A Look-up Table Model Predictive Direct Torque Control of Permanent Magnet Synchronous Generator based on Vienna Rectifier

Sook Yee Yip, Hang Seng Che, Member, IEEE, Chee Pin Tan, Member, IEEE, Wen Tong Chong

Abstract—This paper presents the development and implementation of a Look-Up Table Model Predictive Direct Torque Control (LUT-MPDT) for a Permanent Magnet Synchronous Generator (PMSG) controlled using a Vienna rectifier. Using a machine model in the stator synchronous reference frame, the feasibility of direct torque control-based Model Predictive Control (MPC) for the Vienna rectifier is demonstrated. To reduce the computational burden of the MPDTC, a Look-Up Table (LUT) is proposed based on the stator current position and DC-link voltages. The LUT simplifies the complexity of the MPDTC by reducing the number of switching vectors from eight to five, whilst maintaining the stabilization of the neutral point voltage. The performance and effectiveness of the proposed method are evaluated via Matlab/Simulink and further validated experimentally using a Texas Instrument based TMS320F28335 development board. Simulation and experimental results show that the proposed LUT-MPDTC can successfully improve the torque and stator flux ripples as compared to the classic DTC.

Index Terms—Direct torque control, finite-set model predictive control, permanent magnet synchronous generator, Vienna rectifier

I. INTRODUCTION

Effective control of power electronics converters is an essential part of modern Wind Energy Conversion Systems (WECS). The choice of control is different subject to the type of generators and topology of the converter. Among the different machine types, the Permanent Magnet Synchronous Generator (PMSG) is commonly used in modern WECS since it offers the merits of high efficiency, high power density and is able to operate in direct drive or gearless modes [1]. For low-cost applications, the PMSG can be controlled using a diode-rectifier cascaded with a boost converter [2]. Albeit being cheap and simple, such a topology introduces significant current distortions [3] which translates into higher torque ripples and losses. For better current and torque performance as well as improved controllability, the full six-switch rectifier is generally used for controlling three-phase PMSG [4] by trading cost for performance.

Apart from aforementioned converters, the Vienna rectifier appears to be an interesting alternative: it has better controllability and performance as compared to a diode-rectifier and it is cheaper than a full six-switch rectifier. Since it was first introduced in [5], the Vienna rectifier has attracted the attention in academia as well as the industry, where it had been used in grid-tied rectifiers with power factor correction [6]. Subsequently, it was extended to WECS applications for generator control [7, 8].

For grid-converter applications, the well-known Voltage Oriented Control (VOC) method is commonly adopted whereby control is carried out in the rotating reference frame which is aligned to the grid voltage vector. For the PMSG, however, control is usually performed based on the Field Oriented Control (FOC) method, whereby the rotating frame is aligned with the rotor magnetic vector. In this case, the torque and flux of the PMSG can be controlled by regulating the \(q\) and \(d\)-axis currents respectively, using hysteresis current controllers [8]. For better control performance, Predictive Current Control (PCC) for PMSG with Vienna rectifier have been reported in [9, 10]. In these proposals, the torque and flux currents are controlled using the Model Predictive Control (MPC) approach. Despite being superior to the classical hysteresis control, MPC is known for its high computational burden. Consequently, various works have been dedicated to reduce the computation burden of MPC using the Vienna rectifier. In [10], the concept of neutral point voltage bound was utilized whereby only the candidate switching vectors that fall within the preset bound are used for the cost function calculation of MPC. On the other hand, the authors in [11] designed a hybrid control method that combined a DC-link voltage PI controller and MPC which is...
less computationally burdensome. It had been demonstrated for the Vienna rectifier in the grid-converter application.

Instead of controlling the currents as in FOC, the torque and flux of a generator may be regulated directly using the Direct Torque Control (DTC) approach. Numerous works on DTC for PMSG using the conventional six-switch rectifier have been reported [4, 12-14]. However, DTC of PMSG using the Vienna rectifier has not been discussed as much. In [15], a DTC of PMSG based on Vienna rectifier was presented where a switching vector Look-Up Table (LUT) was developed to allow torque and flux regulation whilst maintaining a balanced DC-link voltages system. Nevertheless, the performance of LUT-DTC is greatly dependent on the LUT, which was developed based on an offline optimization approach. Therefore, the performance is restricted to a limited region of operation for which the offline optimization is conducted. Furthermore, delays due to the digital implementation of LUT-DTC can degrade control performance. Hence, this factor should be considered when designing the LUT.

The goal of this work is to improve the classic LUT-DTC for Vienna rectifier presented in [15] using the MPC approach. A Model Predictive Direct Torque Control (MPDTC) method was developed for the Vienna rectifier to allow online optimization of switching vector selection by minimizing the cost function at every control period. To reduce the computational burden, a switching vector look-up table was utilized in conjunction with a neutral point voltage comparator to directly reduce the number of candidate switching vectors based on neutral point voltage errors. A comparison with the classic LUT-DTC method [15] was conducted to validate the performance of the proposed LUT-MPDTC approach. The core idea is to explore the DTC of PMSG on the Vienna rectifier. Therefore, the comparison of MPDTC and MPCC in performance evaluation for normal systems (which may lead to the similar conclusions as demonstrated by Siami et al. [16]) was not performed in this paper. Nevertheless, the proposed approach in this paper is using difference method of computational effort reduction compared to [10] as a MPC approach, which will be discussed in the paper.

The rest of this paper is structured as follows: Section II first explains the operation principles of the Vienna rectifier, highlighting its feasible switching vectors and operating limits. Subsequently, the PMSG equation, particularly the discrete model in rotating frame used for DTC of the PMSG is described in Section III. Then, Section IV presents the proposed MPDTC and the resulting reduction in computational burden using the LUT. Finally, the simulation and experimental results compared to the existing LUT-DTC are discussed in Section V before conclusions are drawn in Section VI.

II. WORKING PRINCIPLE OF THE VIENNA RECTIFIER

The Vienna rectifier is a three-level unidirectional converter. It consists of three active bi-directional switching units, a diode bridge comprising six diodes and two DC-link capacitors connected in series, as shown in Fig. 1 (a). Since the Vienna rectifier is used as the converter for the PMSG, its equivalent schematic is shown in Fig. 1 (b). The rectifier AC input can be modeled as three AC voltage sources whilst the DC output act as output DC current sources.

- AC side Input Voltage of Vienna Rectifier

The magnitude of the stator phase voltages is governed by the direction of the current flow and switching signals, $S$ of the active switches. Assuming a balanced three-phase system with the equal voltage on the two DC-link capacitors, the three-phase voltage can be expressed as a function of switching signals and current polarity [17]:

$$V_{an} = \frac{V_{dc}}{6} \left[ 2(1-S_a) - (1-S_b) - (1-S_c) \right] 2\text{sign}(I_a) - 1$$

$$V_{bn} = \frac{V_{dc}}{6} \left[ -2(1-S_a) + 2(1-S_b) - (1-S_c) \right] 2\text{sign}(I_b) - 1$$

$$V_{cn} = \frac{V_{dc}}{6} \left[ -2(1-S_a) - (1-S_b) + 2(1-S_c) \right] 2\text{sign}(I_c) - 1$$

(1)

where $V_{an}$, $I_a$, and $S_a$ are the phase voltage, current and switching signal of phase-$i$ respectively. Meanwhile sign($I_i$) is defined as follows:

$$\text{sign}(I_i) = \begin{cases} 1; & I_i \geq 0 \\ 0; & I_i < 0 \end{cases}$$

(2)

It is worth noting that the polarity of the three-phase currents can be used to identify the six sectors (see Fig. 2) as detailed in Table I.
The three-level voltage characteristic of the Vienna rectifier leads to twenty-five switching vectors with space vector diagram as illustrated in Fig. 2. As stated in the legend of Fig. 2, each switching vector is labelled with the corresponding switching state (i.e., P, O or N) together with the switching signal of each phase, \( S_a, S_b, S_c \) (i.e., O is 1 and P or N is 0). The switching vector index, \( n = \{0, 1, 2, ..., 7\} \) is the decimal form of the binary switching signals \( S_a, S_b, S_c \). At any given instance, there are eight feasible switching vectors to be considered, depending on the position of the stator current vector, \( I_s \). For instance, when \( I_s \) falls in sector one, only the switching vectors located in the shaded regions shown in Fig. 2 can be selected. Thus, the phase angle between current and voltage in the Vienna rectifier must be rigorously constrained within the range of \( \pm 30^\circ \) [17]. Nevertheless, this restriction is beneficial for MPDTC applications as the computational burden is naturally reduced.

<table>
<thead>
<tr>
<th>TABLE I</th>
</tr>
</thead>
<tbody>
<tr>
<td>POLARITY OF EACH PHASE CURRENT WITH ITS CORRESPONDING CURRENT SECTOR</td>
</tr>
<tr>
<td>( \text{sign}(l_a) )</td>
</tr>
<tr>
<td>1</td>
</tr>
<tr>
<td>( \text{sign}(l_a) )</td>
</tr>
<tr>
<td>( \text{sign}(l_b) )</td>
</tr>
<tr>
<td>( \text{sign}(l_c) )</td>
</tr>
</tbody>
</table>

III. MATHEMATICAL MODELLING OF PMSG

In order to aid the prediction of the torque and stator flux trajectory in MPC, it is convenient to express the PMSG model in the stator synchronous rotating frame (i.e., x-y) [18] with the stator voltages of the PMSG in xy frame being:

\[
\begin{bmatrix}
V_{sx} \\
V_{sy}
\end{bmatrix} = R_s \begin{bmatrix}
I_{sx} \\
I_{sy}
\end{bmatrix} + \frac{d}{dt} \begin{bmatrix}
\dot{\psi}_{sx} \\
\dot{\psi}_{sy}
\end{bmatrix} + \omega_s \begin{bmatrix}
-\psi_{sy} \\
\psi_{sx}
\end{bmatrix}
\]  

where \( R_s \) is the stator resistance and \( \omega_s \) denotes the speed. \( V_s, I_s, \) and \( \psi_s \) depict the stator voltage, stator current and stator flux, respectively, whereby the subscripts x and y indicate the corresponding variables for the x- and y-axis of the rotating frame.

With the voltage drop across the stator resistance neglected and stator flux aligned to x-axis (such that \( \psi_{sx} = \phi_s \) and \( \psi_{sy} = 0 \)), (6) can be simplified as:

\[
V_{sx} = \frac{d\phi_s}{dt} \]

\[
V_{sy} = \omega_s \phi_s
\]  

The rate of change of the load angle, \( \delta \) which is the difference between stator angle and rotor angle can be defined as the difference between stator speed and rotor speed, \( \omega_s \) as follows:

\[
\frac{d\delta}{dt} = \frac{d}{dt} (\theta_s - \theta_r) = \omega_s - \omega_r
\]  

The steady-state torque, \( T_e \) can be obtained as [14]:

\[
T_e = \frac{3}{2L_s} P \phi_s \phi_r \sin \delta
\]  

Taking the time derivative of (10) gives:

\[
\frac{dT_e}{dt} = \frac{3}{2L_s} P \phi_r \sin \delta + \frac{d\phi_r}{dt} \phi_s \cos \delta
\]  

Equation (11) can be expressed in x-y reference frame as:

\[
\frac{dT_e}{dt} = \frac{3}{2L_s} P \phi_r \sin \delta + (V_{sx} - \phi_r \omega_s) \cos \delta
\]  

From (7) and (12), it is clear that \( V_{sx} \) directly controls the magnitude of stator flux, whilst the torque can be changed by controlling the load angle and stator flux which is governed by \( V_{sx} \) and \( V_{sy} \). Thus, control of Vienna rectifier for the appropriate application of stator voltage is crucial for the PMSG control.

Using Euler’s first order approximation method, the time derivatives of torque and stator flux can be written as:

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\[
\frac{d}{dt} \varphi_s = \frac{1}{s_r} \left[ T_r(t+1) - T_e(t) \right]
\]

(13)

Since the electrical time constant is usually much smaller than the mechanical time constant, the rotor speed can be treated as a constant in the model prediction horizon [10, 19].

Using (7), (12) and (13), the torque and stator flux at time \((t+1)\) can be predicted by using the switching vectors of the Vienna rectifier, \(n\) as follows:

\[
T_r(t+1)_n = T_r(t) + \frac{3}{2L_s} \varphi_s(t) V_{ss,n} \sin \delta(t) + \left( V_{sy,n} - \varphi_s(t + 1) \omega_r \right) \cos \delta(t)
\]

(14)

\[
\varphi_s(t+1)_n = \varphi_s(t) + V_{ss,n} T_s
\]

(15)

IV. PROPOSED LOOK-UP TABLE MODEL PREDICTIVE DIRECT TORQUE CONTROL (LUT-MPDC)

A. Model Predictive Direct Torque Control (MPDTC)

Based on the information in Sections II and III, it is obvious that the selection of switching vector and polarity of stator current will affect the torque and flux change on PMSG as well as the DC-link voltages variation, as defined in (5), (14) and (15). By solving these equations, the effect of each switching vector on the torque, flux, and DC-link voltages can be predicted. In such a case, the best switching vector that can minimize these errors can be determined. This is the fundamental principle of the MPDTC method and its block diagram is shown in Fig. 3 (a). As shown in Fig. 3 (a), the control consists of two cascaded controllers, i.e., an outer PI speed controller and an inner MPDTC controller. Overall, the system is controlled to achieve the following objectives:

1) To minimize electromagnetic torque error.
2) To minimize stator flux error.
3) To maintain neutral point voltage balance.

The evaluation of the optimal switching vector in MPDTC is performed by defining a cost function, \(CF_n\). Due to the similar control objectives, a \(CF_n\) similar to that of a three-level NPC converter [20, 21] can be adopted for the Vienna rectifier as,

\[
CF_n = |T_r^* - T_r(t+1)| + k_\varphi |\varphi_s^* - \varphi_s(t+1)| + k_{DC} |\Delta V_{pn}(t+1)|
\]

(16)

where the first term represents the absolute error of the torque, the second term is the tracking error of the reference stator flux, and lastly, the third term defines the variation of the upper and lower DC-link capacitor voltages. The superscript * denotes the reference value of the control variables and the terms \(k_\varphi\) and \(k_{DC}\) are the weighting factor of flux and DC-link voltage respectively. The weighing factor can be tuned based on the importance of the control objectives [19, 20].

At every control cycle, the cost function for all candidate switching vectors \((n = 0, 1, 2, 3, \ldots, 7)\) are evaluated. The switching vector that gives the lowest cost, which is labelled as \(n_{opt}\) is then selected as the optimal switching vector at the time \((t+1)\) by minimizing the \(CF_n\) as follows:

\[
n_{opt} = \min_{n\in0,1,2,3,\ldots,7} (CF_n)
\]

(17)

B. Proposed LUT-MPDC

As mentioned earlier, the key challenge of MPC is the high computational burden which, in this case, depends on two main factors, i.e.:  
1) Number of candidate switching vectors to be evaluated.
2) Number of control variables (predictive models) in the cost function.

Taking the MPDTC in Section IV-A as an example, with 8 candidate switching vectors and 3 cost function control variables, the 3 predictive models (5), (14) and (15) each need to be evaluated 8 times, giving a total of \(3 \times 8 = 24\) calculations. Subsequently, the cost function \(CF\) (16) will need to be evaluated 8 times in order to find the optimal vector \(n_{opt}\) that minimizes the \(CF\). So, for a general MPC method with \(n\) candidate switching vectors and \(m\) cost function control variables, the total calculation required for the model predictive part is \((n \times m) \times n\). While it is clear that this cannot be directly equated to the cost of computation due the different complexity of each prediction model, it provides some insights on the effect of switching vectors and cost function variables on the overall computational burden of a MPC method. By reducing the number of candidate switching vectors and the number of cost function variables will therefore, help to reduce the computational burden of the MPDTC method.

To do this, a simple neutral point voltage balancing LUT is proposed to reduce the number of candidate switching vectors as well as the number of control variables simultaneously in the...
cost function. The concept and development of this LUT are elaborated here. From Table II, it is obvious that switching vectors 0 and 7 do not have any impact on the neutral point while switching vectors 1 to 6 either decrease or increase $\Delta V_{pm}$ depending on the current sector. At any time, there are always 3 switching vectors that increase $\Delta V_{pm}$ and another 3 vectors that reduce $\Delta V_{pm}$. Hence, given the knowledge of the current sector and the DC-link voltage error, the 3 switching vectors that increase $\Delta V_{pm}$ can be excluded from selection, leaving only 5 possible switching vectors. Based on this concept, a LUT can be constructed as shown in Table III.

The DC-link voltage error (i.e., $V_{pm} = V_p - V_n$) is fed into a comparator defined with an error band of $\pm h$. If $V_{pm} > +h$, only the switching vectors that give either negative or zero change in $\Delta V_{pm}$ are chosen; if the $V_{pm} < -h$, the three switching vectors that give $-\Delta V_{pm}$ are excluded. However, when $V_{pm}$ is within the error band ($+h > V_{pm} > -h$), the vector selection is slightly different. A smaller switching vector is preferred since it gives lower torque and stator flux ripples. Based on this, the big and the medium switching vectors which are located further away from the origin are not considered.

Meanwhile, the redundant switching vectors are also eliminated by removing the vector that is not shared between the two current sectors. As mentioned previously, the operation of the Vienna rectifier is dependent on the polarity of the stator current. Choosing the switching vector with the switching state that is common between the two current sectors can solve the issue of erroneous voltage generated when the current transits from one sector to the subsequent sector [22]. Such erroneous voltages occur due to the change of polarity of any phase of current if the switching signal is in the off state (i.e., switching state P or N depends on the current sector). Thereby, there will be a possibly that the vector cannot be realized for the control and as a result, an erroneous voltage arises. For instance, let’s assume the current locates in sector one, vectors 3 (POO) and 4 (ONN) are the small switching vectors. 3 will be chosen for LUT development as the state in b and c phase are composed of O state.

Since the LUT inherently reduces the DC-link voltage error, the cost function can be simplified to consider only the absolute errors of the torque and flux as follows:

$$C_{F} = |T_{e} - T_{e}(t + 1)_{n}| + k|\varphi_{s} - \varphi_{s}(t + 1)_{n}|$$

where only a weighing factor is used and hence, less tuning effort is required for the cost function. For simplicity, $k$ is fixed as the ratio between the magnitude of the nominal electromagnetic torque and nominal stator flux so that both of them can be measured with the same weight [23].

Furthermore, the optimal switching vectors, $n_{opt}$ selection is also simplified to only five candidate switching vectors as:

$$n_{opt} = \min(C_{F}^{*})$$

The block diagram of the proposed LUT-MPDT is depicted in Fig. 3(b). The combined effect of reducing number of candidate switching vectors (from eight to five) and simplifying the cost function (by removing the DC-link voltage part, i.e., (5), from the cost function) significantly reduces the computation burden of the proposed MPDT method. As compared to the full MPDT (Section IV-A), the number of models to be evaluated at each control cycle is reduced from 24 (3×8) to 10 (2×5) (two control variables, each with five candidate switching vectors to be evaluated). The number of calculations for cost function minimization in (19), is also reduced from 8 to 5. Furthermore, the $C_{F}$ that is simplified from (16) to (18) also eases the selection of weighting factor, where only one factor $k$ is needed in the latter case.

It is worth noting that the MPCC in [10] also presented an approach of computational burden reduction according to the DC-link voltage error. Based on the discussion on the comparison between the MPCC and MPTC in [16], it is expected that the MPCC method in [10] will have similar performance (in both steady-state and transient response) as the proposed LUT-MPDT method with slightly higher torque ripples but less current ripples. However, both methods differ in terms of computational burden, as will be justified as the following. By referring to Fig. 3 (c), the MPCC in [10] always requires the evaluation of the neutral point voltage model ($V_{pu}$) for all the 8 switching vectors at all times in order to reduce the candidate switching vectors from 8 to $n_{filter}$. The flux and torque current models are then evaluated for $n_{filter}$, the remaining switching vectors before the cost functions are evaluated. This gives a total of $8 + 2\times n_{filter} + n_{filter}$ predictive models to be evaluated. It is easy to see that the reduction of candidate switching vector in [10] depends greatly on the selection of boundary setting for the neutral point voltage. The smaller boundary gives a greater reduction in the number of switching vectors (resulting in a lower computational burden) at the expense of high current ripples. On the contrary, when the boundary setting is too big, it is possible that all of the 8 switching vectors will fall within the bound, resulting in no reduction of computational burden. Hence, the number of models to be evaluated in the MPCC method [10] varies from as low as 11 (i.e., $n_{filter}=1, 8+2\times1+1$) to as high as 32 (i.e., $n_{filter}=8, 8+2\times8+8$). The number of models to be evaluated depends on the tuning of neutral point voltage allowable boundary limit, which complicates its implementation.

On the other hand, the proposed LUT-MPDT do not have these issues. It uses LUT to directly select the 5 switching vectors and always evaluates only 15 ($2\times n + n$; number of

<table>
<thead>
<tr>
<th>TABLE III</th>
<th>DEVELOPED LOOK-UP-TABLE FOR COMPUTATION BURDEN REDUCTION</th>
</tr>
</thead>
<tbody>
<tr>
<td>Output state</td>
<td>Current sector</td>
</tr>
<tr>
<td>$V_{pm} &gt; +h$</td>
<td>$n$</td>
</tr>
<tr>
<td>4</td>
<td>S</td>
</tr>
<tr>
<td>5</td>
<td>S</td>
</tr>
<tr>
<td>6</td>
<td>S</td>
</tr>
<tr>
<td>7</td>
<td>Z</td>
</tr>
<tr>
<td>0</td>
<td>B</td>
</tr>
<tr>
<td>$+h &gt; V_{pm} &gt; -h$</td>
<td>$n$</td>
</tr>
<tr>
<td>2</td>
<td>M</td>
</tr>
<tr>
<td>3</td>
<td>S</td>
</tr>
<tr>
<td>5</td>
<td>S</td>
</tr>
<tr>
<td>6</td>
<td>S</td>
</tr>
<tr>
<td>7</td>
<td>Z</td>
</tr>
<tr>
<td>0</td>
<td>B</td>
</tr>
</tbody>
</table>

|$n| = \text{Size of switching vector}; \text{S = Small}; \text{M = Medium}; \text{B = Big}; \text{Z = Zero};$ 

$+h = V_{pm}$ upper error band; $-h = V_{pm}$ lower error band.
switching vector, \( n=5 \) models, i.e., \( n \) calculation of torque and flux model and the other \( n \) times for cost function equation. This means that, \( n_{\text{filter}} \) in [10] will have to be at most \( 2 \) in order to have comparable computational effort as the proposed LUT-MPDTC method; this is highly unlikely, since the value of \( n_{\text{filter}} \) fluctuates with the setting of neutral point voltage boundary.

In order to eliminate three-phase voltage sensors and further reduce the computation burden, the input phase voltage for each feasible switching vector in (1) can be pre-calculated using DC-link voltage and current sector information as shown in the Appendix (Table VI).

C. Model-based Computational Delay Compensation

A model-based delay compensation approach is used to address the computation delay for the real-time digital implementation. While there are various advanced methods to improve the computation delay, such as the trapezoidal mod with active filter [24], the simple one-step look ahead method is usually sufficient. It is done by applying the selected optimal variation in both torque and stator flux in the estimation model. In this way, the parameters at a current instant time, \( t \) is moved one step forward in time. The estimation model for both stator flux and torque are given as follows:

\[
\begin{align*}
T_e'(t) &= T_e(t) + \Delta T_{e,n_{\text{opt}}}(t-1) \\
\varphi_s'(t) &= \varphi_s(t) + \Delta \varphi_{s,n_{\text{opt}}}(t-1)
\end{align*}
\]  

(20)

where the superscript ‘\( \cdot \)’ denotes the revised variables estimation, \( \Delta T_{e,n_{\text{opt}}}(t-1) \) and \( \Delta \varphi_{s,n_{\text{opt}}}(t-1) \) are the variation of torque and stator flux with the optimal vector selected at time \( (t-1) \).

By taking the computation delay into account, the prediction model in (14) and (15) becomes:

\[
\begin{align*}
T_e(t+1) &= T_e(t) + \Delta T_{e,n_{\text{opt}}}(t) \\
\varphi_s(t+1) &= \varphi_s(t) + \Delta \varphi_{s,n_{\text{opt}}}(t)
\end{align*}
\]  

(21)

where \( \Delta T_{e,n_{\text{opt}}}(t) \) and \( \Delta \varphi_{s,n_{\text{opt}}}(t) \) are the variation of torque and stator flux for each switching vectors respectively at instant time, \( t \). They can be expressed as follows:

\[
\begin{align*}
\Delta T_{e,n_{\text{opt}}}(t) &= \frac{3\pi r}{2L_s} p \varphi_r [V_{sX_{n}} \sin \delta + (V_{sY_{n}} - \varphi_s \omega_r) \cos \delta] \\
\Delta \varphi_{s,n_{\text{opt}}}(t) &= V_{sX_{n}} T_s
\end{align*}
\]  

(22)

(23)

Fig. 4 depicts the flowchart of the LUT-MPDTC with the consideration of computation delay compensation.

V. VALIDATION AND IMPLEMENTATION

The developed LUT-MPDTC approach was first implemented in Simulink / Matlab before it was further validated using a lab-scale prototype. The parameters used for both simulation and experiment are summarized in Table IV.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Stator resistance, ( R_s )</td>
<td>2.2 Ohm</td>
</tr>
<tr>
<td>Stator inductance, ( L_s )</td>
<td>8.721 mH</td>
</tr>
<tr>
<td>Pole pairs, ( P )</td>
<td>4</td>
</tr>
<tr>
<td>Inertia, ( J )</td>
<td>0.031 mkg/m²</td>
</tr>
<tr>
<td>Rotor flux, ( \varphi_r )</td>
<td>0.06 Wb</td>
</tr>
<tr>
<td>Rated torque, ( T_{r,n_{\text{opt}}} )</td>
<td>-1.27 Nm</td>
</tr>
<tr>
<td>DC-link capacitance, ( C_1 ) and ( C_2 )</td>
<td>470 ( \mu )F</td>
</tr>
<tr>
<td>Load resistance, ( R_1 ) and ( R_2 )</td>
<td>40 Ohm</td>
</tr>
<tr>
<td>DC-link voltage, ( V_d )</td>
<td>300 V</td>
</tr>
<tr>
<td>PI speed controller and MPDTC</td>
<td></td>
</tr>
<tr>
<td>Proportional gain constant, ( K_p )</td>
<td>0.01</td>
</tr>
<tr>
<td>Integral gain constant, ( K_i )</td>
<td>0.001</td>
</tr>
<tr>
<td>Weighing factor, ( k ) for Cost function</td>
<td>21.167</td>
</tr>
<tr>
<td>Sampling frequency</td>
<td>20 kHz</td>
</tr>
</tbody>
</table>

Three operating conditions were examined for the control approaches in the study, i.e.:  
1) Steady-state operation at both high and low speed.  
2) Transient-state response during torque step change.  
3) Operation with unbalanced DC loads.

For comparison, the existing method that used the classic DTC in [15] was used. Since the real-time implementation was done digitally, model-based computation delay compensation used in the proposed LUT-MPDTC was applied to the classic DTC for a fair comparison. The effect of the error band on the neutral point voltage comparator of the LUT-MPDTC was also identified through simulation to verify the effectiveness of LUT in balancing the neutral point voltage. Parameter detuning was verified in the experiment to investigate the robustness of the proposed control scheme with parameter uncertainties.
A. Simulation Studies

1) Steady-state Operation with Low and High Speed

Firstly, the PMSG was controlled at a constant speed to examine its steady-state operation at both high and low speeds, i.e., 700 rpm and 1500 rpm as shown in Fig. 5 (a) and (b) respectively. The left graph showed the simulation results of the existing DTC and the right graph was for the proposed LUT-MPDTC. It is clear that the speed, torque and stator flux from the proposed control approach were well regulated around their respective references. The upper and lower DC-link capacitance voltages were confined within the range of 148 V to 152 V, which was due to the predefined error band of ±2 V. Additionally, the generated three-phase current for both of the approaches were balanced and sinusoidal. The FFT spectra for the current waveforms under each control approach was obtained using the Matlab FFT tool. The results are shown in Fig. 6. As expected, the FFT results show that both methods gave rise to varying switching frequency with the harmonic components spread over a range of frequencies. In terms of THD, the proposed LUT-MPDTC has better performance as compared to the classic DTC approach with the improvement of 13.5% and 11.7% for operation at 700 rpm and 1500 rpm respectively.

Comparatively, the generated torque from the classic DTC showed a few sudden spikes during high-speed operation. Meanwhile, the stator flux has a negative offset and slightly higher pulsation for both low and high-speed operations. This phenomenon was due to fact that the selection of switching vector for LUT was not optimized in the classical DTC approach.

For a detailed analysis on the torque ripples section, simulation results for the classic DTC were enlarged and focused in the time interval of 0.002 s with the switching vector...
labelled according to [15] for high PMSG operational speed as depicted in Fig. 7. The torque error \( \Delta T \) was calculated using (22) by substituting the selected switching vector. The big switching vector, 17 was selected (i.e., switching vector, 0 in Table III) which gave a huge positive variation in torque during the instant of spikes.

In terms of stator flux offset, it occurred mainly because of the non-optimized selection of switching vector for LUT. For the classic DTC approach in [15], zero vector, 25 (7 in Table III) was the only vector to be selected after each positive variation applied in torque for classic DTC. Therefore, the stator flux exhibited an offset pattern as zero vector produced zero or minimal stator flux. The performance of the classic DTC is highly dependent on the LUT. However, the developed LUT is fixed and does not change based on the PMSG operating conditions. Such simulation results showed that the proposed LUT-MPDTC can choose more optimized switching vectors that track closer to the reference of the control parameters regardless of the PMSG operational speed.

2) Transient-state Response with Torque Step Change

The dynamic response of the proposed LUT-MPDTC as well as classic LUT was demonstrated with the PI speed controller bypassed and the reference torque stepped from -1 Nm to -0.5 Nm at \( t = 1.1 \) s. It was observed that both control approaches are comparable in terms of torque response as they could achieve fast and accurate tracking performance with tracking time of 0.3 ms and 0.15 ms respectively as shown in the top figure in Fig. 8 (a) and (b). There were no ripples seen in the stator flux at the instance of torque change. Throughout the step change in torque, the DC-link voltages were still well regulated within the error band as can be observed in the bottom figures of Fig. 8 for both of the approaches. The same trend was observed in steady-state response for classic DTC approach in which the torque exhibited a spike when it was regulated at -1 Nm. Meanwhile, the stator flux was slightly offset downward from its reference at the value of 0.06 Wb. These results show that the proposed LUT-MPDTC worked well when the reference torque was changed suddenly.

3) Operation with Unbalanced DC Loads

Next, both control schemes were tested under an unbalanced load condition. A resistor (330 Ohms) was connected in parallel to the lower side of the DC link capacitor, \( C_2 \). This will result in an unbalanced load for the capacitor with approximately 10% difference in resistance value between the upper and lower load. The simulation results were displayed in Fig. 9, where both control schemes were capable of maintaining the neutral point voltage balance with the upper and lower DC-link capacitance voltage magnitude close to 150 V. However, the generated torque of the classic DTC showed a distorted waveform as compared to the proposed control, and even more severe ripples were observed in stator flux.

To be more specific, the stator flux at the distorted section was zoomed in and shown in Fig. 10. The selected switching
vectors, the variation of stator flux, $\Delta \phi_s$ and stator flux are displayed from top to bottom. The $\Delta \phi_s$ was calculated using (23). As can be clearly seen in Fig. 10, the vector selected, 18 (vector 1 when current positioned at sector three in Table III) was incorrect, leading to a negative change of stator flux which was opposing the desired value, as indicated with the dotted line at 0.06 Wb. The selected variation of stator flux caused the generated flux to be no longer within the pre-set comparator band (i.e., ±0.005 Wb). Henceforth, the generated stator flux exhibited higher ripples. It is evident from the simulation results that the LUT-MPDTC outperformed the existing DTC under an unbalanced load condition. Overall, the simulation results demonstrated that the use of the LUT-MPDTC method eliminates the difficulty of complex LUT development and allows more optimal selection of switching vectors which is adaptive to the PMSG operation conditions.

4) Error Band Effect on The Neutral Point Voltage Comparator
The effect of error band setting for neutral point voltage comparator was investigated and its results were depicted in Fig. 11. Initially, the band was set at 1 V and is changed to 1.5 V and 2 V at the time, $t = 0.6$ s and 0.8 s respectively. It was observed that the neutral point voltage, $V_{pn}$ was kept closely within the pre-set band regardless of the band magnitude. However, smaller band settings resulted in higher ripples for both torque and stator flux. Switching vectors from LUT of $V_{pn}$ state 0 and 2 tend to be selected more frequently for balancing purposes as shown in the lowest graph in Fig. 11, which means that tighter condition for $V_{pn}$ can result in higher ripples in both torque and stator flux. A higher ripple for torque and stator flux was expected since there was the tendency of big switching vectors to be selected for the neutral point voltage balancing. It is worth to note that the output from neutral point voltage comparator, $V_{pn}$, state is 2 when $V_{pn}$ is larger than the upper error band, 1 when $V_{pn}$ is within the error band and, and 0 when $V_{pn}$ is less than lower error band. When the band was set to 1.5 V and 2 V, ripples exhibited in both torque and stator flux were less than 1 V and they did not show a significant difference for
both of these band settings. This was due to the fact that the
switching vectors designed at \( V_m \) state of 1 gave smaller
magnitude change in both stator flux and torque. Through this
simulation, we can conclude that the neutral point voltage can be
confined to be as small as 1.5 V with a satisfactory output
performance of torque and stator flux. This also demonstrated
the effectiveness of the LUT in balancing the DC-link voltages
without the need of using a cost function.

B. Experimental Results

A laboratory prototype has been implemented with the same
parameters as the simulation (Table IV) to further validate the
performance of the proposed control scheme. The control
algorithm was programmed to be operated at 20 kHz in a Texas
Instrument based TMS320F28335 development board. The
Vienna rectifier was built with the power modules IXYS
VUM25-05E [25]. The generator was driven by a DC motor fed
with a controllable DC power supply. Fig. 12 shows the
overview of the experimental set-up. The DC-link voltage on the
Vienna rectifier was maintained at a constant value of 300
V by an adjustable DC power supply. The generated torque and
stator flux from the PMSG were calculated from the measured
stator current and rotor position and plotted using Matlab. The
studies were carried out while considering the same conditions
applied in the simulation studies. To ensure fair comparison of
the experimental results between the two control approaches,
computational delay due to digital implementation of the classic
DTC was compensated using the similar approach as presented
in Section IV-C. As the results, the classic DTC could be
operated in the same conditions as what was conducted in the
simulation studies.

Fig. 13 shows the experimental results of both control
schemes depending on the PMSG speed. As seen in Fig. 13 (a)
and (b), accurate tracking of rotor speed and torque was
achieved for both control approaches. The generated upper and
lower DC-link voltages were controlled within the limit for both
speeds. Nevertheless, it was clearly seen that the LUT-MPDTC
gave better torque and flux regulation with lower speed error.
Similar to the simulation results, the issue of stator flux offset
was obvious for the classic DTC while the LUT-MPDTC was
able to maintain good performance at both high and low speeds.

The recorded current waveform using oscilloscope were
imported into Matlab to perform FFT analysis. The findings
were similar to those obtained from the simulation. From the
FFT spectra in Fig. 14, it is seen that both the classic DTC and
the LUT-MPDTC gave varying frequency harmonics. When
the speed increased from 700 to 1500 rpm, the harmonics
components shifted more towards the higher frequency
spectrum, which were in agreement with the observations from
the simulation. Overall, the current THDs for the proposed
LUT-MPDTC are lower than those of the classic DTC. The
varying switching frequency is a known issue which can restrict
the application for higher power generators. This issue can be
mitigated using space vector/pulse-width modulation or with
optimal switching sequence selection as discussed in [26-28].
However, they are beyond the scope of this paper which shall
not be discussed further.

The dynamic behaviour of the PMSG was tested and
displayed in Fig. 15. It was observed that the generated torque
from the LUT-MPDTC can track closely to the new reference
torque. Unlike the classic DTC, the obtained torque did not

generate spikes when it was regulated at both -1 Nm and -0.5
Nm respectively. At the same time, the abrupt change did not
affect the LUT-MPDTC in regulating the stator flux as well as
the upper and lower DC-link voltages as shown in Fig. 15.

The experiment results for unbalanced DC load condition
are shown in Fig. 16. It is apparent from the experimental
results that both control approaches were able to balance the
DC-link voltages despite the unbalanced load. The upper and
lower DC-link voltages were confined within the designated
band value. Nevertheless, the torque and stator flux obtained
from the existing DTC method exhibited higher ripples as
compared to the proposed LUT-MPDTC during the imbalanced
load condition. Such phenomena agreed with the simulation
where the proposed method can perform better when there is an
unbalanced at the Vienna rectifier’s DC output.
Fig. 13. Experimental results for the overall response of the classic DTC and of LUT-MPDTC at both low and high speed. (a) 700 rpm. (b) 1500 rpm.

Fig. 14. FFT analysis on generated stator current. (a) DTC at 700 rpm. (b) LUT-MPDTC at 700 rpm. (c) DTC at 1500 rpm. (d) LUT-MPDTC at 1500 rpm.
experimental tests were performed with the LUT-MPDTC using different values of $L_s$ and $\phi_r$. Fig. 17 showed the experimental results during machine parameter variation when the rotor speed was controlled at 1500 rpm. There was no significant increase in torque ripples when stator inductance was reduced by 50%. The torque ripples increased slightly when the variation was up to 200%. In terms of stator flux, reducing stator inductance introduced a slight positive offset, while increasing $L_s$ caused a negative offset. Stator flux ripples also increased with the changes in $L_s$. For the case with rotor flux variation, the produced torque did not exhibit significant ripples, but a small offset was detected in stator flux. A slightly higher ripples appeared in the stator flux was observed when rotor flux difference of 200% was applied, see Fig. 17. From the results, it can be concluded that the proposed LUT-MPDTC is relatively robust against parameter detuning. It is expected to be able to perform reasonably even with the presence of parameters uncertainties.

D. Performance Comparisons for Difference Control Methods

Table V summarizes the differences between the classic DTC and LUT-MPDTC. To further illustrate the advantage of LUT-MPDTC as an improved MPDTC method, the full MPDTC method discussed in Section IV-A is included in the table as well. The computation cost for the three methods were quantified using the Matlab profiler tool. The processor clock speed was 2,094 MHz with clock precision of $3 \times 10^9$ s. The time was recorded for each respective control algorithms after they had run for 20,000 times. It is interesting to see that the LUT-MPDTC gives a 47% reduction in computation burden compared to the full MPDTC. This is due to the reduction in candidate switching vectors and the simplification of the cost function. With only two variables and five candidates switching vectors from the LUT, it leads to only 10 models to be computed (instead of 24) by the LUT-MPDTC at each control period. Furthermore, the computational cost for LUT-MPDTC is actually very close to that of the classic DTC. This is mainly due to the fact that Matlab populates the LUTs every time the code is executed. As a result, higher cost is induced for the classic DTC and the LUT-MPDTC methods which uses LUTs. Despite this, it can be concluded that the LUT-MPDTC provides significant computational cost reduction compared to the full MPDTC.

Quantitative evaluation of the control performance was conducted based on the standard deviation for torque and stator flux as well as THD for the stator current during steady-state condition at both high and low rotational speeds. The main attractive feature of the LUT-MPDTC is its capability to achieve an optimal control performance over a wide range of PMSG speed operation as the standard deviation generated for both torque and stator flux is always lower as compared to classic DTC approach.

Unlike the existing DTC approach, the desired changed of stator flux can be achieved when there was an unbalanced at the DC-link output when compared to the classic DTC method. The stator flux was improved 49% in terms of standard deviation. These comparison results reveal that proposed LUT-MPDTC can give superior performance compared to both classic DTC and full MPDTC.
VI. CONCLUSION

This paper has developed a Model Predictive Direct Torque Control (MPDTC) method for PMSG controlled by a Vienna Rectifier, which has been shown to be an improved DTC method compared to previous similar work [15]. The notable points achieved here in this paper include:

1) The limitations of the classic DTC of PMSG with Vienna Rectifier, i.e., [15] were identified by simulation and experimental results.
2) A MPDTC for PMSG with Vienna Rectifier as an improved DTC method to the work presented in [15] (Rajaei et al, 2011) was successfully implemented.
3) An improved LUT-MPDTC with significant computation burden reduction compared to the full MPDTC was demonstrated.

As compared to the classical DTC approach, the LUT-MPDTC method developed in this paper allows online optimization of switching vector selection, which provides better performance over a wider range of operations. Furthermore, the use of LUT concept simplifies the complexity of the full MPDTC by reducing the number of candidates of switching vector from eight to five. This results in a significant reduction of computational burden at the value of 47% lesser in computational cost. Simulation and experimental results showed that the proposed control scheme allows online optimization control of the torque and stator flux for the PMSG using a Vienna rectifier whilst maintaining a balanced DC-link voltages system. The proposed method outperforms the classic DTC approach especially when there is an unbalanced load at the DC-link output by generating lower torque and stator flux ripples, which reflects lower standard deviation by 33% and 49%, respectively.

TABLE V

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Algorithm complexity</td>
<td>Low</td>
<td>High</td>
<td>Medium</td>
</tr>
<tr>
<td>Implementation complexity</td>
<td>Indirect</td>
<td>Direct</td>
<td>Direct</td>
</tr>
<tr>
<td>Complexity of LUT development</td>
<td>Complex</td>
<td>-</td>
<td>Simple</td>
</tr>
<tr>
<td>Optimization method</td>
<td>Offline</td>
<td>Online</td>
<td>Online</td>
</tr>
<tr>
<td>Computation cost (s)</td>
<td>25.6</td>
<td>48.2</td>
<td>25.5</td>
</tr>
<tr>
<td>Computation burden</td>
<td>Low</td>
<td>High</td>
<td>Medium</td>
</tr>
</tbody>
</table>

Performance evaluation based on experimental results

<table>
<thead>
<tr>
<th>Condition</th>
<th>Parameter</th>
<th>Stator flux (Wb)</th>
<th>Torque (Nm)</th>
<th>Stator flux (Wb)</th>
<th>Torque (Nm)</th>
<th>Stator flux (Wb)</th>
<th>Torque (Nm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1500 rpm, -1Nm</td>
<td>Stator flux</td>
<td>0.0023</td>
<td>0.069</td>
<td>-</td>
<td>0.0021</td>
<td>-</td>
<td>0.055</td>
</tr>
<tr>
<td></td>
<td>Torque</td>
<td>11.18%</td>
<td>0.0572</td>
<td>Current THD</td>
<td>13.22%</td>
<td>-</td>
<td>0.056</td>
</tr>
<tr>
<td>700rpm, -1Nm</td>
<td>Stator flux</td>
<td>0.0055</td>
<td>0.094</td>
<td>Stator flux</td>
<td>0.0055</td>
<td>-</td>
<td>0.063</td>
</tr>
<tr>
<td></td>
<td>Torque</td>
<td>13.22%</td>
<td>0.094</td>
<td>Torque</td>
<td>0.094</td>
<td>-</td>
<td>0.063</td>
</tr>
<tr>
<td></td>
<td>Current THD</td>
<td>10.03%</td>
<td>10.36%</td>
<td>Current THD</td>
<td>10.36%</td>
<td>-</td>
<td>0.063</td>
</tr>
</tbody>
</table>

Fig. 17. Experimental results for parameter detuning study.
(a) Generated torque for $L_s$ variation. (b) Stator flux results for $L_s$ variation. (c) Generated torque for $\phi_r$ variation. (d) Stator flux results for $\phi_r$ variation.
APPENDIX

TABLE VI
PRE-TABULATED STATOR PHASE VOLTAGE USING AVAILABLE SWITCHING VECTORS AT ITS RESPECTIVE CURRENT SECTOR

<table>
<thead>
<tr>
<th>Phase</th>
<th>Index, n</th>
<th>Current sector</th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>a</td>
<td>1</td>
<td>$V_{dc}$, $V_{dc}$, $-V_{dc}$, $-V_{dc}$, $V_{dc}$, $V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>$2$, $6$, $2$, $2$, $6$, $2$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>$V_{dc}$, $V_{dc}$, $-V_{dc}$, $-V_{dc}$, $V_{dc}$, $V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4</td>
<td>$3$, $3$, $3$, $3$, $3$, $3$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>$V_{dc}$, $-V_{dc}$, $-V_{dc}$, $-V_{dc}$, $V_{dc}$, $V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>6</td>
<td>$3$, $3$, $3$, $3$, $3$, $3$</td>
<td></td>
</tr>
<tr>
<td>b</td>
<td>1</td>
<td>$-V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$, $-V_{dc}$, $-V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>$0$, $0$, $0$, $0$, $0$, $0$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>$-V_{dc}$, $-V_{dc}$, $V_{dc}$, $V_{dc}$, $-V_{dc}$, $-V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4</td>
<td>$6$, $6$, $6$, $6$, $6$, $6$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>$V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>6</td>
<td>$6$, $6$, $6$, $6$, $6$, $6$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>7</td>
<td>$0$, $0$, $0$, $0$, $0$, $0$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>8</td>
<td>$0$, $0$, $0$, $0$, $0$, $0$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>9</td>
<td>$2V_{dc}$, $V_{dc}$, $V_{dc}$, $-2V_{dc}$, $-2V_{dc}$, $-2V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>10</td>
<td>$3$, $3$, $3$, $3$, $3$, $3$</td>
<td></td>
</tr>
<tr>
<td>c</td>
<td>1</td>
<td>$0$, $-V_{dc}$, $0$, $0$, $0$, $0$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>2</td>
<td>$-V_{dc}$, $-V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>3</td>
<td>$2$, $2$, $6$, $2$, $2$, $6$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>4</td>
<td>$6$, $6$, $6$, $6$, $6$, $6$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>5</td>
<td>$V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$, $V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>6</td>
<td>$6$, $6$, $6$, $6$, $6$, $6$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>7</td>
<td>$0$, $0$, $0$, $0$, $0$, $0$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>8</td>
<td>$0$, $-V_{dc}$, $0$, $0$, $0$, $0$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>9</td>
<td>$V_{dc}$, $3$, $3$, $3$, $3$, $3$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>10</td>
<td>$0$, $0$, $0$, $0$, $0$, $0$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>11</td>
<td>$2V_{dc}$, $V_{dc}$, $V_{dc}$, $-2V_{dc}$, $-2V_{dc}$, $-2V_{dc}$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>12</td>
<td>$3$, $3$, $3$, $3$, $3$, $3$</td>
<td></td>
</tr>
</tbody>
</table>

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