} (18)

$I_{c1}^k$, $I_{c2}^k$ are the currents flowing across the two capacitors and are calculated using $K_{pi}$, $p = a, b, c$ refers to the inverter leg phase, and $i = 1, 2$ refers to the leg upper switches state, as:

$${\begin{array}{*{20}c}
i_{c1}^k \\
i_{c2}^k \\
\end{array}} = {\begin{array}{*{20}c}
i_{dc} - K_{a1} I_{a1}^k - K_{b1} I_{b1}^k - K_{c1} I_{c1}^k \\
i_{dc} + (1 - K_{a2}) I_{a2}^k + (1 - K_{b2}) I_{b2}^k + (1 - K_{c2}) I_{c2}^k \\
\end{array}}$$  \hspace{1cm} (19)

where $i_{dc}$ is the DC link current at the input of the 3L-NPC inverter.
Minimization of the number of switching transitions: Accounting for the large number of voltage vectors generated by the 3L-NPC inverter and in order to reduce its switching frequency, the incorporation in the cost function of a new term function of the number of the switching transitions $n_{sw}$ is mandatory, as in [24]:

$$n_{sw}^{k+1} = \sum \left| (K_{p1}^{k+1})_i - K_{p1}^k \right| + \left| (K_{p2}^{k+1})_i - K_{p2}^k \right|$$ (20)

Resulting cost function: The cost function of the C-PTC is defined as [24]:

$$g = \frac{(T_{em}^* - T_{em}^{k+2})^2}{T_{emR}^2} + \lambda_f \left( \frac{\|\Psi^* - \|\Psi_s^{k+2}\|}{\|\Psi_{sR}\|^2} \right)^2 + \lambda_{cv} |\Delta V_{c12}^{k+1}| + \lambda_{sw} n_{sw}$$ (21)

where:

- $T_{em}^*$ is the reference electromagnetic torque, which corresponds to the output of speed PI controller.
- $T_{emR}$ is the rated electromagnetic torque.
- $\|\Psi_{sR}\|$ is the rated stator flux.
- $\lambda_f$ is the weighting flux factor, introduced to account for the difference in units and magnitudes of the torque and flux.

In order to gain a high performance over a wide speed range, the three weighting factors $\lambda_f$, $\lambda_{cv}$, and $\lambda_{sw}$ have to be tuned according to the operating point. This could be achieved using several methods based on analytical or empirical approaches [40].

To summarize, a flowchart of the C-PTC is illustrated in Figure 4.

![Flowchart of the C-PTC strategy implemented in a 3L-NPC inverter-fed IM drive.](image-url)
4.2. Introduced PTCs

The above-described C-PTC suffers from the following three major drawbacks:
1. The great effort required to tune three weighting factors ($\lambda_f$, $\lambda_{cv}$, and $\lambda_s$).
2. The huge CPU time spent in the online selection among 27 of the suitable voltage vectors.
3. The excessive torque and stator flux ripple.

The two introduced PTC strategies, to be developed hereunder, have the merit to eradicate the drawbacks of the C-PTC.

4.2.1. SV-PTC1

The SV-PTC1 strategy is based on the bus-clamping approach, which has been applied to space vector pulse width modulation (SV-PWM) techniques [41] and uncontrolled switching frequency DTC strategies [42]. It consists of connecting each IM stator phase to the high or low levels of the DC bus voltage.

As illustrated in Figure 2, the $\alpha\beta$-plane is divided into six equal sectors, defined as follows:

\[
\text{Sector } i : \quad (2i - 3) \frac{\pi}{6} \leq \theta_{\Psi_s} < (2i - 1) \frac{\pi}{6} \quad (22)
\]

where $i$ is an integer, with: $1 \leq i \leq 6$.

The position of the stator flux vector $\Psi_s$ in the $\alpha\beta$-plane is:

\[
\theta_{\Psi_s} = \arctan\left(\frac{\psi_\beta}{\psi_\alpha}\right) \quad (23)
\]

The conventional SV-PWM of the three-phase 3L-NPC converter is based on the approximation of the reference vector $V^*$ by applying the three closest voltage vectors $V_i$ [43]. For instance, in the case when $\Psi_s$ is located in sector 1 and for an anti-clockwise rotation, the corresponding reference voltage vector $V^*$ is 90°-shifted with respect to the reference flux vector $\Psi_s$, according to the expression of the stator voltage vector and by neglecting the voltage drop $r_s I_s$, as illustrated in Figure 5.

![Figure 5. Position of the reference voltage vector $V^*$ in the case where the stator flux vector $\Psi_s$ is located in sector 1.](image-url)

In this case, the approximation of the reference voltage vector $V^*$ is achieved by applying dedicated voltage vectors, according to the location of the extremity of $V^*$, as follows:
with the minimum number of voltage vectors, as the main objective. The control of the neutral point voltage is no longer incorporated in the cost function equation through \( \Delta V_{cl2} \).

The second introduced strategy SV-PTC2 considers balancing the neutral point voltage with the minimum number of voltage vectors, as the main objective. The control of the neutral point voltage is no longer incorporated in the cost function equation through \( \Delta V_{cl2} \).

### 4.2.2. SV-PTC2

The second introduced strategy SV-PTC2 considers balancing the neutral point voltage with the minimum number of voltage vectors, as the main objective. The control of the neutral point voltage is no longer incorporated in the cost function equation through \( \Delta V_{cl2} \).

**Table 2. Selected vectors and the corresponding clamped stator phases considered in the SV-PTC1 strategy in the case of an anti-clockwise rotation of \( \Psi_s \).**

| Sector | Selected Vectors | Clamped Phase |
|--------|------------------|---------------|
| 1      | \( V_8, V_9, V_{10}, V_{14}, V_{16}, V_{17}, V_{22} \) | \( c^- \) |
| 2      | \( V_2, V_3, V_4, V_7, V_{17}, V_{18}, V_{23} \) | \( b^+ \) |
| 3      | \( V_{10}, V_{11}, V_{12}, V_{14}, V_{15}, V_{19}, V_{24} \) | \( a^- \) |
| 4      | \( V_4, V_5, V_6, V_7, V_{19}, V_{20}, V_{25} \) | \( c^+ \) |
| 5      | \( V_8, V_{12}, V_{13}, V_{14}, V_{15}, V_{20}, V_{26} \) | \( b^- \) |
| 6      | \( V_1, V_2, V_6, V_7, V_{15}, V_{16}, V_{21} \) | \( a^+ \) |

Referring to Table 2, one can notice that only 7 voltage vectors are evaluated in the cost function per sector, instead of 27 in the case of the C-PTC. This limited choice of vectors per sector offers the mitigation of the number of commutations by clamping the stator phase to the high and low levels of the DC bus voltage. Such clamping makes simplifying the cost function and the corresponding clamped phases, in the case of an anti-clockwise rotation of \( \Psi_s \).
and its weighting factor $\lambda_{cv}$. This has been made possible considering the shift between the two DC bus capacitor voltages, at the beginning of each sector, as:

- $V_{dc1} > V_{dc2}$: the upper switches are switched ON to achieve the balance;
- $V_{dc1} < V_{dc2}$: the lower switches are switched ON to achieve the balance.

Thus, for each sector, the two sets of selected voltage vectors detailed in the case of SV-PTC1 are predefined, to maintain the DC-link at high or low levels, depending on the sign of $(V_{dc1} - V_{dc2})$. Thus, unlike the SV-PTC1 strategy, the clamping of each phase is no longer regular. For instance, in the cases of sectors 1 and 2, if the DC bus voltage error $(V_{dc1} - V_{dc2})$ is positive, the b-phase would be clamped to the high level of the DC bus voltage during 120°.

Table 3 summarizes the sets of the selected vectors and the corresponding clamped stator phases depending on the position of the $\Psi_s$ and the comparison between $V_{dc1}$ and $V_{dc2}$.

### Table 3. Selected vectors and the corresponding clamped stator phases considered in the SV-PTC2 strategy in the case of an anti-clockwise rotation of $\Psi_s$

| Sector | Condition | Selected Vectors | Clamped Phase |
|--------|-----------|------------------|---------------|
| 1      | $V_{dc1} > V_{dc2}$ | $V_2, V_3, V_7, V_{16}, V_{17}, V_{22}, V_4$ | $b^+$ |
|        | $V_{dc1} < V_{dc2}$ | $V_8, V_9, V_{10}, V_{14}, V_{16}, V_{17}, V_{22}$ | $c^-$ |
| 2      | $V_{dc1} > V_{dc2}$ | $V_2, V_3, V_4, V_7, V_{17}, V_{18}, V_{23}$ | $b^+$ |
|        | $V_{dc1} < V_{dc2}$ | $V_{10}, V_{11}, V_{14}, V_{17}, V_{18}, V_{23}, V_{12}$ | $a^-$ |
| 3      | $V_{dc1} > V_{dc2}$ | $V_4, V_5, V_7, V_{18}, V_{19}, V_{24}, V_6$ | $c^+$ |
|        | $V_{dc1} < V_{dc2}$ | $V_{10}, V_{11}, V_{12}, V_{14}, V_{18}, V_{19}, V_{24}$ | $a^-$ |
| 4      | $V_{dc1} > V_{dc2}$ | $V_4, V_5, V_6, V_7, V_{19}, V_{20}, V_{25}$ | $c^+$ |
|        | $V_{dc1} < V_{dc2}$ | $V_{12}, V_{13}, V_{14}, V_{19}, V_{20}, V_{25}, V_8$ | $b^-$ |
| 5      | $V_{dc1} > V_{dc2}$ | $V_4, V_6, V_7, V_{15}, V_{20}, V_{20}, V_2$ | $a^+$ |
|        | $V_{dc1} < V_{dc2}$ | $V_8, V_{12}, V_{13}, V_{14}, V_{15}, V_{20}, V_{26}$ | $b^-$ |
| 6      | $V_{dc1} > V_{dc2}$ | $V_4, V_2, V_6, V_7, V_{15}, V_{16}, V_{21}$ | $a^+$ |
|        | $V_{dc1} < V_{dc2}$ | $V_8, V_9, V_{14}, V_{15}, V_{16}, V_{21}, V_{10}$ | $c^-$ |

In the manner of the SV-PTC1 strategy, just seven vectors were evaluated in the cost function, which was reduced to:

$$
S = \left( \frac{T_{em}^s - T_{em}^{k+2}}{T_{emR}^2} \right)^2 + \lambda f \left( \|\Psi_s^k\| - \|\Psi_s^{k+2}\| \right)^2
$$

### 5. Case Study

The performances of both introduced PTC strategies were investigated through a set of simulation works, including steady-state and transient behaviors of a 3L-NPC inverter-fed IM drive, and compared to the C-PTC ones. Then experiments were carried out on a developed test bench in order to validate the simulation results.

The IM under study had a rated power of 1.1 kW, a rated voltage of 220 V, a rated current of 2.5 A, and a rated speed of 2820 rpm at a stator frequency of 50 Hz. The parameters are listed in Table 4.

### Table 4. IM parameters

| $r_s$ | $l_s$ | $M$ | $J$ | $f$ |
|-------|-------|-----|-----|-----|
| 6.32Ω | 0.692H | 0.666H | $3.5e^{-3}$ Kg·m² | $9e^{-3}$ N·m·s |
| 7.36Ω | 0.692H | $N_p$ | 1 | }
The different weighting factors for the three PTC strategies are provided in Table 5.

### Table 5. Weighting factor values of the three PTC strategies.

| Weighting Factor | C-PTC | SV-PTC1 | SV-PTC2 |
|------------------|-------|---------|---------|
| $\lambda_f$      | 100   | 100     | 100     |
| $\lambda_{cv}$   | 1     | 1       | 0       |
| $\lambda_s$      | $10^{-6}$ | 0       | 0       |

The DC bus voltage $V_{dc}$ is kept constant, equal to 400 V. The amplitude $\|\Psi_s\|$ of the reference stator flux is equal to 0.947 Wb, and the sampling period $T_s$ is equal to 100 $\mu$s.

### 5.1. Simulation Results

The first part of the simulation works dealt with the transient behavior of the 3L-NPV inverter-fed IM under the three PTC strategies, during a start-up until reaching 190 rpm, followed by an acceleration to 1700 rpm, then a deceleration to 1150 rpm, under a load torque $T_l = 3.56$ N·m. The investigated features were the rotor speed and its reference, the neutral point voltage, the stator phase current, and the electromagnetic and load torques. The obtained results are illustrated in Figure 6.

From the analysis of the results of Figure 6, one can notice that:

- The rotor speed (in blue) followed its reference (in red) under the three PTC schemes.
- The neutral point voltage was well-balanced under the three schemes, with a slight increase limited to 1.25% of the DC bus voltage at low speeds under the SV-PTC2.
- A reduction of the stator current harmonic distortion and damping of the electromagnetic torque ripple were gained thanks to the implementation of SV-PCT1 and SV-PCT2.

![Figure 6](image-url)

**Figure 6.** Transient behavior of the 3L-NPV inverter-fed IM drive during a start-up to reach 190 rpm, followed by an acceleration to 1700 rpm, then a deceleration to 1150 rpm, for $T_l = 3.56$ N·m.

**Legend 1** (a): C-PTC, (b): SV-PTC1, (c): SV-PTC2. **Legend 2** top to bottom: rotor speed and its reference, neutral point voltage, stator phase current, and electromagnetic and load torques.
The second part of the simulation was devoted to the investigation of selected features under a steady-state operation of the IM fed by the 3L-NPC inverter controlled by the three PTCs under comparison. Figure 7 shows the simulation results for a steady-state operating point characterized by a speed of 286 rpm and a load torque of 3.56 N.m. The analysis of these results reveals that:

- The DC bus capacitance voltages were quite balanced under the three strategies with the neutral point voltage not exceeding 0.2 V for C-PTC, 2 V for the SV-PTC1, and 2.2 V for SV-PTC2.
- The two proposed SV-PTCs mitigated the switching commutation thanks to the clamping of the stator phases to the high and low levels of DC bus voltages, which were regular in the case of SV-PTC1 and arbitrary in the case of SV-PTC2.
- Compared to C-PTC, SV-PTC1 and SV-PTC2 strategies exhibited higher performances in terms of the reduction of the ripple of the stator flux and electromagnetic torque. This statement was confirmed by a firm comparison; the results are provided in Table 6.

Figure 7. Steady-state operation of the 3L-NPV inverter-fed IM drive at a speed of 286 rpm and a load torque of 3.56 N.m. Legend 1 (a): C-PTC, (b): SV-PTC1, (c): SV-PTC2. Legend 2 top to bottom: sectors succession, the sum $K_{a1} + K_{a2}$, DC bus capacitance voltages, stator phase current, stator flux and its reference, and electromagnetic and load torques.
Table 6. Comparison study with emphasis on the stator flux and electromagnetic torque ripple at a steady-state operating point characterized by a speed of 286 rpm and a load torque of 3.56 N·m.

|            | Δ||Ψ_s|| (Wb) | ΔT_{em} (N·m) |
|------------|--------------|---------------|
| C-PTC      | 0.0987       | 3.58          |
| SV-PTC1    | 0.0711       | 2.62          |
| SV-PTC2    | 0.06         | 2.53          |

5.2. Experimental Validation

5.2.1. Test Bench Description

For the sake of the experimental validation of the simulation results, a test bench was built around a 3L-NPC inverter-fed IM drive, as shown in Figure 8. Manufactured by SEMIKRON, the 3L-NPC inverter was equipped with three current sensors at its output, and two voltage sensors enabled the measurement of the voltages across the DC bus capacitors in its input. The latter was made up of the parallel connection of six equal capacitors of 610 µF, yielding a total of 3660 µF per DC bus capacitor. A SEMIKRON full bridge diode rectifier fed the 3L-NPC inverter with a constant DC voltage of 400 V. The IM had the same ratings and parameters as the machine considered in the simulation. It was mechanically-coupled to a brushless DC generator whose armature was connected to a variable three-phase resistor. The mechanical sensor mounted on the IM second shaft terminal was a rotary incremental optical encoder HEIDENHAIN ROD 420D yielding 1024 pulses per revolution. The used digital control platform was a dSPACE-1104 linked to a desk computer through a PCI connector. The developed PTC strategies were implemented using a Matlab–Simulink software package.

![Figure 8. Developed test bench built around a 3L-NPV inverter-fed IM drive.](image)

5.2.2. Transient Behavior

Figure 9 characterizes the transient behavior of the 3L-NPV inverter-fed IM drive under a ramp-shaped reference speed, from 300 to 1800 rpm, and back to 300 rpm:

- Figure 9a–c show the scopes illustrating the waveforms of the motor speed and the electromagnetic torque yielded by C-PTC, SV-PTC1, and SV-PTC2, respectively;
- Figure 9d–f show the scopes illustrating the waveforms of the stator flux and the electromagnetic torque yielded by C-PTC, SV-PTC1, and SV-PTC2, respectively.

Referring to the experimental results shown in Figure 9, one can remark that the speed follows its reference under the three PTC schemes during acceleration and deceleration. Moreover, a significant reduction of the electromagnetic torque ripple is gained following
the implementation of the two proposed PTCs, in comparison with the one measured following the implementation of the C-PTC. Finally, the stator flux had almost a ripple-free waveform under the three PTCs.

![Figure 9](image.png)

**Figure 9.** Experimental characterization of the transient behavior of the 3L-NPV inverter-fed IM drive under a ramp-shaped reference speed, from 300 to 1800 rpm, and back to 300 rpm, and under a constant load torque of 3.56 N·m. **Legend 1:** (C-PTC) left scopes, (SV-CPT1) middle scopes, (SV-CPT2) right scopes. **Legend 2:** (a–c) the rotor speed in the top with \( \Omega_r \ (500 \text{ tr/min/\text{div}}) \) and the electromagnetic torque \( T_{em} \) in the bottom with (2 N·m/\text{div}) and the electromagnetic torque \( T_{em} \) in the bottom with (2 N·m/\text{div}).

### 5.2.3. Steady-State Operation

Figure 10 characterizes the steady-state operation of the 3L-NPV inverter-fed IM drive at a speed of 190 rpm and a load torque of 3.56 N·m:

- Figure 10a–c show the scopes illustrating the sector succession and the sum of the control signals of \( K_{a1} + K_{a2} \) yielded by C-PTC, SV-PTC1, and SV-PTC2, respectively,
- Figure 10d–f show the scopes illustrating the waveforms of the two DC bus capacitor voltages \( V_{dc1} \) and \( V_{dc2} \) yielded by C-PTC, SV-PTC1, and SV-PTC2, respectively.
- Figure 10g–i show the scopes illustrating the locus described by the extremities of the applied voltage vectors \( V_i \) yielded by C-PTC, SV-PTC1, and SV-PTC2, respectively.
- Figure 10j–l show the scopes illustrating the waveforms of the stator phase voltage and current yielded by C-PTC, SV-PTC1, and SV-PTC2, respectively.

In the manner of the simulation results, the experimental ones confirm the superiority of the proposed PTCs compared to C-PTC. Such superiority has been gained thanks to the minimization of the number of selected voltage vectors without compromising the neutral point voltage balance.

The experimental characterization of the steady-state operation of the 3L-NPV inverter-fed IM drive, at a speed of 190 rpm and a load torque of 3.56 N·m, was extended to further features. The obtained results are shown in Figure 11. These clearly validate the significant reduction of the torque ripple gained thanks to the implementation of the introduced PTC strategies, compared to those yielded by C-PTC.

Finally, the steady-state operation of the 3L-NPV inverter-fed IM drive was experimentally-investigated in a second point characterized by a speed of 1720 rpm and a load torque of 3.56 N·mf. The obtained results are illustrated in Figure 12.
Figure 10. Experimental characterization of the steady-state operation of the 3L-NPV inverter-fed IM drive at a speed of 190 rpm and a load torque of 3.56 N·m. **Legend 1:** (C-PTC) left scopes, (SV-PTC1) middle scopes, (SV-PTC2) right scopes. **Legend 2:** (a–c) the sector succession at the top with (2 sectors/div) and the sum of the control signals $K_{a1} + K_{a2}$ in the bottom, (d–f): the two DC bus capacitor voltages $V_{dc1}$ and $V_{dc2}$ (100 V/div), (g–i): the locus described by the extremities of the applied voltage vectors $V_s$ in the $a\beta$-plane (100 V/div), (j–l): the stator phase voltage in the top (200 V/div) and the stator phase current in the bottom (2 A/div).
5.3. Simulation versus Experimental Results

In order to characterize the degree of agreement between the simulation and experimental results, a simulation was carried out at a steady-state operation under the same experimental conditions of Figures 10 and 11, with emphasis on the stator current and the electromagnetic torque waveforms. In order to amplify the waveforms making clear the differences between the simulation and experimental results, the investigation has been limited to a single strategy: SV-PTC1 for instance. The obtained results are shown in Figure 13. For the sake of comparison, the experimental waveform of the stator current shown in Figure 10 and one of the electromagnetic torques depicted in Figure 11, are recalled. One can notice that both simulated and measured stator phase currents are slightly distorted sharing almost the same peak-to-peak value of 6 A. Concerning the electromagnetic torque, the simulated waveform has a ripple $\Delta T_{em}$ of 2.57 N·m, while the one yielded by the experimental tests was 2.60 N·m. This comparison proved the acceptable agreement between simulation and experimental results.
Figure 12. Experimental characterization of the steady-state operation of the 3L-NPV inverter-fed IM drive for a speed of 1720 rpm and a load torque of 3.56 N·m. **Legend 1:** (C-PTC) left scopes, (SV-PTC1) middle scopes, (SV-PTC2) right scopes. **Legend 2:** (a–c) the stator phase voltage in the top (200 V/div) and the stator phase current in the bottom (2 A/div), (d–f): the stator flux $\Psi_s$ in the top (0.5 Wb/div), and the electromagnetic torque $T_{em}$ in the bottom (2 N·m/div).

Figure 13. Steady-state operation of the 3L-NPV inverter-fed IM drive under the control of the SV-PTC1 for a speed of 190 rpm and a load torque of 3.56 N·m. **Legend 1:** (top) simulation results (bottom) experimental results, **Legend 2:** (left) stator phase current, (right) electromagnetic torque

6. Conclusions
An investigation of the performances of two novel PTC strategies intended to the control of 3L-NPC inverter-fed IM drives and their comparisons with those exhibited by the conventional PTC ones were treated in this paper. The introduced PTC strategies have the merit to minimize the number of selected voltage vectors to 7, resulting in a significant reduction of the complexity and computational burden, while 27 voltage vectors are involved in the conventional PTC.
Simulation works, validated by experiments carried out on a developed test bench, enabled the investigation and the comparison of the performances of the three PTCs during the transient and steady-state operations of a 3L-NPC inverter-fed IM drive.

It has been found that, following their comparisons with the conventional PTC strategy, the two proposed ones offer the following benefits:

- The reduction of the switching transitions (thanks to the phases clamping);
- The mitigation of the current distortion and the torque and flux ripples,
- The minimization of the number of weighting factors in the cost function without disturbing the neutral point voltage balance.

The performances exhibited by the proposed PTC strategies have been reached thanks to the accurate identification of the IM parameters. However, it is well known that these parameters vary with the operating conditions (heating, saturation, skin effects, vibration).

It is clear that an online identification of the machine parameters is key to improving the robustness of the PTC strategies. This represents one of the major outlooks of the present work.

**Author Contributions:** Conceptualization, W.Z. and B.E.B.; Formal analysis, B.E.B.; Funding acquisition, W.Z. and B.E.B.; Investigation, W.Z. and B.E.B.; Methodology, B.E.B.; Project administration, A.M.; Resources, A.M.; Software, W.Z., I.N.E.B. and B.E.B.; Supervision, B.E.B.; Writing—review & editing, A.M. All authors have read and agreed to the published version of the manuscript.

**Funding:** This research received no external funding.

**Institutional Review Board Statement:** Not applicable.

**Informed Consent Statement:** Not applicable.

**Data Availability Statement:** Not applicable.

**Conflicts of Interest:** The authors declare no conflict of interest.

**References**

1. Takahashi, I.; Noguchi, T. A new quick-response and high-efficiency control strategy of an induction motor. *IEEE Trans. Ind. Appl.* 1986, 22, 820–827. [CrossRef]

2. Depenbrock, M. Direct self-control (DSC) of inverter-fed induction machine. *IEEE Trans. Power Electron.* 1988, 3, 420–429. [CrossRef]

3. Casadei, D.; Profumo, F.; Serra, G.; Tani, A. FOC and DTC: Two viable schemes for induction motors torque control. *IEEE Trans. Power Electron.* 2002, 17, 779–787. [CrossRef]

4. Xie, H.; Wang, F.; He, Y.; Rodriguez, J.; Kennel, R. Encoderless parallel predictive torque control for induction machine using a robust model reference adaptive system. *IEEE Trans. Energy Convers.* 2022, 37, 232–242. [CrossRef]

5. Alsofiyani, I.M.; Lee, K.-B. A unidirectional voltage vector preselection strategy for optimizing model predictive torque control with discrete space vector modulation of IPMSM. *IEEE Trans. Indus. Electron.* 2022, 69, 12305–12315. [CrossRef]

6. Fang, G.; Ye, J.; Xiao, D.; Xia, Z.; Emadi, A. Low-ripple continuous control set model predictive torque control for switched reluctance machines based on equivalent linear SRM model. *IEEE Trans. Indus. Electron.* 2022, 69, 12480–12495. [CrossRef]

7. He, Z.; Shi, Q.; Wei, Y.; Zheng, J.; Gao, B.; He, L. A torque demand model predictive control approach for driving energy optimization of battery electric vehicle. *IEEE Trans. Veh. Technol.* 2021, 70, 3232–3242. [CrossRef]

8. Kousalya, V.; Rai, R.; Singh, B. Sliding model-based predictive torque control of induction motor for electric vehicle. *IEEE Trans. Ind. Appl.* 2022, 58, 742–752.

9. Badawy, M.O.; Sharma, M.; Hernandez, C.; Elrayyah, A.; Guerra, S.; Coe, J. Model predictive control for multi-port modular multilevel converters in electric vehicles enabling HEDSs. *IEEE Trans. Energy Convers.* 2022, 37, 10–23. [CrossRef]

10. Liu, H.; Zhang, L.; Wang, P.; Chen, H. A real-time NMPC Strategy for electric vehicle stability improvement combining torque vectoring with rear-wheel steering. *IEEE Trans. Trans. Electrif.* 2022, 8, 3825–3835. [CrossRef]

11. Oncken, J.; Chen, B. Real-time model predictive powertrain control for a connected plug-in hybrid electric vehicle. *IEEE Trans. Veh. Technol.* 2022, 69, 8420–8432. [CrossRef]

12. Pereira, D.F.; da Costa Lopes, F.; Watanabe, E.H. Nonlinear model predictive control for the energy management of fuel cell hybrid electric vehicles in real time. *IEEE Trans. Ind. Electron.* 2021, 68, 3213–3223. [CrossRef]

13. Griefnow, P.; Jakoby, M.; Dorschel, L.; Andert, J. Nonlinear model predictive control of mild hybrid powertrains with electric supercharging. *IEEE Trans. Veh. Technol.* 2021, 70, 8490–8504. [CrossRef]
14. Zhang, Y.; Huang, Y.; Chen, Z.; Li, G.; Liu, Y. An optimal control strategy for plug-in hybrid electric vehicles based on enhanced model predictive control with efficient numerical method. *IEEE Trans. Trans. Electrific.* 2022, 8, 2516–2530. [CrossRef]

15. Kou, P.; Wang, J.; Liang, D. Powered yaw control for distributed electric propulsion aircraft: A model predictive control approach. *IEEE Trans. Trans. Electrific.* 2022, 7, 3006–3020. [CrossRef]

16. Zhang, J.; Roumeliotis, I.; Zolotas, A. Nonlinear model predictive control-based optimal energy management for hybrid electric aircraft considering aerodynamics–propulsion coupling effects. *IEEE Trans. Trans. Electrific.* 2022, 8, 2640–2653. [CrossRef]

17. Gomez, I.A.G.; Lourenço, L.E.N.; Grilo, A.P.; Salles, M.B.C.; Meegahapola, L.; Sguarezi Filho, A.J. Primary frequency response of microgrid using doubly fed induction generator with finite control set model predictive control plus droop control and storage system. *IEEE Access* 2020, 8, 189298–189312. [CrossRef]

18. Jlassi, I.; Marques Cardoso, A.J. Enhanced and computationally efficient model predictive flux and power control of PMSG drives for wind turbine applications. *IEEE Trans. Ind. Electron.* 2021, 68, 6574–6583. [CrossRef]

19. Zarei, M.E.; Ramirez, D.; Prodanoic, M.; Arana, G.M. Model predictive control for PMSG-based wind turbines with overmodulation and adjustable dynamic response time. *IEEE Trans. Ind. Electron.* 2022, 69, 1573–1585. [CrossRef]

20. Lin, X.; Huang, W.; Jiang, W.; Zhao, Y.; Zhu, S. Predictive torque control for PMSM based on weighting factor elimination and fast voltage vector selection. *IEEE J. Emerg. Sel. Top. Power Electron.* 2020, 8, 3736–3750. [CrossRef]

21. Zhang, Y.; Bai, Y.; Yang, H.; Zhang, B. Low switching frequency model predictive control of three-level inverter-fed IM drives with speed-sensorless and field-weakening operations. *IEEE Trans. Ind. Electron.* 2019, 66, 4262–4272. [CrossRef]

22. Wang, F.; Li, Z.; Tong, X. Modified predictive control method of three-level simplified neutral point clamped inverter for common-mode voltage reduction and neutral-point voltage balance. *IEEE Access* 2019, 7, 119476–119485. [CrossRef]

23. Jin, N.; Dai, D.; Xie, H.; Wu, J.; Guo, L. Virtual vector-based FCS-MPC for NPC three-level grid-tied inverter without weighting factor of neutral-point voltage balancing. *IEEE Access* 2020, 12, 72806–72814. [CrossRef]

24. Habibullah, M.; Lu, D.D.-C.; Xiao, D.; Rahman, M.F. Finite-state predictive torque control of induction motor supplied from a three level NPC voltage source inverter. *IEEE Trans. Power. Electron.* 2017, 32, 479–489. [CrossRef]

25. Habibullah, M.; Lu, D.D.C.; Xiao, D.; Osman, I.; Rahman, M.F. Selected prediction vectors based FS-PTC for 3L-NPC inverter fed motor drives. *IEEE Trans. Ind. Appl.* 2017, 53, 3588–3597. [CrossRef]

26. Habibullah, M.; Lu, D.D.-C.; Xiao, D.; Fletcher, J.E.; Rahman, M.F. Predictive torque control of induction motor sensorless drive fed by a 3L-NPC inverter. *IEEE Trans. Ind. Inform.* 2017, 13, 60–70. [CrossRef]

27. Zhang, Z.; Tian, W.; Xiong, W.; Kennel, R. Predictive torque control of induction machines fed by 3L-NPC converters with online weighting factor adjustment using fuzzy logic. In Proceedings of the 2017 IEEE Transportation Electrification Conference and Expo (ITEC), Chicago, IL, USA, 22–24 June 2017.

28. Osman, I.; Xiao, D.; Rahman, M.F. Two-stage optimization based finite state predictive torque control for 3L-NPC inverter fed IM drives. *IET Electr. Power Appl.* 2018, 13, 64–72. [CrossRef]

29. Bandy, K.; Stumpf, P. Model predictive torque control for multilevel inverter fed induction machines using sorting networks. *IEEE Access* 2021, 9, 13800–13813. [CrossRef]

30. Xiao, D.; Alam, K.S.; Osman, I.; Akter, M.P.; Shakib, S.M.S.I.; Rahman, M.F. Low complexity model predictive flux control for three-level neutral-point clamped inverter-fed induction motor drives without weighting factor. *IEEE Trans. Ind. Appl.* 2020, 56, 6496–6506. [CrossRef]

31. Xiao, D.; Akter, M.P.; Alam, K.; Dutta, R.; Mekhilef, S.; Rahman, M.F. Cascaded predictive flux control for a 3-L active NPC fed IM drives without weighting factor. *IEEE Trans. Energy Convers.* 2021, 36, 1797–1807. [CrossRef]

32. Osman, I.; Xiao, D.; Rahman, M.F.; Norambuena, M.; Rodriguez, J. An optimal reduced-control-set model predictive flux control for 3L-NPC fed induction motor drive. *IEEE Trans. Energy Convers.* 2021, 36, 2967–2976. [CrossRef]

33. Jun, E.-S.; Nguyen, M.H.; Kwak, S.-S. Model predictive control method with NP voltage balance by offset voltage injection for three-phase three-level NPC inverter. *IEEE Access* 2021, 8, 172175–172195. [CrossRef]

34. Wang, F.; Lin, G.; He, Y. Passivity-based model predictive control of three-level inverter-fed induction motor. *IEEE Trans. Power. Electron.* 2021, 36, 1984–1993. [CrossRef]

35. Begh, M.A.W.; Karamanakos, P.; Geyer, T. Gradient-based predictive pulse pattern control of medium-voltage drives—Part I: Control, concept, and analysis. *IEEE Trans. Power Electron.* 2022, 37, 14222–14236. [CrossRef]

36. Begh, M.A.W.; Karamanakos, P.; Geyer, T. Gradient-based predictive pulse pattern control of medium-voltage drives—Part II: Performance assessment. *IEEE Trans. Power Electron.* 2022, 37, 14237–14251. [CrossRef]

37. Zouari, W.; El Badsi, B. Comparative investigation of bus-clamping PTC strategies for IM drives. In Proceedings of the 2017 Twelfth International Conference on Ecological Vehicles and Renewable Energies (EVER), Monte Carlo, Monaco, 11–13 April 2017.

38. Mohan, D.; Zhang, X.; Foo, G.H.B. Three-level inverter-fed direct torque control of IPMSM with constant switching frequency and torque ripple reduction. *IEEE Trans. Ind. Electron.* 2016, 63, 7908–7918. [CrossRef]

39. Zhang, Y.; Yang, H.; Xia, B. Model predictive torque control of induction motor drives with reduced torque ripple. *IET Electr. Power Appl.* 2015, 9, 595–604. [CrossRef]

40. Habibullah, M.; Lu, D.D.-C.; Xiao, D.; Rahman, M.F. A simplified finite-state predictive direct torque control for induction motor drive. *IEEE Trans. Ind. Electron.* 2016, 63, 3964–3975. [CrossRef]

41. Basu, K.; Prasad, J.S.S.; Narayanan, G.; Krishnamurthy, H.K.; Ayyanar, R. Reduction of torque ripple in induction motor drives using an advanced hybrid PWM technique. *IEEE Trans. Ind. Electron.* 2010, 57, 2085–2091. [CrossRef]
42. El Badsi, B.; Bouzidi, B.; Masmoudi, A. Bus-clamping-based DTC: An attempt to reduce harmonic distortion and switching losses. *IEEE Trans. Ind. Electron.* **2013**, *60*, 873–884. [CrossRef]

43. Li, K.; Wei, M.; Xie, C.; Deng, F.; Guerrero, J.M.; Vasquez, J.C. Triangle carrier-based DPWM for three-level NPC inverters. *IEEE J. Emerg. Sel. Top. Power Electron.* **2018**, *6*, 1966–1978. [CrossRef]