Optimal switching sequence model predictive control for three-level NPC grid-connected inverters

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Abstract
In order to concentrate the frequency spectrum of the output voltage and improve the quality of grid currents for the three-level neutral point clamped inverter with the model predictive control, this paper proposes an optimal switching sequence model predictive control algorithm. Based on the increments of grid currents and the neutral-point voltage, the predictive model of the inverter is established in $\alpha\beta$ frame. Moreover, including grid currents and the neutral point voltage tracking, the cost function is designed and simplified by arranging the voltage vector sequence in a sampling cycle appropriately. Meanwhile, the calculation of the optimal dwell time for each voltage vector sequence is derived by Lagrange multiplier method, and its solving process is simplified to reduce online computations. Furthermore, according to different voltage vector sectors, a voltage vector sequence preselection principle is introduced in this paper, besides, considering redundant small vectors in a pair have opposite affections on the neutral point voltage, the amount of voltage vector sequence that needs to be verified is further cut down, for obtaining the optimal voltage vector sequence. Finally, the experimental results show that the proposed method can concentrate the output frequency spectrums, which maintains a good quality of grid currents with a small neutral-point voltage fluctuation.

1 | INTRODUCTION

Currently, with the rapid development of digital signal processors (DSP), some complex control algorithms have been researched extensively as effective approaches for controlling power converters and electrical drives, such as sliding model control, fuzzy logic control, model predictive control (MPC), etc. [1,2]. Among non-linear control strategies, the MPC is preferred in power electronic application, which has advantages like simple implementation, fast dynamic response and strong robustness [3,4].

As one of the most representative MPC methods, the finite control set MPC (FCS-MPC) method has been widely researched. In [5], an FCS-MPC method is proposed for a three-phase neutral-point clamped (NPC) inverter. However, some shortcomings of the FCS-MPC restrict its application in power converters. When conventional FCS-MPC method is applied to power converters, only an output voltage vector that minimizes the cost function is applied during the whole sampling cycle, which leads to the unfixed switching frequency of the inverter output voltage [6–8]. The unfixed switching frequency of output voltage not only affects the design of filtering inductor for grid-connected inverters [9] but also introduces more harmonics into grid currents [10,11]. In addition, the conventional FCS-MPC also uses a high sampling frequency to achieve good performances, which leads to larger inductors. These issues have been researched in some literature. In [12], a control algorithm consists in setting a fixed switching frequency and dividing it into smaller evaluation steps is proposed to fix the switching frequency, but the control performance for grid currents is unsatisfactory. Hassine et al. [13] propose a method of increasing the virtual voltage vector with fixed amplitude and phase angle to improve the control accuracy of the system and concentrate the frequency spectrum of the output voltage. However, the non-zero average steady-state tracking error produced by above methods cannot be eliminated, because that these FCS-MPC methods generate the output voltage for the inverter by selecting one optimal vector during a sampling cycle essentially.
In order to solve the problems mentioned above, some modified FCS-MPC methods, introducing the idea of traditional space vector modulation (SVM) method, are proposed. In [14–18], a modulated MPC algorithm is proposed for the indirect matrix converter to obtain the fixed switching frequency by using a similar output mode with SVM method to control the input instantaneous reactive power and output current. Zhou et al. [19] propose an extended control set model predictive control (ECS-MPC) algorithm using a new voltage vector synthesis method to improve the control precision for the MPC strategy. Optimal switching sequence model predictive control (OSS-MPC) is a kind of modified FCS-MPC [20], which applies an optimal vector sequence in a control cycle instead of an optimal output vector. Thereby, the OSS-MPC has a better performance than conventional FCS-MPC. Moreover, the OSS-MPC has the advantage of a constant switching frequency [21]. Vazquez et al. [22] propose a predictive optimal switching sequence direct power control (OSS-DPC) algorithm for grid-connected converters, and the OSS-DPC can provide the desired power references by calculating globally optimal switching sequences. In [23], two adjacent voltage vectors and zero voltage vector are utilized in one sampling cycle and the optimal duty cycles for all sectors are calculated to obtain a good control performance.

Most of the literature mentioned above focus on modifying OSS-MPC for the two-level converter topologies, but, the three-level (3-L) NPC inverter has a neutral-point voltage unbalancing problem, which appears due to the capacitance error, the inconsistent characteristics of switching devices, the unbalanced operation of three-phase inverters, etc. Vazquez et al. [24] extended the MPC method using the optimal switching sequence to a single-phase NPC inverter, and the fixed switching frequency in the output voltage of inverter can be obtained. In [21], an OSS-MPC method for the three-phase Vienna rectifier is proposed. However, the dc-link voltage is controlled by a proportional-integral controller without illustrating the neutral-point voltage control. In [25], a cascaded-OSS-MPC strategy is proposed for a 3-L NPC converter, which uses two MPC controllers in series to achieve each goal separately without a weighting factor. However, a tuning parameter is introduced to trade the grid current tracking error versus control input effort, which actually affects the grid current control.

Above all, the control of the neutral-point voltage and grid currents both have an effect on grid currents quality, therefore, it is necessary to research the coordinated control for the neutral-point voltage and grid currents of the 3-L NPC inverter with OSS-MPC. However, it is hard to obtain the duty cycle for each optimal voltage vector sequence (VVS) directly, according to the control target that minimizing the neutral-point voltage fluctuation meanwhile obtaining a good performance on the grid current control.

This paper, based on the derived predictive model, which indicates the increments of grid currents and the neutral point voltage in \( \alpha \beta \) frame, proposes a modified OSS-MPC method to concentrate the frequency spectrum of the output voltage and improve the quality of grid currents for three-level NPC inverter, decreasing the neutral-point voltage fluctuation. In Section 2, according to the increments of grid currents and neutral-point voltage, the predictive model of three-level NPC inverter with the proposed OSS-MPC method is established in \( \alpha \beta \) frame, and the cost function is designed. Besides, based on the effects of different output voltage vectors on the neutral-point voltage increment, the arrangements for voltage vectors synthesis in each VVS are set to reduce the online computation for the cost function. In Section 2, considering the dwell time constraints for an output VVS, the possible solutions are derived to solve the optimal dwell time for a VVS by Lagrange multiplier method. And the solving process is also simplified to reduce the online calculation. In Section 2, regarding different voltage vector sectors, a vector sequences preselction method is introduced to decrease the amount of the voltage vector sequences that needs to be evaluated to obtain the optimal one. Meanwhile, the amount of voltage vector sequences is finally set to five for online computation by judging the sign of the neutral-point voltage and grid currents in a control cycle. The output optimal VVS with its dwell time are applied to achieve the coordinated control for the neutral-point voltage and grid currents, and the entire flowchart is given in this paper. To verify the effectiveness of the proposed method, a three-level NPC grid-connected inverter prototype is built in Section 5. The experimental results show that compared with conventional FCS-MPC method in [5] and ECS-MPC method in [19], the proposed OSS-MPC method can concentrate output frequency spectrums for the inverter, which has a better grid currents quality with a smaller neutral-point voltage fluctuation.

### 2 Model of three-level NPC inverter

The topology of a three-level NPC grid-connected inverter is shown in Figure 1, where \( I_a = I_b = I_c = I \) is the filter...
inductance and $R$ ($R_a = R_b = R_c = R$) is the parasitic resistance of filter inductor. $e_i$ and $i_i$ ($i = a, b, c$) represent the grid voltage and grid current, respectively. There are three switching states for each phase leg, the switching state of each phase $S_r^i$ is defined as

\[
S_r^i = \begin{cases} 
1 & S_{r1} = \text{on}, S_{r2} = \text{on}, S_{r3} = \text{off}, S_{r4} = \text{off} \\
0 & S_{r1} = \text{off}, S_{r2} = \text{on}, S_{r3} = \text{on}, S_{r4} = \text{on} \\
-1 & S_{r1} = \text{off}, S_{r2} = \text{off}, S_{r3} = \text{on}, S_{r4} = \text{on} 
\end{cases} \tag{1}
\]

where $S_r^i$ ($i \in \{a, b, c\}$ and $r \in \{1, 2, 3, 4\}$) denote the power switch in the main circuit, “on” and “off” represent that the power switch is turned on and turned off, respectively.

Thereby, the three-level NPC inverter has $3^3 = 27$ switching state combinations. Each switching state combination corresponds to one output voltage vector, and the three-level NPC inverter can generate 27 output voltage vectors, which are shown in Figure 2. The output voltage vector $v_o$ can be expressed as

\[
v_o = \frac{2}{3} \left( u_{aoe} e^{j0} + u_{boe} e^{j\frac{2\pi}{3}} + u_{coe} e^{j\frac{4\pi}{3}} \right), \quad n = 1, \ldots, 27 \tag{2}
\]

where $V_{dc}$ represents the dc-link bus voltage and $u_{ao}$, $i \in \{a, b, c\}$ represents the three phase output voltage of the inverter.

Besides, three-phase output voltage can be expressed as

\[
u_{io} = S_r^i \cdot \frac{V_{dc}}{2} \quad i = a, b, c. \tag{3}
\]

The corresponding components of $u_{io}$ in $\alpha\beta$ frame can be derived as

\[
\begin{bmatrix} u_{io} \\ u_{i\beta} \end{bmatrix} = T_{3-2} \begin{bmatrix} u_{io} \\ u_{i\alpha} \end{bmatrix} \tag{4}
\]

where $u_{io}$ and $u_{i\beta}$ represent the output voltage of three-level NPC inverter in $\alpha\beta$ frame, $T_{3-2}$ represents the transformation matrix of the Clarke transformation, which is given by

\[
T_{3-2} = \frac{2}{3} \begin{bmatrix} 1 & -1/2 & -1/2 \\ \sqrt{3}/2 & -\sqrt{3}/2 \\ 0 & 0 & 1 \end{bmatrix}. \tag{5}
\]

Besides, define $i_a$, $i_b$ and $e_a$, $e_b$ represent the grid current and grid voltage in $\alpha\beta$ frame respectively, which are transformed from $e_i$ and $i_i$ ($i = a, b, c$).

Thereby, the model of the main circuit in $\alpha\beta$ frame can be described as

\[
\begin{cases} 
L_a \frac{di_a}{dt} = u_{io \alpha} - R_i e_a \\
L_b \frac{di_b}{dt} = u_{io \beta} - R_i e_b 
\end{cases} \tag{6}
\]

where $i_a$, $i_b$ and $e_a$, $e_b$ represent the grid current and grid voltage in $\alpha\beta$ frame respectively; $u_{io \alpha}$ and $u_{io \beta}$ represent the output voltage of the inverter in $\alpha\beta$ frame.

According to Figure 1, by transforming three phase grid $i_i$ ($i = a, b, c$) to $i_a$, $i_b$ using Equation (5), the neutral point current of three-phase three-level NPC inverter $i_0$ can be derived as

\[
\begin{bmatrix} i_a \\ i_b \\ i_0 \end{bmatrix} = \frac{1}{3} \left[ (1 - |S_r^i|) \left(1 - |S_r^i|\right) \right] \cdot T_{3-2}^{-1} \begin{bmatrix} i_a \\ i_b \\ i_0 \end{bmatrix} \tag{7}
\]

where $i_{a1}$ and $i_{b2}$ are currents through capacitor $C_1$ and $C_2$, respectively. $v_r$ represents voltage difference of $C_1$ and $C_2$, $C$ is their capacitance.

If the sampling frequency is much higher than the fundamental frequency, the variation of grid current and neutral-point voltage can be viewed as constants in a sampling cycle. By substituting Equations (3) and (4) into Equation (6) and combined with Equation (7), and an output voltage vector $v_o$ is applied, the increments of grid current and neutral-point voltage at $k$th
simplification of cost function can be expressed as

\[
\begin{align*}
    f_{αα} &= \frac{dα}{dν} \bigg|_{αα} = \frac{1}{L} \left[ \frac{V_α}{3} \left( S_α - 2S_β - \frac{1}{2}S_γ \right) - R_{αα}(k) - e_α(k) \right] \\
    f_{ββ} &= \frac{dβ}{dν} \bigg|_{ββ} = \frac{1}{L} \left[ \frac{V_β}{3} \left( \sqrt{3}S_β - \sqrt{3}S_β \right) - R_{ββ}(k) - e_β(k) \right] \\
    f_{νν} &= \frac{dν}{dν} \bigg|_{νν} = -\frac{1}{C} \left( |S_ν| - \frac{1}{2}|S_ν| - \frac{1}{2}|S_ν| \right) i_α(k) \\
    &\quad - \frac{1}{C} \left( \sqrt{3} |S_ν| - \sqrt{3} |S_ν| \right) i_β(k)
\end{align*}
\]

∀n ∈ {1, ..., 27}

(8)

where \( f_{αα}, f_{ββ}, f_{νν} \) represent the increments of grid current in \( αβ \) frame and the neutral-point voltage respectively when the output voltage vector \( ν_n \) is applied.

### 2.2 Predictive model and cost function

In order to concentrate the frequency spectrum of the output voltage, improve the quality of grid currents, and decrease the neutral-point voltage fluctuation of three-level NPC inverter, the output signal of the proposed OSS-MPC method is modified to a three-voltage-vector sequence instead of one voltage vector with conventional FCS-MPC in a sampling cycle. As shown in Figure 2, the distribution diagram of the output voltage vectors can be divided into six big sectors, and there are four small triangle areas in each big sector. Besides, the voltage vectors for constituting a three-voltage-vector sequence are selected according to the principle similar to the nearest sector vector space pulse width modulation. An arbitrary output VVS is described as

\[ V_ν = \{ ν_{i1}, ν_{i2}, ν_{i3} \}, i1, i2, i3 ∈ \{1, ..., 27\} \]

(9)

where the output voltage vectors \( ν_{i1}, ν_{i2}, ν_{i3} \) are applied during a sampling cycle, which are random three different voltage vectors that belong to the same small triangle area in each sector.

The dwell time of the three voltage vectors \( ν_{i1}, ν_{i2}, ν_{i3} \) is defined as \( t_1, t_2, t_3 \), respectively, and the dwell time should satisfy the following equation:

\[ t_1 + t_2 + t_3 = T_s \]

(10)

where \( T_s \) is the sampling period.

As assumed above, the sampling frequency is much higher than the fundamental frequency. Therefore, the measured values of grid current and neutral-point voltage at \( k \)th sampling instant can be viewed as constants during a sampling cycle \( T_s \). Then, the predicted values of grid currents and the neutral-point voltage at \((k+1)\)th sampling instant can be calculated as

\[
\begin{align*}
    i_α^n(k+1) &= i_α(k) + \sum_{j=1}^{3} f_{αα} t_j \\
    i_β^n(k+1) &= i_β(k) + \sum_{j=1}^{3} f_{ββ} t_j \\
    ν_i^n(k+1) &= ν_i(k) + \sum_{j=1}^{3} f_{νν} t_j
\end{align*}
\]

(11)

where \( i_α(k), i_β(k), ν_i(k) \) are the measured values of grid currents and the neutral-point voltage at \( k \)th sampling instant, \( i_α^n(k+1), i_β^n(k+1) \) and \( ν_i^n(k+1) \) represent the predicted values of grid currents and the neutral-point voltage at \((k+1)\)th sampling instant, respectively.

In order to achieve the coordinated control of grid currents and the neutral-point voltage, the cost function for the proposed OSS-MPC method is defined as

\[ g = (ε_α^{k+1})^2 + (ε_β^{k+1})^2 + λ(ε_ν^{k+1})^2 \quad (λ ≥ 0) \]

(12)

where \( ε_α^{k+1} \) and \( ε_β^{k+1} \) are the tracking errors of grid currents in \( αβ \) frame, \( ε_ν^{k+1} \) is the neutral-point voltage tracking error at \((k+1)\)th sampling instant, \( λ \) is a weighting factor, and the tracking errors can be expressed as

\[
\begin{align*}
    ε_α^{k+1} &= i_α^n(k+1) - i_α^n(k) = i_α^n(k+1) - i_α(k) - \sum_{j=1}^{3} f_{αα} t_j \\
    ε_β^{k+1} &= i_β^n(k+1) - i_β^n(k) = i_β^n(k+1) - i_β(k) - \sum_{j=1}^{3} f_{ββ} t_j \\
    ε_ν^{k+1} &= 0 - ν_i^n(k+1) = -ν_i(k) - \sum_{j=1}^{3} f_{νν} t_j
\end{align*}
\]

(13)

where \( i_α^n(k+1) \) and \( i_β^n(k+1) \) are grid current reference values at \((k+1)\)th sampling instant. \( f_{αα}, f_{ββ}, f_{νν} \) are defined as the initial tracking errors of grid currents at \( k \)th sampling instant.

### 2.3 Simplification of cost function

Substituting Equation (13) into Equation (12), the detailed expression of the cost function can be obtained. However, the
TABLE 1 Voltage vector arrangements and increments calculation

| Three voltage vectors combination types | Arrangements of voltage vectors sequence \( V_t \) | Increments calculation \( f_{v1} \) | \( f_{v2} \) | \( f_{v3} \) |
|----------------------------------------|---------------------------------------------|-----------------|----------|----------|
| (1) \( v_s, v_m, v_b \)               | \( \{v_s, v, v_b\} \)                        | \( \sqrt{ } \)  | \( \sqrt{ } \) | 0        |
| (2) \( v_1, v_2, v_3 \)               | \( \{v_1, v_2, v_3\} \)                      |                 |          |          |
| (3) \( v_1, v_2, v_m \)               | \( \{v_m, v_1, v_2\} \) or \( \{v_m, v_2, v_1\} \) | \( \sqrt{ } \)  |          |          |

A detailed expression of the cost function is too complex to facilitate subsequent calculations. For convenience, the cost function can be expressed simply, which includes the increments of grid currents and the neutral-point voltage respectively, tracking errors and the vector dwell time values:

\[
g = G \left( e^{0}_v, e^{0}_g, v_{t}, f_{a\,j}, f_{b\,j}, f_{c\,j}, t_j \right) \quad j \in \{1, 2, 3\}. \quad (14)
\]

In order to reduce the online calculation burden, based on the different effects of big, medium, small, zero output voltage vectors on the neutral-point voltage, the cost function can be simplified by arranging the VVS in a sampling cycle appropriately. For a three-voltage-vector sequence, according to (9), the three voltage vectors \( v_1, v_2, v_3 \) are applied during a sampling cycle, the rules of voltage vector arrangements and the neutral-point voltage increments calculation are shown in Table 1, where the \( v \), \( v_m \), \( v_b \), \( v_\delta \) and \( v_\xi \) represent the small vector, middle vector, big vector and zero vector respectively, among the entire output voltage vectors in Figure 2.

As shown in Table 1, in a control cycle, the output voltage vector sequence \( V_t \) has three combination types, and based on which, the arrangements of VVS are set here. In each VVS arrangement, the first and the second voltage vector have impact on the neutral-point voltage, which generate increments for adjusting the neutral-point voltage. Besides, when the last voltage vector is a big vector or a zero vector, its corresponding neutral-point voltage increment is zero. Meanwhile, if the last voltage vector is a small vector, its corresponding neutral-point voltage increment can also be zero, by introducing its paired positive or negative small vector, which has the opposite effects on the neutral-point voltage. And the total dwell time of small vector in pair is the same as that of the original small vector. Therefore, by this modification, the control of the grid currents is not affected. Based on the voltage vector arrangements in Table 1, considering the redundant small vectors, there are 16 voltage vector sequences for each sector in Figure 2.

Above all, for each three-voltage-vector sequence, the calculation number for the increments of neutral-point voltage can be decreased to 2, the cost function can be modified as

\[
g = G \left( e^{0}_v, e^{0}_g, v_{t}, f_{a\,j}, f_{b\,j}, f_{c\,j}, f_{v1\,2}, t_j \right) \quad j \in \{1, 2, 3\}. \quad (15)
\]

Besides, the control of the grid currents is not affected by this simplification.

3 | ANALYSIS AND CALCULATION OF VECTOR DWELL TIME

3.1 Basic principle of vector dwell time calculation

Based on the analysis in Section 2, when a voltage vector sequence \( V_t \) is applied at \( k \)th sampling instant, and the grid currents and the neutral-point voltage are acquired. Supposing that their tracking errors and increments are obtained, by combining with the dwell time calculation of the voltage vectors, then the optimal VVS can be found, which achieves the minimum results of the cost function in (15), among the selected voltage vector sequences in a control cycle.

Therefore, from (15), it can be seen that when a voltage vector sequence \( V_t \) is applied, \( t_1, t_2, t_3 \) of \( V_t \) need to be calculated previously, then the results of the cost function can be calculated, and the optimal \( V_t \) can be solved.

In order to obtain the calculation expressions of the dwell times for a VVS, the cost function can be used for the optimal problem that calculating the vector dwell time values that minimize the cost function. Besides, the constraints of the vector dwell time should be considered, which are described as follows:

\[
\begin{cases}
0 \leq t_1 \leq T_s \\
0 \leq t_2 \leq T_s \\
0 \leq t_3 \leq T_s \\
t_3 = T_s - t_1 - t_2
\end{cases}. \quad (16)
\]

Combined (15) and (16), the optimal programming problem of the cost function can be formulated as

\[
\min g = G \left( t_1, t_2 \right) \quad t_1 \in R, t_2 \in R
\]

s.t. \( f_1 \left( t_1, t_2 \right) = t_1 \geq 0 \)

\( f_2 \left( t_1, t_2 \right) = t_2 \geq 0 \)

\( f_3 \left( t_1, t_2 \right) = T_s - t_1 - t_2 \geq 0 \).

\( f_1, f_2 \) and \( f_3 \) represent the constraint functions of vector dwell time. By solving the optimal programming problem of the cost function described in Equation (17), the vector dwell time for three voltage vectors of \( V_t \) can be obtained.

3.2 Solving for the optimal problem of cost function

The Hessian matrix determinant of the cost function is as follows:

\[
\det \left( H_G \right) = \begin{vmatrix}
\frac{\partial^2 G}{\partial t_1^2} & \frac{\partial^2 G}{\partial t_1 \partial t_2} \\
\frac{\partial^2 G}{\partial t_2 \partial t_1} & \frac{\partial^2 G}{\partial t_2^2}
\end{vmatrix}
\]

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\[= 4\lambda \left[ f_{x1} \left( f_{x2} - f_{x3} \right) - f_{x2} \left( f_{x1} - f_{x3} \right) \right]^2 + 4\lambda \left[ f_{x1} \left( f_{x2} - f_{x3} \right) - f_{x2} \left( f_{x1} - f_{x3} \right) \right]^2 + 4 \left[ f_{x1} \left( f_{x2} - f_{x3} \right) - f_{x2} \left( f_{x1} - f_{x3} \right) + f_{x3} \left( f_{x1} - f_{x2} \right) \right]^2 \]

where

\[
\frac{\partial^2 G}{\partial t_i^2} = 2 \left[ \lambda f_{x1}^2 + \left( f_{x1} - f_{x3} \right)^2 + \left( f_{x2} - f_{x3} \right)^2 \right].
\]

As shown in Equation (18), all principal minors of the Hessian matrix for the cost function are non-negative, so the Hessian matrix of the cost function is positive semi-definite, which indicates that the cost function is a convex function. And the optimal programming problem described in Equation (17) is a convex programming for its constraint functions \( f_1, f_2, f_3 \) are all concave functions. Therefore, the global optimal solution for the optimal programming problem of cost function with inequality constraints can be obtained by finding the solution which meet the K-T necessary conditions.

In this paper, the Lagrange multiplier method is preferred to find the global optimal solution for the optimal programming problem of cost function with inequality constraints. The Lagrange function is defined as

\[
L \left( t_1, t_2, w_1, w_2, w_3 \right) = G \left( t_1, t_2 \right) - w_1 f_1 \left( t_1, t_2 \right) - w_2 f_2 \left( t_1, t_2 \right) - w_3 f_3 \left( t_1, t_2 \right)
\]

where \( w_1, w_2 \) and \( w_3 \) are the Lagrange multipliers corresponding to the inequality constraint functions.

Besides, the first-order necessary conditions of the optimal programming problem are defined as

\[
\begin{align*}
\frac{\partial L \left( t_1, t_2, w_1, w_2, w_3 \right)}{\partial t_1} &= 0 \\
\frac{\partial L \left( t_1, t_2, w_1, w_2, w_3 \right)}{\partial t_2} &= 0 \\
w_1 f_1 \left( t_1, t_2 \right) &= 0 \\
w_2 f_2 \left( t_1, t_2 \right) &= 0 \\
w_3 f_3 \left( t_1, t_2 \right) &= 0 \\
t_1 &= 0 \\
t_2 &= 0 \\
T - t_1 - t_2 &= 0 \\
w_1, w_2, w_3 &= 0
\end{align*}
\]

As shown in Equation (20), there are equality and inequality constraints in the first-order necessary conditions of the optimal solution. In order to obtain the detailed calculation expression of the optimal solutions and facilitate the digital implementation of the algorithm, the system of equations in (20) is solved primarily to obtain the expression of all the alternative solutions, and then the alternative solutions are verified that whether they satisfy the inequality constraints in Equation (20) to obtain the optimal solution.

Define the alternative solutions as \( S_y \), \( y = 0, 1, 2, \ldots, 6 \). The expressions of all the seven alternative solutions are derived as follows:

1. **The expression of alternative solution \( S_0 \)** is

\[
\begin{align*}
t_1 &= \frac{(A_{x1} \cdot E_0 + B_{x1} \cdot F_0) \lambda + C_{x1} \cdot D_0}{(E_0^2 + F_0^2) \lambda + D_0^2} \\
t_2 &= \frac{(A_{x2} \cdot E_0 + B_{x2} \cdot F_0) \lambda + C_{x2} \cdot D_0}{(E_0^2 + F_0^2) \lambda + D_0^2} \\
w_1 &= 0, w_2 = 0, w_3 = 0
\end{align*}
\]

where

\[
\begin{align*}
A_{x1} &= \left( e^0_{x1} - T_{x1} f_{x3} \right) f_{x2} + \left( f_{x2} - f_{x3} \right) v_i \left( k \right) \\
B_{x1} &= \left( e^0_{x2} - T_{x2} f_{x3} \right) f_{x1} + \left( f_{x1} - f_{x3} \right) v_i \left( k \right) \\
C_{x1} &= \left( e^0_{x3} - T_{x3} f_{x3} \right) f_{x1} + \left( f_{x1} - f_{x3} \right) v_i \left( k \right) \\
A_{x2} &= \left( e^0_{x1} - T_{x1} f_{x3} \right) f_{x3} + \left( f_{x1} - f_{x3} \right) v_i \left( k \right) \\
B_{x2} &= \left( e^0_{x2} - T_{x2} f_{x3} \right) f_{x3} + \left( f_{x2} - f_{x3} \right) v_i \left( k \right) \\
C_{x2} &= \left( e^0_{x3} - T_{x3} f_{x3} \right) f_{x3} + \left( f_{x2} - f_{x3} \right) v_i \left( k \right)
\end{align*}
\]

In fact, in Equation (21), \( w_1, w_2 \) and \( w_3 \) are equal to 0, thereby, the expressions of \( t_1 \) and \( t_2 \) can also be calculated based on solving minimum value of the cost function in (17) without constraint functions \( f_1, f_2, f_3 \).

2. **The expression of alternative solution \( S_1 \)** is

\[
\begin{align*}
\begin{cases}
t_1 = 0, t_2 = 0 \\
w_1 = \frac{2}{\lambda} f_{x1}, w_2 = \frac{2}{\lambda} f_{x2}, w_3 = 0
\end{cases}
\end{align*}
\]

where

\[
\begin{align*}
A_1 &= \left( f_{x1} - f_{x3} \right) \left( T_{x1} f_{x3} - e^0_{x1} \right) + \left( f_{x1} - f_{x3} \right) \left( T_{x1} f_{x3} - e^0_{x2} \right) + \lambda f_{x1} v_i \left( k \right) \\
B_1 &= \left( f_{x2} - f_{x3} \right) \left( T_{x2} f_{x3} - e^0_{x1} \right) + \left( f_{x2} - f_{x3} \right) \left( T_{x1} f_{x3} - e^0_{x2} \right) + \lambda f_{x2} v_i \left( k \right).
\end{align*}
\]
3. The expression of alternative solution \( S_2 \) is

\[
\begin{align*}
    & t_1 = T_s, t_2 = 0 \\
    & w_1 = 0, w_2 = 2 A_2, w_3 = 2 B_2 \\
\end{align*}
\]

where

\[
A_2 = (f_{a1} - f_{a2}) (e_{\alpha}^0 - T_s f_{a1}) + (f_{\beta1} - f_{\beta2}) (e_{\beta}^0 - T_s f_{\beta1}) \\
- \lambda (f_{a1} - f_{a2}) (T_s f_{a1} + v_i (k)) \\
B_2 = (f_{a2} - f_{a3}) (e_{\alpha}^0 - T_s f_{a1}) + (f_{\beta1} - f_{\beta3}) (e_{\beta}^0 - T_s f_{\beta1}) \\
- \lambda f_{a1} (T_s f_{a1} + v_i (k)) \\
\]

(23)

4. The expression of alternative solution \( S_3 \) is

\[
\begin{align*}
    & t_1 = -\frac{A_{31} + \lambda f_{a1} v_i (k)}{E_3}, t_2 = 0 \\
    & w_1 = 0, w_2 = 2 \left[ \frac{\lambda (A_{32} + B_3) + C_3 \cdot D_3}{T_s E_3} \right], w_3 = 0 \\
\end{align*}
\]

where

\[
A_{31} = (f_{a1} - f_{a3}) (T_s f_{a3} - e_{\alpha}^0) + (f_{\beta1} - f_{\beta3}) (T_s f_{\beta3} - e_{\beta}^0) \\
A_{32} = [(T_s f_{a3} - e_{\alpha}^0) f_{a1} + (f_{a3} - f_{a1}) v_i (k)] \cdot [(f_{a2} - f_{a3}) f_{a1} \]
- \lambda (f_{a1} - f_{a3}) f_{a2} \\
B_3 = [(T_s f_{\beta3} - e_{\beta}^0) f_{a1} + (f_{\beta3} - f_{\beta1}) v_i (k)] \cdot [(f_{\beta2} - f_{\beta3}) f_{a1} \]
- \lambda (f_{\beta1} - f_{\beta3}) f_{\beta2} \\
C_3 = [(f_{\beta2} - f_{\beta3}) f_{a1} + (f_{\beta1} - f_{\beta2}) f_{a3} - (f_{\beta1} - f_{\beta3}) f_{\beta2}] \\
D_3 = [(T_s f_{\beta3} f_{a3} - f_{a3} f_{\beta3}) + (f_{\beta1} - f_{\beta3}) (e_{\beta}^0 - (f_{a1} - f_{a3}) e_{\beta}^0) \\
E_3 = \alpha^2 \alpha^2 + (f_{a1} - f_{a3})^2 + (f_{\beta1} - f_{\beta3})^2. \\
\]

(24)

5. The expression of alternative solution \( S_4 \) is

\[
\begin{align*}
    & t_1 = 0, t_2 = T_s \\
    & w_1 = 2 A_4, w_2 = 0, w_3 = -2 B_4 \\
\end{align*}
\]

where

\[
A_4 = \lambda (f_{a1} - f_{a3}) (f_{a3} T_s + v_i (k)) + (f_{a1} - f_{a2}) (T_s f_{a2} - e_{\alpha}^0) \\
+ (f_{\beta1} - f_{\beta2}) (T_s f_{\beta2} - e_{\beta}^0) \\
B_4 = (T_s f_{a3} + v_i (k)) \lambda f_{a2} + (f_{a3} - f_{a1}) (T_s f_{a2} - e_{\alpha}^0) \\
+ (f_{\beta2} - f_{\beta3}) (T_s f_{\beta2} - e_{\beta}^0). \\
\]

(25)

6. The expression of alternative solution \( S_5 \) is

\[
\begin{align*}
    & t_1 = 0, t_2 = -\frac{A_{51} + \lambda f_{a1} v_i (k)}{E_5}, t_2 = 0 \\
    & w_1 = -\frac{2 \lambda (A_{52} + B_5) + C_5 \cdot D_5}{T_s E_5}, w_2 = 0, w_3 = 0 \\
\end{align*}
\]

where

\[
A_{51} = (T_s f_{a3} - e_{\alpha}^0) (f_{a2} - f_{a3}) + (T_s f_{\beta3} - e_{\beta}^0) (f_{\beta2} - f_{\beta3}) \\
A_{52} = [(T_s f_{a3} - e_{\alpha}^0) f_{a2} - (f_{a2} - f_{a3}) v_i (k)] \cdot [(f_{a2} - f_{a3}) f_{a1} \\
\cdot f_{a1} - (f_{a1} - f_{a3}) f_{a2} \\
B_5 = [(T_s f_{\beta3} - e_{\beta}^0) f_{a2} - (f_{\beta2} - f_{\beta3}) v_i (k)] \cdot [(f_{\beta2} - f_{\beta3}) f_{a1} \]
\cdot f_{a1} - (f_{\beta1} - f_{\beta3}) f_{\beta2} \\
C_5 = [(f_{\beta3} - f_{\beta2}) f_{a1} + (f_{\beta2} - f_{\beta1}) f_{a3} - (f_{\beta1} - f_{\beta3}) f_{a2}] \\
D_5 = T_s (f_{a3} f_{a2} - f_{a2} f_{a3}) + (f_{\beta2} - f_{\beta3}) (e_{\beta}^0 - (f_{a1} - f_{a3}) e_{\beta}^0) \\
E_5 = \alpha^2 \alpha^2 + (f_{a2} - f_{a3})^2 + (f_{\beta2} - f_{\beta3})^2. \\
\]

(26)

7. The expression of alternative solution \( S_6 \) is

\[
\begin{align*}
    & t_1 = -\frac{A_{61}}{E_6}, t_2 = \frac{A_{62}}{E_6}, t_2 = 0 \\
    & w_1 = 0, w_2 = 0, w_3 = -\frac{2 \lambda (A_{63} + B_6) + C_6 \cdot D_6}{E_6}, w_3 = 0 \\
\end{align*}
\]

where

\[
A_{61} = (T_s f_{a2} - e_{\alpha}^0) (f_{a1} - f_{a2}) + (T_s f_{\beta2} - e_{\beta}^0) (f_{\beta1} - f_{\beta2}) \\
+ \lambda (f_{a1} - f_{a2}) (v_i (k) + T_s f_{a3}) \\
A_{62} = [(f_{a1} - f_{a3}) f_{a2} - (f_{a2} - f_{a3}) f_{a1}] \cdot [v_i (k) (f_{a1} - f_{a2}) \\
+ T_s (f_{a1} f_{a2} - f_{a2} f_{a1}) + e_{\alpha}^0 (f_{a1} - f_{a2})] \\
B_{61} = (T_s f_{a1} - e_{\alpha}^0) (f_{a1} - f_{a2}) + (T_s f_{a2} - e_{\beta}^0) (f_{\beta1} - f_{\beta2}) \\
+ \lambda (f_{a1} - f_{a2}) (v_i (k) + T_s f_{a3}) \\
B_{62} = [(f_{\beta1} - f_{\beta2}) f_{a1} - (f_{\beta2} - f_{\beta3}) f_{a3}] \cdot [v_i (k) (f_{\beta1} - f_{\beta2}) \\
+ T_s (f_{\beta1} f_{\beta2} - f_{\beta2} f_{\beta3}) + e_{\beta}^0 (f_{a1} - f_{a2})] \\
C_6 = (f_{\beta2} - f_{\beta3}) f_{a1} + (f_{\beta1} - f_{\beta2}) f_{a3} - (f_{\beta1} - f_{\beta3}) f_{a2} \\
D_6 = T_s (f_{a1} f_{a2} - f_{a2} f_{a1}) + (f_{\beta1} - f_{\beta2}) (e_{\alpha}^0 - (f_{a1} - f_{a2}) e_{\beta}^0) \\
E_6 = \alpha^2 \alpha^2 + (f_{a1} - f_{a2})^2 + (f_{\beta1} - f_{\beta2})^2. \\
\]

(27)

Above all, the expressions of all the alternative solutions \( S_0 - S_6 \) are derived by Lagrange multiplier method, according to the cost function with equality constraints in (20). Furthermore, it needs to be verified that whether \( t_1, t_2 \) and \( w_1, w_2, w_3 \) meet the inequality constraints in Equation (20) respectively. Actually, for the voltage vector sequences described in Section 2.3, the cost function in (20) is a strict convex function, which means the optimal solution for the optimal programming problem of cost function is unique.
Therefore, for each VVS, only one calculation expression is the optimal solution among the alternative solutions $S_0$–$S_5$, which meets equality and inequality constraints of the optimal programming problem in (20).

### 3.3 Further simplification of the solution verification

In order to simplify the verification of the optimal solution among the alternative solutions $S_0$–$S_5$, the amount of alternative solutions for the verification is reduced by region division. The detailed simplification process is described as follows.

As shown in Figure 3, take $t_1$–$t_2$ as the frame axis, the $t_1$–$t_2$ plane is divided into 7 regions (which are marked by Region ①–⑦) by the boundaries of inequality constraints in (16), and the alternative solutions $S_0$–$S_5$ that located in Region ② meet the inequality constraints in (16). Actually, as it is described above, the solution $S_t$ in equation (21) meets the minimum value $g_{t0}$ of the cost function in (17) without constraint functions $f_1$, $f_2$, $f_3$.

According to the convex property of the cost function, in the $t_1$–$t_2$ plane, the cost function values of the points, which are closer to $S_t$ in the same direction pointing to $S_{0t}$, are smaller [26]. When a voltage vector sequence $V_t$ is applied, supposing that the corresponding $S_t$ is at Region ② in the $t_1$–$t_2$ plane, as shown in Figure 3, the cost function value of point $P_t$ is smaller than the cost function value of $S_{0t}$. The similar conclusions can be reached at points $P_2$ and $P_3$, and the possible optimal solution set for the optimal programming problem in (20) is $\{S_1, S_4, S_5\}$. Furthermore, $S_1$, $S_4$, $S_5$ can be calculated by (22), (25) and (26) respectively, and the optimal solution would be the one in $\{S_1, S_4, S_5\}$, which meets the inequality constraints in Equation (20).

In summary, based on the region division, according to the location of $S_t$ in the $t_1$–$t_2$ plane, the possible optimal solution set is preselected and the amount of alternative solutions for the verification process is reduced. The detailed possible optimal solution set are listed in Table 2, which primarily depends on the region that $S_t$ falls in.

### 4 VECTOR SEQUENCES PRESELECTION

According to the optimization principle of traditional model predictive control and the arrangements of voltage vectors sequence described in Section 2.3, there are 96 voltage vectors sequences need to be calculated for the optimization problem for the grid currents and the neutral-point voltage in each sampling cycle, which would bring great calculation burden for the signal processor. Therefore, it is preferred to reduce the amount of voltage vector sequences that need to be calculated in each sampling cycle.

The reference voltage vector $v^\ast(k)$ can be introduced to reduce the amount of vector sequences. The reference voltage vector $v^\ast(k)$ can be calculated by

$$
v^\ast(k) = \begin{bmatrix} u^\ast_x(k) \\ u^\ast_y(k) \\ u^\ast_\beta(k) \end{bmatrix} = \frac{L}{\tau} \begin{bmatrix} \ell^x_x(k+1) \\ \ell^y_x(k+1) \end{bmatrix} + \begin{bmatrix} \ell^x_x(k) \\ \ell^y_x(k) \end{bmatrix} + \frac{1}{\tau} \begin{bmatrix} \ell^x_\beta(k) \\ \ell^y_\beta(k) \end{bmatrix},
$$

(28)

Determined by the position of $v^\ast(k)$ in the distribution diagram of output voltage vectors, only one sector can be selected from sector A–F, instead of calculating all the vector sequences in six sectors totally, and then the online computation is greatly reduced. In the subsequent optimization calculation process, only the voltage vector sequences in the preselected sector would be calculated, so that the amount of voltage vector sequences that needs to be calculated can be reduced from 96 to 12 in each sampling cycle.

Moreover, the redundant positive or negative small vectors in a pair have the same effects on the grid-connected currents and opposite effects on the neutral-point voltage, and from Table 1, it can be seen that each VVS includes the small vectors. Thereby, according to the sign of the neutral-point voltage and grid currents, only the voltage vector sequences that have better effect on controlling the neutral-point voltage will be
### TABLE 3 Parameters for experimental test bench

| Parameter                  | Symbol | Value  |
|----------------------------|--------|--------|
| Parasitic resistance       | \( R_a, R_b, R_c \) | 0.5 \( \Omega \) |
| Filter inductance          | \( L_a, L_b, L_c \) | 5 mH |
| DC−link capacitance        | \( C_1, C_2 \) | 150 \( \mu F \) |
| DC−link bus voltage        | \( V_d \) | 240 V |
| Amplitude of grid phase voltage | \( e_{a, peak}, e_{b, peak}, e_{c, peak} \) | 100 V |
| Sampling frequency         | \( f_s \) | 10 kHz |
| Weight coefficient         | \( \lambda \) | 0.05 |

The parameters are preselected. Besides, the preselected voltage vector sequences generate a voltage vector sequences set for selecting the final optimal output VVS online. Then, the amount of voltage vector sequences can be decreased. The detailed analysis is as follows.

Taking the redundant vectors \( v_3(0 -1 -1) \) and \( v_4(1 0 0) \) as example, when \( v_3 \) is applied, according to (8), the increment \( f_{vc}(v_3) \) of the neutral-point voltage is

\[
f_{vc}(v_3) = -\frac{1}{C} \left( |0| - \frac{1}{2} \times |1| - \frac{1}{2} \times |1| \right) i_\alpha (k) = \frac{i_\alpha (k)}{C}.
\]

By using \( v_4 \), the \( f_{vc}(v_4) \) can obtained as

\[
f_{vc}(v_4) = -\frac{1}{C} \left( |1| - \frac{1}{2} \times |0| - \frac{1}{2} \times |0| \right) i_\alpha (k) = -\frac{i_\alpha (k)}{C}.
\]

Supposing that \( i_\alpha (k) \geq 0 \) and \( v_{\alpha}(k) \geq 0 \), then \( f_{vc}(v_3) \geq 0 \) and \( f_{vc}(v_4) \leq 0 \), it is better to choose the voltage vector sequences that including \( v_4 \) instead of \( v_3 \). The vector sequence \{\( v_{15} \ v_4 \ v_{21} \)\} is thereby preselected between two vector sequences in \( \triangle 43 \) triangle area of Figure 2, and the preselection principle can be applied to all the sectors. Therefore, the amount of voltage vector sequence that need to be calculated online can be decreased from 12 to 5 finally, by judging the sign of the neutral-point voltage and grid currents in a control cycle. The flowchart of the proposed OSS-MPC strategy is shown in Figure 4.

### 5 EXPERIMENTAL RESULTS

In order to verify the validity of the proposed OSS-MPC method, a three-level NPC inverter prototype has been built. The parameters of the experimental test bench are listed in Table 3.

In the experimental test bench, as shown in Figure 5, a control system consists of a TMS320F28377D DSP and a EPM1270T CPLD. The DSP is designed for the implementation of control algorithm mainly, and the sampling of grid phase voltages, grid phase currents as well as the capacitor voltage.
voltage are also completed by the DSP. Besides, the CPLD is applied to arrange digital signals and realize protections for power switches.

5.1 Experimental results under grid-connected condition

According to experimental parameters under grid-connected conditions in Table 3, the experimental results are shown in Figure 6. From Figure 6a, it can be seen that the proposed OSS-MPC method has good performances on the control of grid currents, and the THD (up to 100th harmonic order) of grid currents is very low, which is about 2.421%. Figure 6b shows that proposed OSS-MPC method can also control the neutral-point voltage $v_n$ effectively. $v_n$ fluctuates less than 5V without a DC deviation. From the fast Fourier transform (FFT) analysis result of the grid current of phase A in Figure 6c, it can be seen that the frequency spectrum of the grid current is clustered at the sampling frequency $f_s$ (harmonic order 200), which indicates that the switching frequency of the grid current for the inverter is almost fixed.

In order to show the dynamic performance of the proposed OSS-MPC method, the transient experimental waveforms under grid-connected conditions are shown in Figure 7. In Figure 7a, when the peak value of grid current reference steps from 9 to 4.5 A, the proposed method performs good dynamic and stead-state characteristics on the control of grid currents. Besides, Figure 7b shows that the proposed method can also response quickly and maintain a good control effect on the neutral-point voltage.

5.2 Experimental results under different modulation ratio conditions

In order to further show the performance of the proposed method, the ECS-MPC method in [19] and conventional FCS-MPC method in [5] are compared with the proposed method in terms of output current quality, the neutral-point voltage control and system efficiency. The experiment at different modulation ratio ($M$) is performed on a NPC three-level inverter prototype with a three-phase balanced RL load ($R_{\text{load}} = 10 \, \Omega$ and $L_{\text{load}} = 5 \, \text{mH}$).

In the experiment, $M$ is adjusted by changing the amplitude of the reference ac current, the experimental parameters are set based on Table 3. The experimental results with
the three control methods at $M = 0.7$, $0.5$, $0.3$ and $0.1$ are shown in Figures 8–11, respectively, which are compared in Table 4. In Figure 8a, Figure 9a, Figure 10a and Figure 11a, the THD (up to 100th harmonic order) of the load current with proposed OSS-MPC method is very low at different $M$. It also can be seen that with the proposed method, the frequency spectrums of load currents are clustered in the domain of the high frequency around $f_s$ under different $M$ conditions.

From Figures 8–11 and Table 4, it can be seen that compared with the other two methods in [5] and [19], the proposed OSS-MPC method can concentrate the frequency spectrum and improve the quality of output currents effectively with lower THD value. Furthermore, the proposed OSS-MPC method has a better control effect for the neutral-point voltage with smaller ripples under varies modulation ratio conditions.

Meanwhile, the power losses of inverter with proposed OSS-MPC method, and the other two control methods in [5] and [19] are analyzed and experimentally verified in this section.

Commonly, the efficiency of the PWM converter mainly depend on the conduction loss and the switching loss of power devices, which is affected by PWM generations using different methods [27,28]. In this paper, supposing that the neutral-point voltage is balanced, the calculation for the conduction loss and the switching loss of power devices is analyzed as follows:
FIGURE 10  Experimental waveforms of load currents and the corresponding frequency spectrums with three methods at $M = 0.3$. (a) Proposed OSS-MPC ($M = 0.3$). (b) ECS-MPC ($M = 0.3$). (c) Conventional FCS-MPC ($M = 0.3$)

FIGURE 11  Experimental waveforms of load currents and the corresponding frequency spectrums with three methods at $M = 0.1$. (a) Proposed OSS-MPC ($M = 0.1$). (b) ECS-MPC ($M = 0.1$). (c) Conventional FCS-MPC ($M = 0.1$)

TABLE 4  Experimental results comparison of the three methods

| Modulation ratio | The ripple of $v_c$ | THD value of output currents |
|------------------|----------------------|-----------------------------|
|                  | Proposed OSS-MPC | ECS-MPC in [19] | Conventional FCS-MPC in [5] | Proposed OSS-MPC | ECS-MPC in [19] | Conventional FCS-MPC in [5] |
| $M = 0.7$        | 10.0 V            | 14.0 V            | 24.0 V            | 2.501%           | 2.783%           | 5.839%           |
| $M = 0.5$        | 6.5 V             | 9.5 V             | 15.5 V            | 1.900%           | 2.612%           | 7.081%           |
| $M = 0.3$        | 4.5 V             | 7.5 V             | 7.5 V             | 2.175%           | 3.487%           | 12.341%          |
| $M = 0.1$        | 5.5 V             | 6.0 V             | 9.0 V             | 3.374%           | 4.260%           | 46.705%          |
The conduction losses of power devices are calculated as [29]

$$P_{\text{cond}(\text{IGBT})} = \frac{1}{T} \int_0^T V_{CE} \cdot I_C \cdot \text{duty}_{\text{IGBT}} \, dt \quad (31)$$

$$P_{\text{cond}(\text{diode})} = \frac{1}{T} \int_0^T V_d \cdot I_F \cdot \text{duty}_{\text{diode}} \, dt \quad (32)$$

where $T$ is the period of the fundamental frequency; $V_{CE}$ is IGBT saturation voltage; $I_c$ is IGBT collector current; duty$_{\text{IGBT}}$ is the duty ratio of IGBT, which is obtained from modulation signals generated by the control methods; $V_d$ is diode voltage drop; $I_F$ is diode forward current; duty$_{\text{diode}}$ is the duty ratio of diode.

The switching losses of power devices are obtained by [29]

$$P_{sw}(\text{IGBT}) = \frac{1}{T} \sum_{t=0}^{T} \left[ E_{\text{on}}(V_{ce}, I_c) + E_{\text{off}}(V_{ce}, I_c) \right] \quad (33)$$

$$P_{sw}(\text{diode}) = \frac{1}{T} \sum_{t=0}^{T} E_{\text{on}}(V_d, I_F) \quad (34)$$

where $E_{\text{on}}$ is turn-on energy loss, $E_{\text{off}}$ is turn-off energy loss and $E_{\text{on}}(V_d)$ is the reverse recovery energy loss.

According to Equations (31)-(34), the calculated total power losses of the inverter with three methods are shown in Figure 12, where $M = 0.7$ and root-mean-square (RMS) value of phase currents $I_{x_{\text{rms}}} \ (x = a, b, c) = 6.76 \, A$.

From Figure 12, it can be seen that compared with conventional FCS-MPC method in [5] and ECS-MPC method in [19], the total power loss of power devices with proposed OSS-MPC method is higher.

The experimental results for the inverter efficiency with three different methods above are shown in Figure 13, where $\eta$ represents the output efficiency.

From Figure 13, it can be seen that compared with conventional FCS-MPC and ECS-MPC methods, although the inverter efficiency is a bit sacrificed with the proposed OSS-MPC method. However, the proposed method can also maintain a high system efficiency at high modulation ratios ($M = 0.5, 0.7$).

Above all, compared with conventional FCS-MPC and ECS-MPC method, although the efficiency is a bit sacrificed, especially at low modulation ratio, the proposed OSS-MPC method has advantages in improving the output current quality and controlling the neutral-point voltage with a smaller ripple at various modulation ratios. Besides, the proposed method can concentrate the frequency spectrum of output currents, which is beneficial to the filter design for the inverter.

6 | CONCLUSIONS

In order to improve the quality of grid currents and concentrate the frequency spectrum of the output leg voltage for the three-level NPC inverter, an OSS-MPC algorithm is proposed in this paper. Based on the increments of grid current and neutral-point voltage, the predictive model and cost function for the proposed OSS method are derived. After that, a method to calculate the vector dwell time by using Lagrange multiplier method to solve the cost function optimal problem with inequality constraints is proposed, and the calculation process is simplified to reduce the computation. Besides, a vector sequences preselection method is introduced to further reduce the online computational burden. Experimental results show that compared with conventional FCS-MPC and ECS-MPC method, the proposed OSS-MPC method has better performances on improving the output current quality and controlling the neutral-point voltage. And the frequency spectrum of the inverter output voltage with proposed method becomes concentrated, which is beneficial for decreasing the current distortion and designing the filtering inductor of inverter.

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