

Article

Model-Free Predictive Synthesis Performance Optimization of DAB Converters Based on an Ultra-Local Model

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Abstract

The dual-active-bridge (DAB) converter is the core component of the DC micro-grid system; it has the advantages of topological structure symmetry, high efficiency, and high-power density. Model predictive control (MPC) is often employed to improve the dynamic response characteristics of the system, but its strong parameter dependence is a key factor limiting the development of MPC. Therefore, a model-free predictive control (MFPC) method combining an ultra-local model with model predictive control is proposed to solve the problem of strong dependence of traditional MPC on system model parameters. Firstly, establish the ultra-local mathematical model of the DAB converter. The system's lumped disturbances are identified using the residual prediction method and substituted into the discrete model of the system at the next time step to achieve model-free prediction. Secondly, a minimum back-flow power constraint is added to the cost function to improve the steady-state performance of the converter. Thirdly, in the extended phase shift modulation, the Lagrange multiplier method (LMM) is proposed to reduce the current stress, ultimately achieving the collaborative optimization of the comprehensive performance of the DAB. Finally, a simulation model is built using MATLAB/Simulink, and compared with traditional control methods, the voltage ripple has been reduced by 51.3%, 89.1%, and 85.1%, respectively; the current stress significantly decreases both when the output voltage reference value changes and when the load resistance changes abruptly, and both can basically achieve zero back-flow power operation. The validity and superiority of the proposed strategy have been verified.

Keywords: DAB converter; model-free predictive control; ultra-local model; LMM; steady-state performance; dynamic performance



Academic Editor: Krzysztof Górecki

Received: 10 April 2026

Revised: 15 May 2026

Accepted: 15 May 2026

Published: 18 May 2026

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1. Introduction

With the rapid development of renewable energy, the scale and application range of DC micro-grids have also been constantly expanding [1]. To achieve flexible configuration and optimal utilization of renewable energy in DC micro-grids, the application of DC-DC converters has been promoted, providing support and guarantee for the development of DC micro-grids [2]. The current topologies of bidirectional DC-DC converters can be roughly classified into two categories: isolated and non-isolated bidirectional DC-DC converters. In the isolated bidirectional DC-DC converter, the dual-active-bridge (DAB) converter has attracted widespread attention owing to its capabilities—including bidirectional power

transfer, a wide voltage conversion ratio, and zero-voltage soft switching [3]. When the DAB converter is applied to DC micro-grids, it is necessary to effectively deal with operating conditions such as sudden load changes and sudden voltage changes. Therefore, the robustness and dynamic performance of the control system constitute critical metrics for achieving efficient control [4].

The steady-state optimization objectives of the DAB converter include inductor current, output voltage, system loss [5], and return power [6]. By modulating the DAB converter, the steady-state performance can be optimized [7]. In addition to the steady-state characteristics, actual systems may encounter working conditions with sudden changes in source and load, causing complex transient characteristics in the DAB converter [8]. This shows that dynamic performance is crucial for the stable operation of the system. The control of the DAB converter can optimize the transient performance of the system, with control targets including input/output voltage, inductor current, etc. Under traditional proportional integral (PI) control, the control system is limited by the integral link, resulting in a slow dynamic response speed, which severely restricts the application of the DAB converter in the field of new energy [9,10]. In recent years, model predictive control (MPC) has emerged as a nonlinear control strategy. It can improve the transient response speed in power converters while maintaining robustness and can well meet the requirements of power fluctuations and impact loads in DC distribution systems. With the rapid advancement of high-performance DSP, model predictive control (MPC) has been widely adopted in applications such as power electronic converters and motor control systems [11]. Since MPC predicts the future behavior of the controlled system based on the prediction model of the system, passive device parameters, dead time, high-frequency variable reduced-order models, and sampling noise can all lead to static errors in MPC, reducing the performance of the converter. Therefore, the strong dependence on parameters is one of the important reasons restricting the development of MPC [12]. To improve the parameter robustness of MPC, reference [13] proposed an MPC scheme with a power balance function, suitable for input-independent output-parallel DAB systems. However, MPC still needs to use PI control to compensate for model errors, thus sacrificing dynamic performance. In reference [14], researchers used the recursive least squares method to identify DAB inductor parameters online, improving the parameter robustness of MPC. However, before implementing the above parameter identification methods, it is still necessary to establish a more accurate system model to obtain the relationship between system identification variables and input/output variables.

Reference [15] proposed the concept of model-free control (MFC). This theory uses an ultra-local model to replace the model of traditional control systems. Only the determined variables in the system are placed in the model assumptions, while other variables can be calculated through lumped disturbances. This method can significantly reduce the impact of parameter accuracy on the system. Therefore, an MFC based on the ultra-local model was proposed to control the controlled object [16]. This lumped disturbance method can also be applied to DC converters to overcome the dependence of MPC on parameters. Reference [17] proposed a model-free predictive power control, which effectively enhanced the robustness of the system. Compared with traditional power predictive control, this method has a higher power steady-state ripple, mainly due to significant errors in the estimation of the dynamic part in the ultra-local model. Reference [18] proposed the ultra-local control of DAB converters and used a sliding mode observer to estimate and compensate for the dynamic part in the ultra-local model, improving the stability and robustness of the converter. However, the designed hyper-local controller essentially still adopts PI control, which is difficult to meet the control requirements of converters under nonlinear and time-varying conditions.

In response to the above problems, a collaborative optimization method combining Model-Free Predictive Control (MFPC) with the LMM method is proposed in this paper. Firstly, the operating principle and basic working characteristics of the DAB converter under EPS modulation are analyzed; then, a large-signal reduced-order model of the DAB converter is established, and the output voltage prediction equation of the system is constructed. Finally, the innovative contributions of the strategy in this paper are summarized as follows:

- (1) Based on the residual prediction method, the centralized disturbance introduced in the next time step is calculated in the model to implement an ultra-local model. Compared with the traditional MPC model, it has the advantages of parameter insensitivity and self-balancing of transmission power, achieving model-free predictive voltage control.
- (2) To reduce the back-flow power, the minimum back-flow power constraint is added using the cost function to calculate the optimal shift ratio combination (D_1 , D_2), thereby enhancing the performance of the DAB.
- (3) The two modulation modes of EPS are analyzed, and the LMM optimization method is adopted to reduce the inductor current stress. The optimal inner and outer shift ratio combinations (D_1 , D_2) within the full power range and the corresponding maximum current stress under both conditions are solved.

From what has been discussed above, the combination of the external shift ratio obtained by solving the cost function with back-flow power constraints and the internal shift ratio obtained by the LMM method is jointly applied to the EPS driver module of the DAB converter to reduce the back-flow power and current stress of the system and achieve the comprehensive performance optimization of the DAB converter. Compared with traditional methods, the proposed control method has the advantages of strong robustness, fast dynamic response, and low computational complexity.

2. Topological Structure and EPS Modulation of the DAB Converter

2.1. Topological Structure

The typical circuit structure of the DAB converter is shown in Figure 1. The H-bridges on both sides are connected through the intermediate high-frequency transformer. U_i and U_o represent the DC voltage sources on both the left and right sides of DAB, n is the transformation ratio of the transformer, L_d is the sum of the leakage inductance of the transformer and the external series inductor converted to the primary side, and C_i and C_o are the DC-side supporting capacitors.

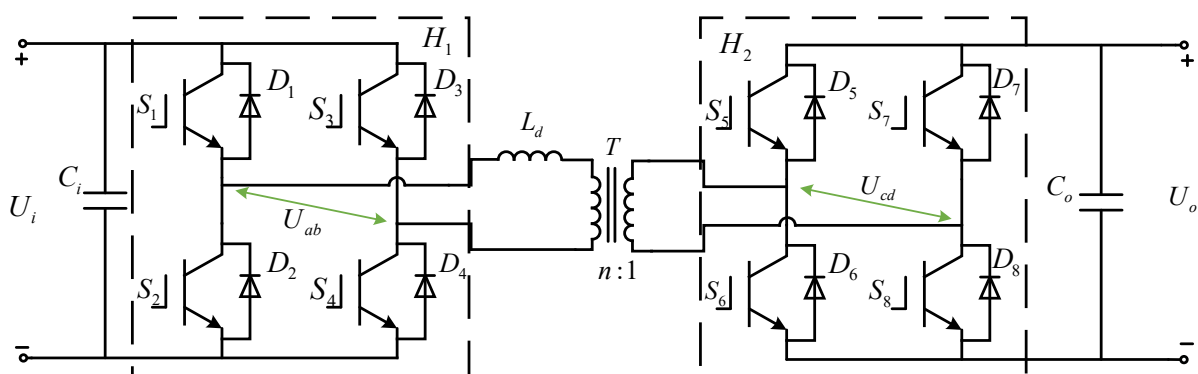


Figure 1. DAB converter topology structure.

2.2. EPS Modulation

EPS modulation, as an improved method of SPS modulation, as shown in Figure 2, extends the phase shift control compared to the traditional single phase shift control. It

increases the phase shift angle within the H-bridge on the i_{L_d} side, which doubles the control degrees of freedom and thereby enhances the control flexibility. Here, the internal phase shift time is represented as $D_1 T_{hs}$, and the external phase shift time is represented as $D_2 T_{hs}$.

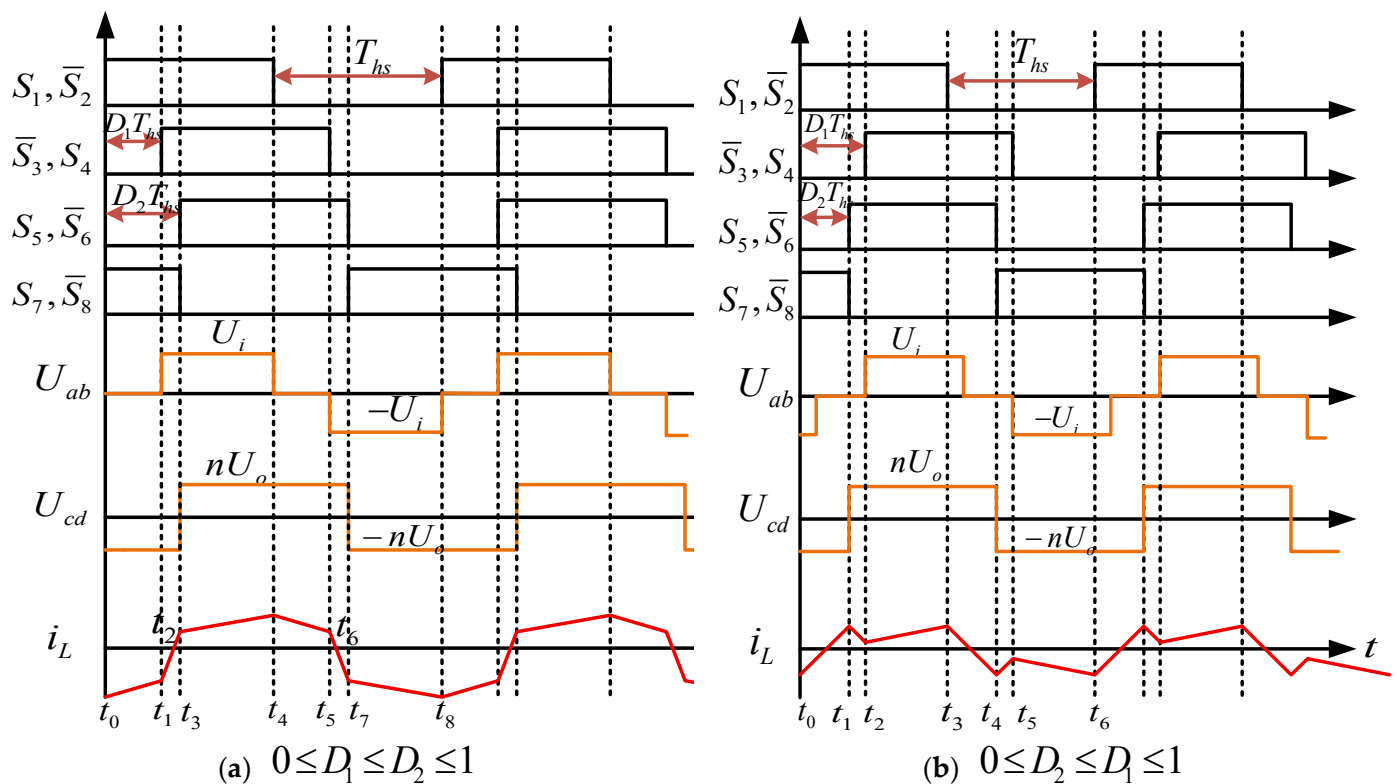


Figure 2. Two modes of EPS modulation.

According to the working waveform diagram in Figure 2a, it can be seen that in each switching cycle, the working process of the DAB converter is divided into eight states, as shown in Figure 3 below. The analysis method of Figure 2b is similar to that of Figure 2a and will not be elaborated further in the following text.

$$i_{L_d}(t) = \begin{cases} i_{L_d}(t_0) + \frac{nU_o}{L_d}(t - t_0), & t_0 \leq t \leq t_1 \\ i_{L_d}(t_1) + \frac{U_i + nU_o}{L_d}(t - t_1), & t_1 \leq t \leq t_2 \\ i_{L_d}(t_2) + \frac{U_i + nU_o}{L_d}(t - t_2), & t_2 \leq t \leq t_3 \\ i_{L_d}(t_3) + \frac{U_i - nU_o}{L_d}(t - t_3), & t_3 \leq t \leq t_4 \\ i_{L_d}(t_4) + \frac{-nU_o}{L_d}(t - t_4), & t_4 \leq t \leq t_5 \\ i_{L_d}(t_5) + \frac{-U_i - nU_o}{L_d}(t - t_5), & t_5 \leq t \leq t_6 \\ i_{L_d}(t_6) + \frac{-U_i - nU_o}{L_d}(t - t_6), & t_6 \leq t \leq t_7 \\ i_{L_d}(t_7) + \frac{-U_i + nU_o}{L_d}(t - t_7), & t_7 \leq t \leq t_8 \end{cases} \quad (1)$$

In Figure 2a, let $t_0 = 0$; then, at each moment, they can be expressed as: $t_1 = D_1 T_{hs}$, $t_3 = D_2 T_{hs}$, $t_4 = T_{hs}$, $t_5 = (1 + D_1) T_{hs}$, $t_7 = (1 + D_2) T_{hs}$, and $t_8 = 2T_{hs}$. When the circuit stabilizes, based on the symmetry of the inductor current, we can obtain $i_{L_d}(t_0) = -i_{L_d}(t_3)$ and $i_{L_d}(t_2) = -i_{L_d}(t_5)$, and combined with Equation (1), the expression of the current in the first half cycle is:

$$\begin{cases} i_{L_d}(t_0) = -\frac{nU_o}{4f_sL_d}[k(1-D_1)+2D_2-1] \\ i_{L_d}(t_1) = -\frac{nU_o}{4f_sL_d}[k(1-D_1)+2D_2-2D_1-1] \\ i_{L_d}(t_3) = \frac{nU_o}{4f_sL_d}[k(2D_2-1-D_1)+1] \\ i_{L_d}(t_4) = \frac{nU_o}{4f_sL_d}[k(1-D_1)+2D_2-1] \end{cases} \quad (2)$$

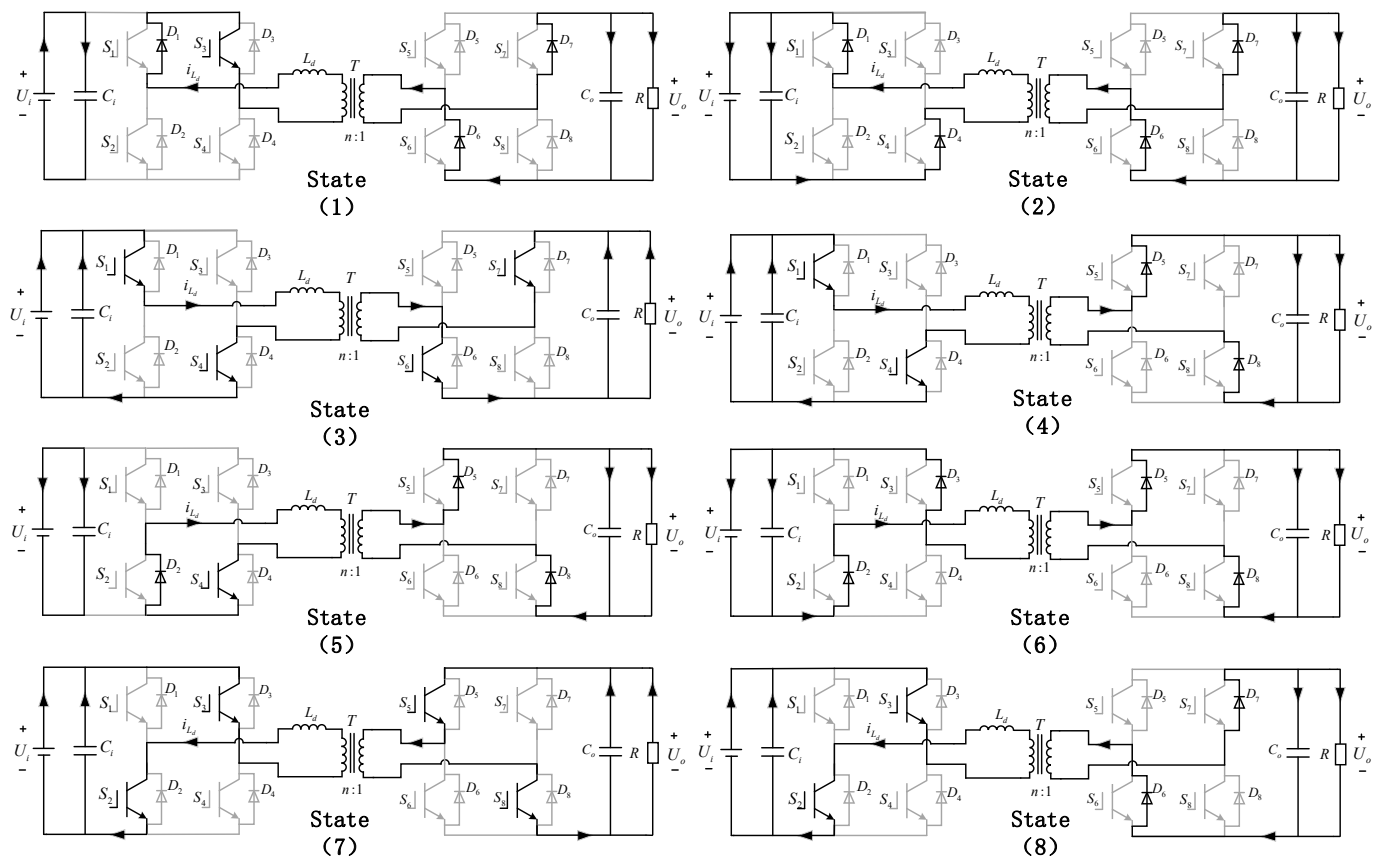


Figure 3. The eight states of $0 \leq D_1 \leq D_2 \leq 1$.

Using the same analytical method, the current expressions for each component in the first half cycle of the pattern shown in Figure 2b can be obtained as follows:

$$\begin{cases} i_{L_d}(t_0) = -\frac{nU_o}{4f_sL_d}[k(1-D_1)+2D_2-1] \\ i_{L_d}(t_1) = \frac{nU_o}{4f_sL_d}(kD_1+1-k) \\ i_{L_d}(t_3) = \frac{nU_o}{4f_sL_d}[k(D_1-1)+2D_2-2D_1+1] \\ i_{L_d}(t_4) = \frac{nU_o}{4f_sL_d}[k(1-D_1)+2D_2-1] \end{cases} \quad (3)$$

Based on Equations (2) and (3), the average input current for both modes is obtained as:

$$I_{L_d} = \begin{cases} \frac{nU_o}{4f_sL_d}(-D_1^2+2D_1D_2-D_1-2D_2^2+2D_2), & 0 \leq D_1 \leq D_2 \leq 1 \\ \frac{nU_o}{4f_sL_d}(D_1^2-2D_1D_2-D_1+2D_2), & 0 \leq D_2 \leq D_1 \leq 1 \end{cases} \quad (4)$$

The corresponding transmission power is:

$$P_{EPS} = \begin{cases} \frac{nU_iU_o}{4f_sL_d}(-D_1^2+2D_1D_2-D_1-2D_2^2+2D_2), & 0 \leq D_1 \leq D_2 \leq 1 \\ \frac{nU_iU_o}{4f_sL_d}(D_1^2-2D_1D_2-D_1+2D_2), & 0 \leq D_2 \leq D_1 \leq 1 \end{cases} \quad (5)$$

Select the maximum transmission power P_N under SPS control and perform normalization:

$$P_{EPS.N} = \begin{cases} 2(-D_1^2 + 2D_1D_2 - D_1 - 2D_2^2 + 2D_2), & 0 \leq D_1 \leq D_2 \leq 1 \\ 2(D_1^2 - 2D_1D_2 - D_1 + 2D_2), & 0 \leq D_2 \leq D_1 \leq 1 \end{cases} \quad (6)$$

3. Model-Free Predictive Control (MFPC) of DAB

3.1. The Mathematical Model of DAB

Based on the state space averaging method [19,20], all variables in the state space model are replaced by their average values. This paper establishes a linearized large-signal reduced-order model for the DAB converter. The dynamic characteristics of the inductor current in the converter are mainly ignored, and this omission has a relatively small impact on the control accuracy of the system when controlling the output voltage [21]. Through the above analysis, the large-signal reduced-order model of the converter is shown in Figure 4.

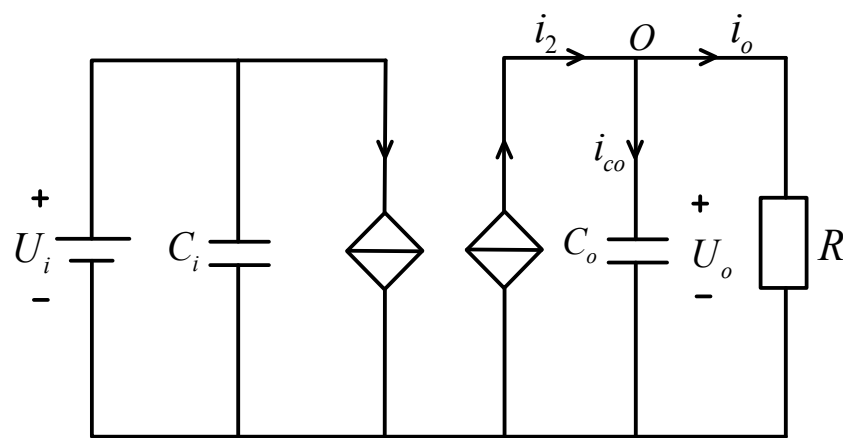


Figure 4. The average value equivalent circuit model of the DAB converter.

According to Kirchhoff's Current Law (KCL), the current equation at point O can be expressed as:

$$i_2 = i_{co} + i_o \quad (7)$$

The secondary side current transmission can be calculated based on the transformer ratio n without considering the characteristics of the components themselves. The average value of the secondary side current controlled by EPS can be obtained by calculating the transmission power according to Equation (5):

$$i_2 = ni_L = n \frac{P}{U_{ab}} = \begin{cases} \frac{nT_s U_i (2D_2(1-D_2+D_1) - D_1 - D_1^2)}{4L_d}, & 0 \leq D_1 \leq D_2 \leq 1 \\ \frac{nT_s U_i (D_1(D_1-1-2D_2) + 2D_2)}{4L_d}, & 0 \leq D_2 \leq D_1 \leq 1 \end{cases} \quad (8)$$

3.2. Ultra-Local Model

The design of the model-free predictive control method under DAB converter EPS modulation is similar to the previous content. By combining Equation (5), the output-side current i_2 under EPS modulation can be obtained as:

$$i_2 = \frac{P_{EPS}}{U_o} = \frac{nT_s U_i}{2L_d} D_{EPS} \quad (9)$$

In Equation (9):

$$D_{EPS} = \begin{cases} \frac{(2D_2(1-D_2)-D_1(1+D_1)+2D_2D_1)}{2}, & 0 \leq D_1 \leq D_2 \leq 1 \\ \frac{(2D_2(1-D_1)+D_1(D_1-1))}{2}, & 0 \leq D_2 \leq D_1 \leq 1 \end{cases} \quad (10)$$

Based on the average equivalent circuit of the DAB, the output voltage state equation can be obtained as follows:

$$\frac{dU_o}{dt} = \frac{nT_s U_i}{2L_d C_o} D_{EPS} - \frac{U_o}{RC_o} \quad (11)$$

The forward Euler method is adopted to discretize Equation (16) and simplify it accordingly. Then, the voltage prediction model of the DAB converter at time $k + 1$ after modulation by EPS can be obtained as:

$$U_o(k+1) = (1 - \frac{T_s}{RC_o})U_o(k) + \frac{nT_s^2 U_i(k)}{2L_d C_o} D_{EPS}(k) \quad (12)$$

Therefore, by combining Equation (16), the ultra-local model of the DAB system can be obtained as:

$$\begin{cases} \dot{x} = \alpha u + \hat{f} \\ y = x \end{cases} \quad (13)$$

Here, $nT_s^2 U_i(k)/2L_d C_o$ is denoted as α , representing the controller gain of the system; D_{EPS} is denoted as u , representing the control force input of the EPS system; and $\frac{dU_o}{dt}$ is denoted as $\dot{x}$. Thus, the ultra-local model in Equation (13) can be further derived as:

$$\begin{cases} \frac{dU_o}{dt} = \alpha D_{EPS} + \hat{f} \\ \hat{f} = -\frac{U_o}{RC_o} + f \end{cases} \quad (14)$$

Among them, f and $\hat{f}$ represent the unmodeled unknown disturbance terms and the lumped disturbance terms under EPS modulation, respectively. By discretizing (14) using the forward Euler method, the discrete output voltage prediction ultra-local model of the DAB converter at time $k + 1$ under EPS modulation can be obtained as:

$$U_o(k+1) = U_o(k) + T_s(\alpha(k)D_{EPS}(k) + \hat{f}(k)) \quad (15)$$

To enhance the robustness and tracking performance of the MPC, this paper employs the “model residual + first-order prediction (EWMA)” method to obtain the collective disturbances. By rewriting the model at time $f_{lu}(k)$ in the previous part, the collective disturbance of the system can be expressed as:

$$f_{lu}(k-1) = \frac{U_o(k) - U_o(k-1)}{T_s} - \alpha(k-1)D(k-1) \quad (16)$$

Use first-order exponential weighting to obtain the predicted value for the next time step:

$$\hat{f}_{lu}(k) = (1 - \alpha)\hat{f}_{lu}(k-1) + \alpha f_{lu}(k-1) \quad (17)$$

The first-order prediction method combines the “newly measured residuals” and the “previous prediction value” according to weights to obtain the next prediction value, which is the predicted value of the system’s lumped disturbance $\hat{f}_{lu}(k)$, and it is used to replace the actual value $f_{lu}(k)$ and is directly substituted into the local model of the next time step. Among them, α represents the residual prediction coefficient (also known as the forgetting factor or smoothing factor), which is used to balance the weight of the current measured

residual and the previous prediction value in the prediction of the next step. Based on experience, α is usually taken to be between 0.1 and 0.5.

3.3. Parameter Sensitivity Analysis

In Equation (12), it can be concluded that the model parameter errors of the energy storage inductor and the output capacitor in the circuit directly affect the accuracy of the prediction model. $U_o(k+1)$ represents the output voltage of the controller at time $k+1$, and U_o^* is the reference value of the output voltage. L_d and C_o are the model values of the energy storage inductor and the output capacitor in the controller. Define $U_o^{re}(k+1)$ as the output voltage of the actual circuit at time $k+1$, and L_{dm} and C_{om} are the actual circuit model values of the energy storage inductor and the output capacitor, considering the errors. Assuming that the output voltage value of the controller at time k is the actual value, then the output voltage of the controller at time $k+1$ is as shown in Equation (12), while the output voltage of the actual circuit at time $k+1$ is:

$$U_o^{re}(k+1) = (1 - \frac{T_s}{RC_o})U_o(k) + \frac{nT_s^2 U_i(k)}{2L_{dm}C_{om}}D_{EPS}(k) \quad (18)$$

When the system is operating stably, it can be considered that:

$$U_o^* = U_o(k+1) = U_o(k) \quad (19)$$

By combining Equations (12), (18), and (19), we can obtain:

$$U_o^{re}(k+1) = U_o^* + \frac{i_o(k)T_s(L_d - L_{dm})}{L_{dm}C_{om}} \quad (20)$$

In Equation (20), it can be seen that the error in the parameters of the energy storage inductor is the dominant factor affecting the deviation value of the system's output voltage. When the parameters of the energy storage inductor are accurate ($L_d = L_{dm}$), the output voltage has no deviation and is not affected by the error in the parameters of the output capacitor; when there is an error in the parameters of the energy storage inductor ($L_d \neq L_{dm}$), the output voltage has a deviation, and the error in the capacitor model will affect the output voltage.

3.4. Minimize the Back-Flow Power

Taking advantage of the multi-objective optimization feature of model predictive control, the minimum back-flow power constraint is added to the cost function to achieve the optimization of the back-flow power. In this paper, to ensure that the output voltage can quickly and accurately track the reference value while maintaining the minimum back-flow power, the cost function is constructed as follows:

$$g = [P_{EPS,h} - 0]^2 + \gamma[U_{oref}^* - U_o(k+1)]^2 \quad (21)$$

Here, γ represents the constraint multiplier of the equation, and $\gamma \neq 0$. In order to enable the output voltage to quickly and accurately track the reference value, let $\frac{\partial g}{\partial \gamma} = 0$, and solve for it:

$$D_2 = \frac{1 + D_1 - \sqrt{1 - D_1^2 - 4\Delta}}{2} \quad (22)$$

In the equation, $\Delta = \frac{U_{oref}^* - U_o(k) - T_s f_{lu}(k)}{\alpha(k)T_s}$. In order to minimize the back-flow power, by setting $\frac{\partial g}{\partial D_1} = \frac{\partial g}{\partial D_2} = 0$, the solution is obtained as:

$$\begin{cases} D_1 = \frac{k(k+1) - \sqrt{2(k+1) - 4\Delta(k^2+2k+2)}}{k^2+2k+2} \\ D_2 = \frac{2(k^2+k+1) - (k+2)\sqrt{2(k+1) - 4\Delta(k^2+2k+2)}}{2(k^2+2k+2)} \end{cases}, 0 \leq D_1 \leq D_2 \leq 1 \quad (23)$$

Similarly,

$$\begin{cases} D_1 = \frac{k - \sqrt{k(1-k) + 1 - 2\Delta(k+1)}}{k+1} \\ D_2 = \frac{2k+1 - (k+2)\sqrt{k(1-k) + 1 - 2\Delta(k+1)}}{2(k+1)} \end{cases}, 0 \leq D_2 \leq D_1 \leq 1 \quad (24)$$

4. Lagrange Multiplier Method (LMM) for Optimizing Current Stress

For the given transmission power conditions, to find the optimal internal shift and the optimal external shift that minimize the inductor current stress, this type of problem can be summarized as solving the problem of the optimal solution under equality constraints. The Lagrange multiplier method (LMM), as a commonly used optimization algorithm for solving equality constraints, is more suitable for solving the minimum current stress under EPS modulation, and it is faster and requires less computation compared to the traditional derivative-based methods [22]. This paper employs the LMM numerical optimization current stress method to solve the internal migration problem and compares it with the combination of D_1 and model-free prediction to enhance the overall performance of DAB.

The problem of determining the optimal current stress under the modulation of EPS with a given transmission power as a constraint condition can be formulated as follows:

$$\begin{cases} \min i_{L_d.\max}(D_1, D_2) \\ s.t., P_{EPS}(D_1, D_2) = P' \end{cases} \quad (25)$$

The general form of the Lagrange multiplier method is:

$$F_E(x, y, \mu) = f_E(x, y) + \mu \zeta(x, y) \quad (26)$$

Among them, $F_E(x, y, \mu)$ represents the constructed Lagrange function; $f_E(x, y)$ represents the target function relationship to be solved; $\zeta(x, y)$ represents the equality constraint condition; and μ is the Lagrange multiplier.

By specifically integrating the LMM method into the EPS modulation to solve the problem of optimal current stress, we obtain:

$$E(D_1, D_2, \lambda) = i_{L_d.\max}(D_1, D_2) + \lambda(P_{EPS} - P') \quad (27)$$

In the above equation, λ is the Lagrange multiplier designed for solving the optimal current stress optimization problem under EPS modulation, and P' is the given reference transmission power. To solve for the minimum current stress, taking the partial derivative of it gives:

$$\begin{cases} \frac{\partial E}{\partial D_1} = 0 \\ \frac{\partial E}{\partial D_2} = 0 \\ \frac{\partial E}{\partial \lambda} = 0 \end{cases} \quad (28)$$

Based on the normalized current stress and transmission power expressions of the two states, and by combining with (27) and (28), the optimal shift ratio combination (D_1, D_2) under (a) $0 \leq D_1 \leq D_2 \leq 1$ and (b) $0 \leq D_2 \leq D_1 \leq 1$ is obtained as:

$$\left\{ \begin{array}{l} (a) \left\{ \begin{array}{l} D_1 = \frac{(k-1)\sqrt{(1-P_{EPS})(k^2-2k+2)}}{k^2-2k+2} \\ D_2 = \frac{1}{2} + \frac{(k-2)\sqrt{(1-P_{EPS})(k^2-2k+2)}}{2(k^2-2k+2)} \end{array} \right. , \frac{2k-2}{k^2} \leq P_{EPS} \leq 1 \\ (b) \left\{ \begin{array}{l} D_1 = 1 - \sqrt{\frac{P_{EPS}}{2(k-1)}} \\ D_2 = \frac{k-1}{2} + \frac{2-k}{2} \left(1 - \sqrt{\frac{P_{EPS}}{2(k-1)}} \right) \end{array} \right. , 0 \leq P_{EPS} \leq \frac{2k-2}{(3k-2)^2} \end{array} \right. \quad (29)$$

However, the current stress obtained in the case of mode (b) $0 \leq D_2 \leq D_1 \leq 1$ under EPS modulation does not meet the requirements for the converter in certain operating conditions, so mode (b) is further divided into $0 \leq D_2 \leq D_1 \leq 2D_2 \leq 1$ and $0 \leq D_2 \leq 2D_2 \leq D_1 \leq 1$ for more detailed analysis.

When $0 \leq D_2 \leq D_1 \leq 2D_2 \leq 1$, if $D_2 < D_1 \leq \frac{D_2+k-1}{k}$, then the maximum current stress is obtained at time t_4 ; that is:

$$i_{L_d}(t_4) = \frac{nU_o}{4f_s L_d} [k(1 - D_1) + 2D_2 - 1] \quad (30)$$

When $0 \leq D_2 \leq 2D_2 \leq D_1 \leq 1$, if $\frac{D_2+k-1}{k} < D_1 \leq 2D_2$, then the maximum current stress is obtained at time t_1 ; that is:

$$i_{L_d}(t_1) = \frac{nU_o}{4f_s L_d} (kD_1 + 1 - k) \quad (31)$$

Similarly, the maximum transmission power of SPS is used as the benchmark to unitize Equations (30) and (31):

$$i_{L_d, \max} = \begin{cases} 2[k(1 - D_1) + 2D_2 - 1], & D_2 < D_1 \leq \frac{D_2+k-1}{k} \\ 2(kD_1 + 1 - k), & \frac{D_2+k-1}{k} < D_1 \leq 2D_2 \end{cases} \quad (32)$$

Similarly, the optimal shift ratio combination calculated based on the LMM optimized current stress for the two subdivided cases in mode (b) $0 \leq D_2 \leq D_1 \leq 1$ can be obtained as:

$$(b) \left\{ \begin{array}{l} \left\{ \begin{array}{l} D_1 = \frac{4k-3-\sqrt{1-(4k-2)P_{EPS}}}{4k-2} \\ D_2 = \frac{3k-2-k\sqrt{1-(4k-2)P_{EPS}}}{4k-2} \end{array} \right. , 0 \leq P_{EPS} < \frac{2k-2}{(3k-2)^2} \\ \left\{ \begin{array}{l} D_1 = 1 - \sqrt{\frac{P_{EPS}}{2(k-1)}} \\ D_2 = \frac{k-1}{2} + \frac{2-k}{2} \left(1 - \sqrt{\frac{P_{EPS}}{2(k-1)}} \right) \end{array} \right. , \frac{2k-2}{(3k-2)^2} \leq P_{EPS} < \frac{2k-2}{k^2} \end{array} \right. \quad (33)$$

Combining Equations (29), (32), and (33), the minimum inductance current stress based on LMM optimization within different normalized transmission power ranges P_{EPS} under EPS modulation is:

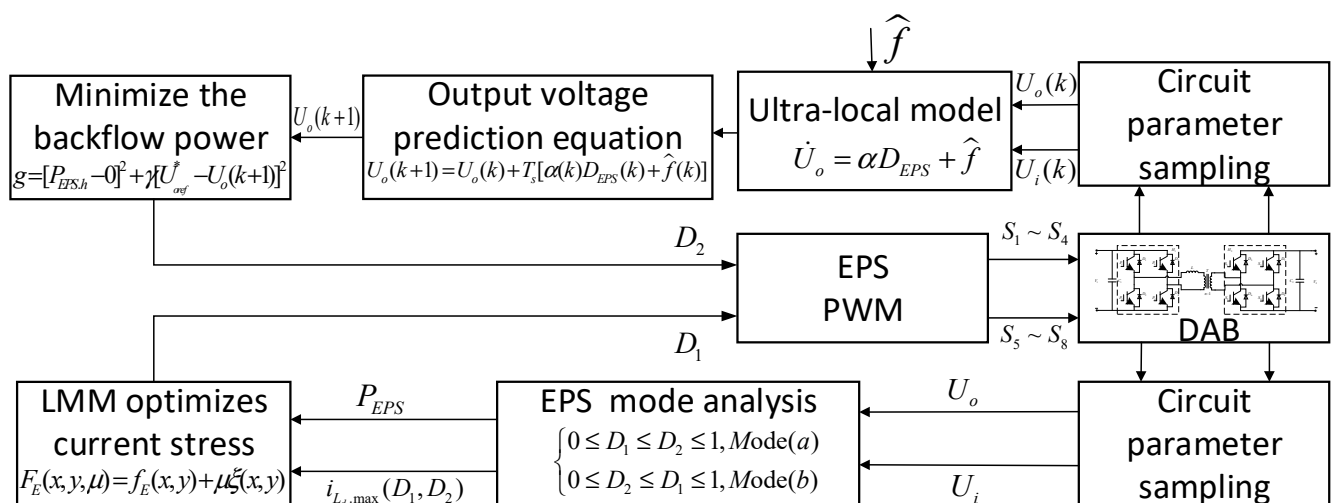
$$i_{L_d, \max} = \begin{cases} 2k - 2\sqrt{(1 - P_{EPS})(k^2 - 2k + 2)}, & \frac{2k-2}{k^2} \leq P_{EPS} \leq 1 \\ 4(k-1)\sqrt{\frac{P_{EPS}}{2(k-1)}}, & \frac{2k-2}{(3k-2)^2} \leq P_{EPS} < \frac{2k-2}{k^2} \\ \frac{3k-2-k\sqrt{1-(4k-2)P_{EPS}}}{2k-1}, & 0 \leq P_{EPS} < \frac{2k-2}{(3k-2)^2} \end{cases} \quad (34)$$

In conclusion, by combining Equations (29) and (33), the optimal shift ratio obtained by optimizing the current stress based on the LMM method under EPS modulation is as shown in Table 1. The corresponding current stress expression in this case is also presented in Table 1.

Table 1. Expression for the minimum current stress and shift ratio within different transmission power ranges under EPS modulation.

P_{EPS}	$0 \leq P_{EPS} \leq \frac{2k-2}{(3k-2)^2}$	$\frac{2k-2}{(3k-2)^2} \leq P_{EPS} \leq \frac{2k-2}{k^2}$	$\frac{2k-2}{k^2} \leq P_{EPS} \leq 1$
D_1	$\frac{4k-3-\sqrt{1-(4k-2)P_{EPS}}}{4k-2}$	$1 - \sqrt{\frac{P_{EPS}}{2(k-1)}}$	$\frac{(k-1)\sqrt{(1-P_{EPS})(k^2-2k+2)}}{k^2-2k+2}$
D_2	$\frac{3k-2-k\sqrt{1-(4k-2)P_{EPS}}}{4k-2}$	$\frac{k-1}{2} + \frac{2-k}{2} \left(1 - \sqrt{\frac{P_{EPS}}{2(k-1)}}$	$\frac{1}{2} + \frac{(k-2)\sqrt{(1-P_{EPS})(k^2-2k+2)}}{2(k^2-2k+2)}$
$i_{L_d,max}$	$\frac{3k-2-k\sqrt{1-(4k-2)P_{EPS}}}{2k-1}$	$4(k-1)\sqrt{\frac{P_{EPS}}{2(k-1)}}$	$2\sqrt{(1-P_{EPS})(k^2-2k+2)}$

The previous text proposed the use of the LMM method to optimize the current stress. Compared with the traditional current stress optimization methods, this approach reduces computational complexity and volume and can effectively lower the current stress. In fact, the optimal shift obtained by optimizing the current stress based on the LMM method is only compared with the D_1 combination, and the external shift compared with D_2 , although it is the stress-optimal solution, cannot guarantee stable voltage transmission from the primary side to the secondary side; that is, the output voltage cannot achieve precise following of the given reference voltage value. The second part of this paper proposes the MFPC control strategy under EPS modulation. By solving the optimal D_2 , it achieves precise, fast, and stable output voltage. Therefore, in order to reduce the current stress of the converter, improve the system transmission efficiency, and simultaneously achieve the improvement of the system's dynamics and robustness, this section redesigns the MFPC algorithm under EPS modulation and combines it with the LMM current stress optimization algorithm to propose a multi-objective optimization control strategy based on the combination of MFPC and LMM, to optimize the current stress and simultaneously improve the system's dynamic characteristics and robustness. The control block diagram of the model-free predictive control combined with the LMM optimization method under the DAB converter EPS modulation is shown in Figure 5.

**Figure 5.** Control block diagram of model-free predictive control for comprehensive performance optimization of the DAB system.

5. Simulation Verification

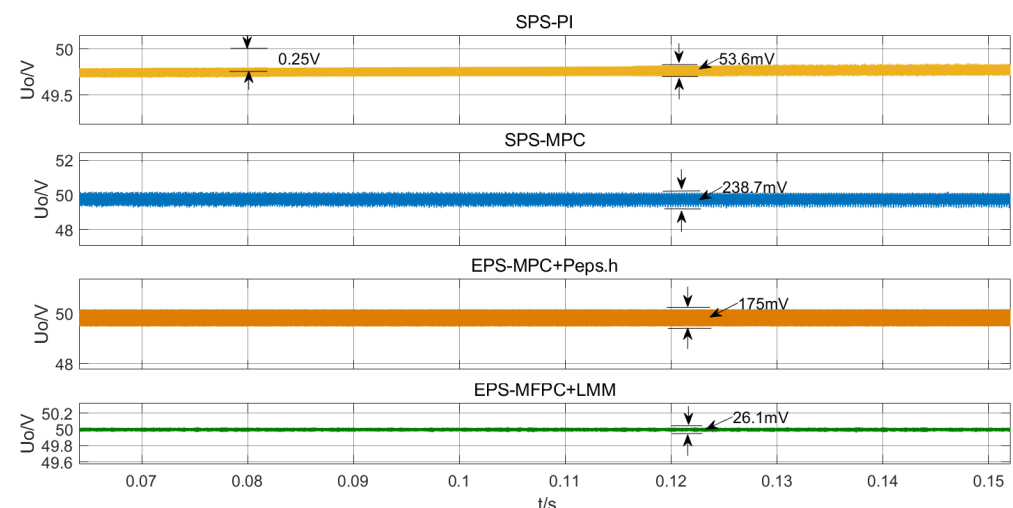
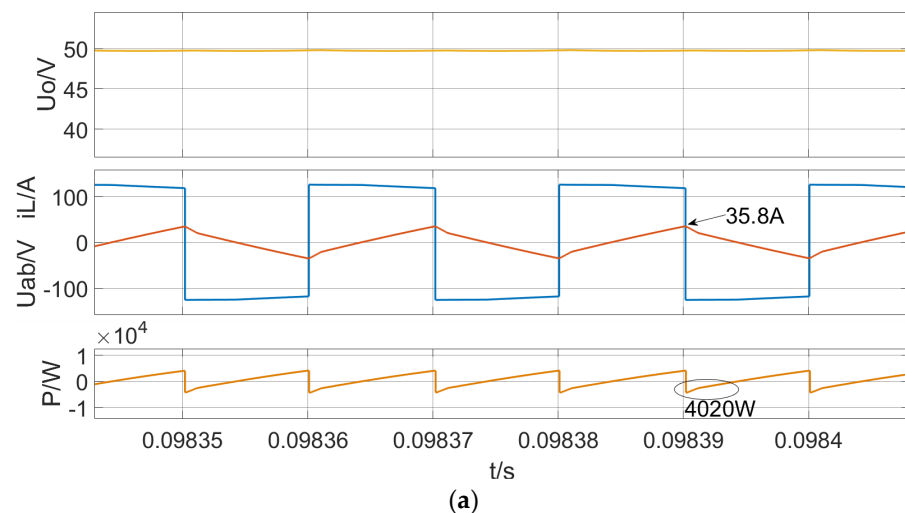
To further verify the correctness and feasibility of the theoretical analysis, a simulation model was established in MATLAB2024(a)/Simulink. The model parameters are shown in Table 2.

Table 2. Main parameters of the DAB simulation model.

Parameters	Numerical Value
Input voltage U_i /V	125
Output voltage U_o /V	50
Switching frequency f_s /kHz	50
Equivalent inductance L_d / μ H	12
Input capacitor C_i /mF	1
Output capacitor C_o /mF	1
High-frequency transformer ratio n	1
Load R / Ω	5

5.1. Verification of Steady-State Performance

Using the SPS-PI, SPS-MPC, and EPS-MPC + Peps.h methods as references, the steady-state performance of the proposed EPS-MFPC + LMM strategy is compared and verified. Figures 6 and 7 show the simulation waveforms of the output voltage U_o , the primary-side H-bridge output voltage U_{ab} , the inductor current i_L , and the calculated DAB converter transmission power P under the control of SPS-PI, SPS-MPC, and EPS-MPC + Peps.h, as well as the strategy proposed in this paper. When the input voltage is $U_i = 125$ V, the output voltage reference value is $U_o = 50$ V, and the load resistance is $R = 5$ Ω .

**Figure 6.** Comparison of output voltages for four control strategies.**Figure 7.** Cont.

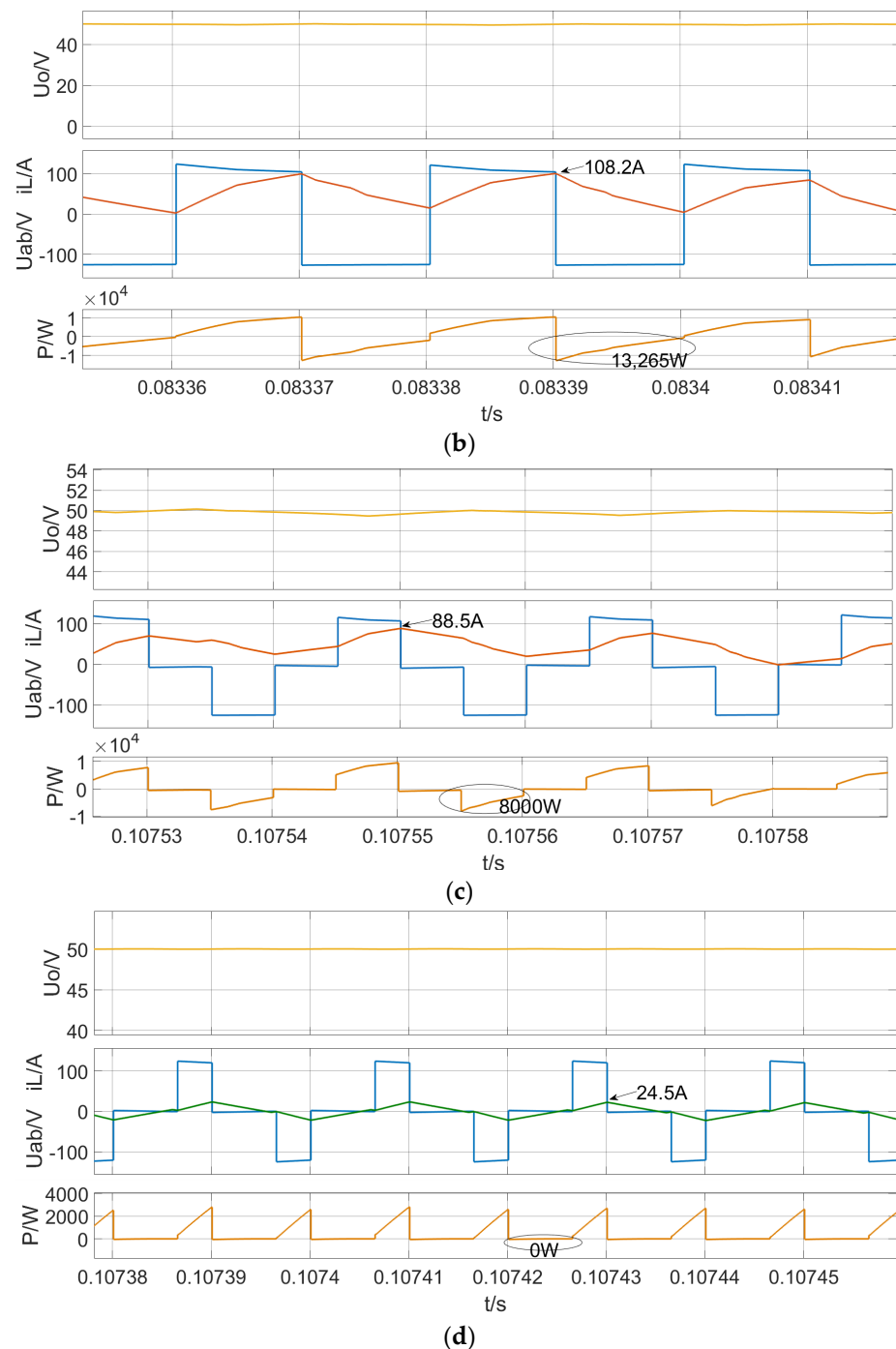


Figure 7. Comparison of steady-state performance among three methods of the DAB converter: (a) SPS-PI; (b) SPS-MPC; (c) EPS-MPC + Peps.h; (d) EPS-MFPC + LMM.

5.1.1. Output Voltage Ripple Comparison

The comparison of output voltage ripples during the steady-state operation of the converter is shown in Figure 5. In the stable state, there is a static error ($\Delta U_o = 0.25$ V) between the output voltage U_o of the SPS-PI method and the desired output voltage U_{oref} . The output voltages of the other three control algorithms can accurately track the reference value. The output voltage ripples of SPS-PI, SPS-MPC, EPS-MPC + Peps.h, and EPS-MFPC + LMM are 53.6 mV, 238.7 mV, 175 mV, and 26.1 mV, respectively. The method proposed in this paper has the smallest voltage ripple. Compared with the other three methods, it has, respectively, reduced them by 51.3%, 89.1%, and 85.1%, demonstrating the superiority of this method.

5.1.2. Back-Flow Power and Current Stress Comparison

When the converter operates in a steady state, for the current stress, as can be seen from the four graphs in Figure 7, the strategy proposed in this paper has the minimum current stress (24.5 A). Compared with the other three methods, it has, respectively, reduced them by 31.6%, 77.4%, and 72.3%. For the back-flow power, when other methods are employed, the transmission power contains negative values in each switching cycle, and there is a relatively large back-flow power in the other three control methods. In this paper, the control strategy of MFPC combined with LMM is adopted. In Figure 7d, the back-flow power of the system is basically optimized to zero, which verifies the correctness and effectiveness of the proposed control strategy. The main performance indicators of the four strategies in the steady state are shown in the Table 3 below.

Table 3. Comparison of the main simulation results at a steady state.

Indicators Strategy	Steady-State Voltage Ripple (mV)	Current Stress (A)	Back-Flow Power (W)	Settling Error (V)
SPS-PI	53.6	35.8	4020	0.25
SPS-MPC	238.7	108.2	13,265	0
EPS-MPC + Peps.h	175	88.5	8000	0
EPS-MFPC + LMM	26.1	24.5	0	0

Meanwhile, the inductor current and primary node voltage of the DAB converter under the four control methods remained stable. Therefore, the experimental results indicate that all the control algorithms have good steady-state performance.

5.2. Transient Performance Verification

To verify the dynamic characteristics of the traditional SPS-PI, SPS-MPC, and EPS-MPC + Peps.h control algorithms and the control algorithm proposed in this paper, experiments were conducted under the conditions of output voltage sudden change and load sudden change. The specific algorithm comparison simulation results are presented below.

5.2.1. Output Voltage Reference Value Step Change

Figures 8 and 9 show the reference value mutation waveform of the output voltage of the DAB converter. The total input voltage $U_i = 125$ V and the output voltage U_o changed from 50 V to 75 V and 25 V, respectively.

In Figure 8, it can be seen that all four control methods can accurately track the reference value and maintain stability without any static errors. The current stresses of the three methods, SPS-PI, SPS-MPC, EPS-MPC + Peps.h, and EPS-MFPC + LMM, are 69.1 A, 122 A, 71.5 A, and 34.2 A, respectively. This strategy has reduced them by 50.5%, 72.0%, and 52.2% to a greater extent compared to the other three strategies. It can be observed that the strategy proposed in this paper can significantly reduce the current stress of the converter when the reference voltage fluctuates. In terms of back-flow power, the instantaneous 566 W under the proposed strategy is also much smaller than the 7678 W, 12,280 W, and 7917 W under the other three control methods, and zero back-flow power operation is resumed after reaching the steady state.

As can be seen in Figure 9, the dynamic adjustment times of the SPS-PI and SPS-MPC methods are relatively long, approximately 10 ms; the EPS-MPC + Peps.h method is approximately 3.3 ms, while the EPS-MFPC + LMM method is approximately 9 ms. None of the four methods has static errors. For the current stress of the converter, the strategy proposed in this paper is approximately 15.5 A, which is significantly lower than the 43.8 A, 43.5 A, and 38.2 A under the other three control methods; the reduction amounts are 64.6%,

64.4%, and 59.4%, respectively. For the back-flow power, the strategy proposed in this paper has basically achieved zero back-flow power.

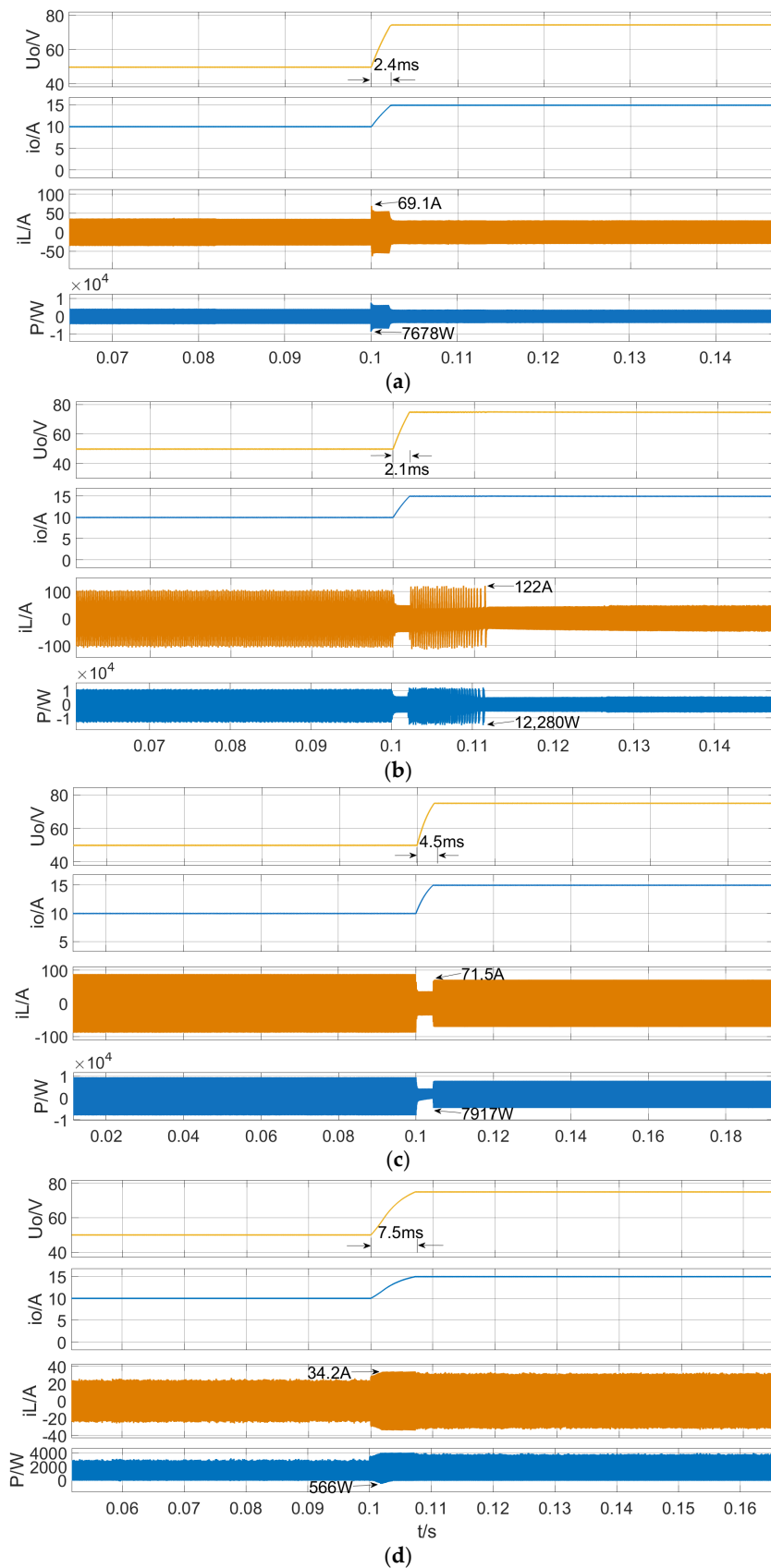


Figure 8. Spike in the output voltage reference value: (a) SPS-PI; (b) SPS-MPC; (c) EPS-MPC + Peps.h; (d) EPS-MFPC + LMM.

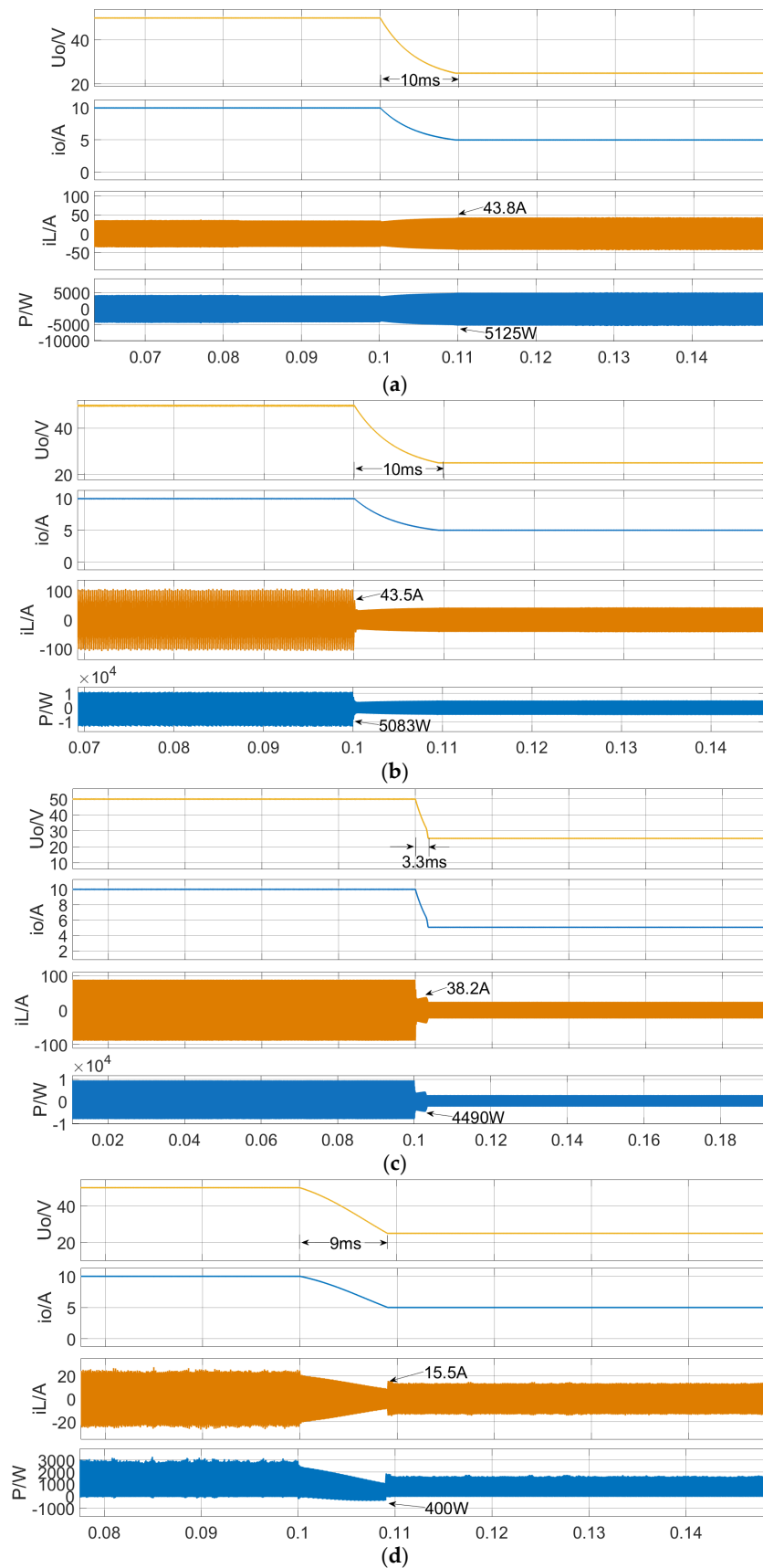


Figure 9. Sudden decrease in output voltage reference value: (a) SPS-PI; (b) SPS-MPC; (c) EPS-MPC + Peps.h; (d) EPS-MFPC + LMM.

5.2.2. Load Resistance Step Change

To compare the load disturbance resistance suppression capabilities of the traditional control methods with the proposed strategy, a step comparison experiment of the output-side load resistance under four strategies was conducted. The total input voltage $U_i = 125$ V, the output voltage $U_o = 50$ V, and the load resistance changed from $5\ \Omega$ to $10\ \Omega$ and $2.5\ \Omega$, respectively.

As shown in Figure 10, the adjustment time of the traditional SPS-PI method to restore the latest steady state is about 2 ms. When the proposed method is applied, the response speed has almost no adjustment time compared to the SPS-MPC and EPS-MPC + Peps.h methods. The output voltage fluctuation value and the current stress are smaller (compared with the other three methods, there were reductions of 10.5%, 73.4%, and 69.3%, respectively), the reference voltage can still be tracked quickly, and the steady state can be recovered faster when the load increases suddenly.

As shown in Figure 11, when the output-side load resistance of the DAB converter suddenly changes from $5\ \Omega$ to $2.5\ \Omega$, under the SPS-PI control, there is a static error ($\Delta U_o = 1$ V), and the current stress is approximately 41.4 A; under the SPS-MPC control, there is no obvious adjustment time, but there is still a large current stress of approximately 91.4 A; and under the EPS-MPC + Peps.h control, there is a steady-state output voltage error of 2.65 V and a large current stress of approximately 64.8 A. The strategy in this paper has an output voltage drop of approximately 1 V, and it takes approximately 5.5 ms to restore to a new stable state. It still has the smallest current stress of approximately 39.5 A among the four strategies (compared with the other three methods, there were reductions of 4.6%, 56.8%, and 39.0%, respectively). The main parameters of the simulation under the four strategies are summarized in Table 4 as follows.

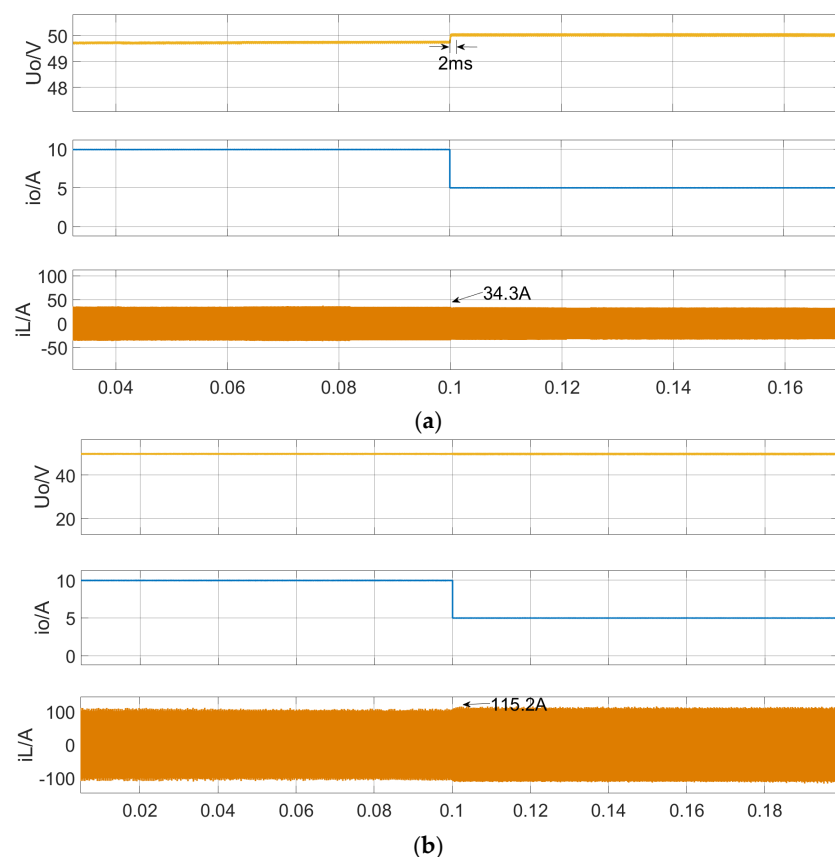


Figure 10. Cont.

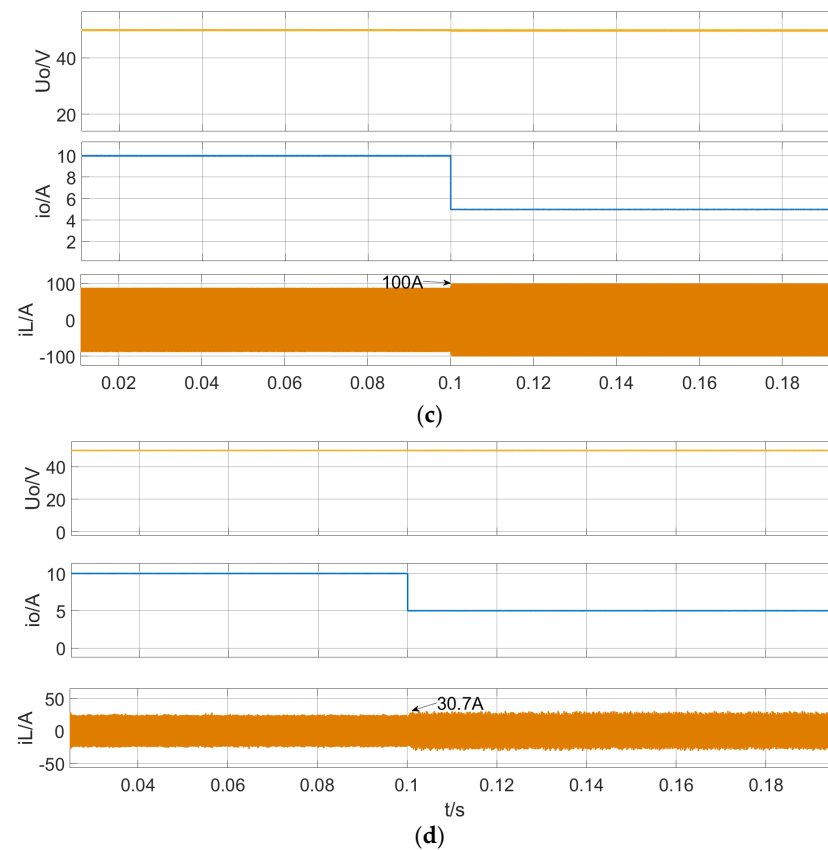


Figure 10. Sudden increase in load resistance: (a) SPS-PI; (b) SPS-MPC; (c) EPS-MPC + Peps.h; (d) EPS-MFPC + LMM.

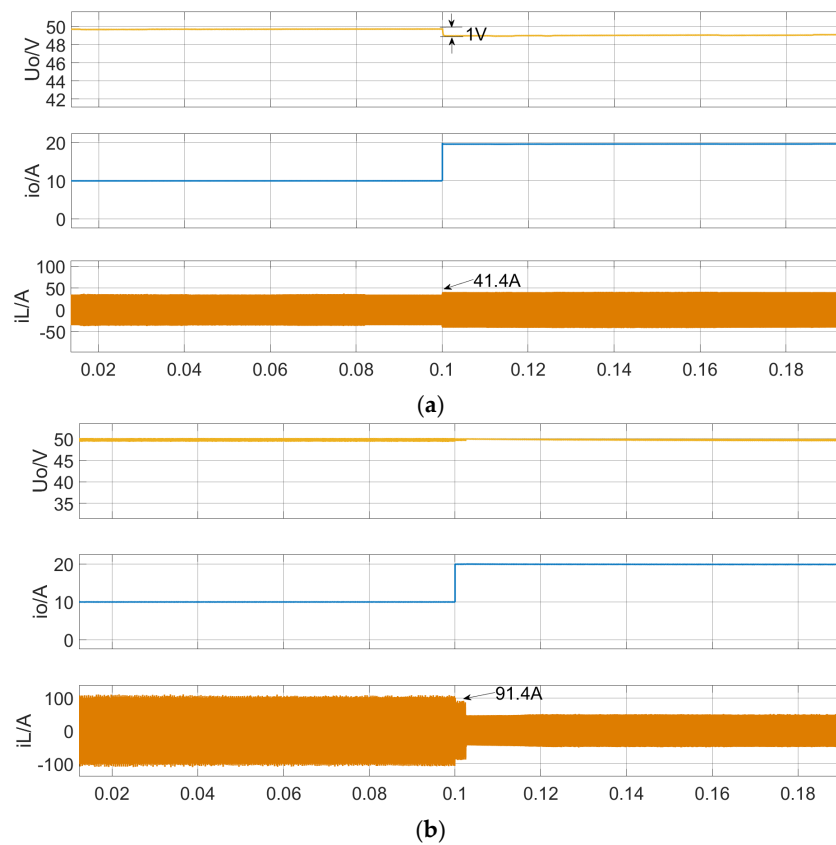


Figure 11. Cont.

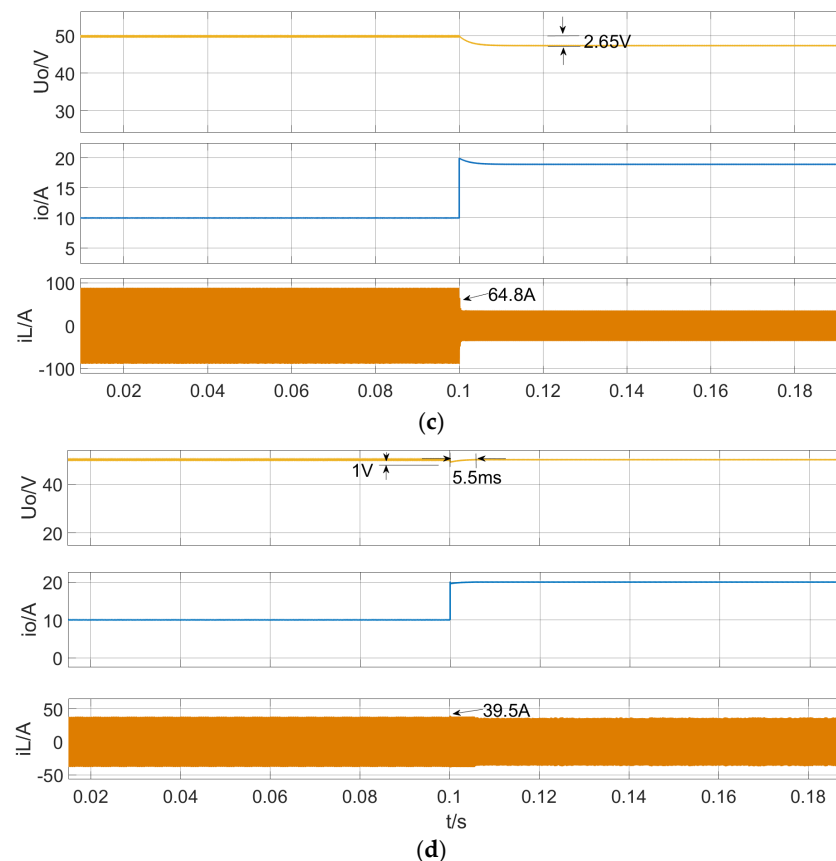


Figure 11. Sudden reduction in load resistance: (a) SPS-PI; (b) SPS-MPC; (c) EPS-MPC + Peps.h; (d) EPS-MFPC + LMM.

Table 4. Comparison of the main parameters of the dynamic performance simulation results.

Strategy	Post-Increase Output Voltage Reference Value			Post-Reduction Output Voltage Reference Value			Post-Increase Resistance Load		Post-Reduction Resistance Load		
	Recovery Time (ms)	Current Stress (A)	Back-Flow Power (W)	Recovery Time (ms)	Current Stress (A)	Back-Flow Power (W)	Recovery Time (ms)	Current Stress (A)	Recovery Time (ms)	Current Stress (A)	Steady-State Error (V)
SPS-PI	2.4	69.1	7678	10	43.8	5125	2	34.3	0	41.4	1
SPS-MPC	2.1	122	12,280	10	43.5	5083	0	115.2	0	91.4	0
EPS-MPC + Peps.h	4.5	71.5	7917	3.3	38.2	4490	0	100	0	64.8	2.65
EPS-MFPC + LMM	7.5	34.2	566	9	15.5	400	0	30.7	5.5	39.5	0

6. Conclusions

Aiming at the problem of strong parameter dependence of the DAB converter under the model predictive control strategy, a model-free control method based on ultra-local model combination and MPC is proposed in this paper, which solves the disadvantage of model predictive control relying too much on model parameters and improves the performance of the DAB converter. From the simulation results, it can be seen that compared with traditional control methods, the voltage ripple has been reduced by 51.3%, 89.1%, and 85.1%, respectively. The strategy proposed in this paper has superior steady-state performance. The current stress of the DAB converter decreases in the case of the output voltage reference value step (compared with the other three strategies, there were reductions of 50.5%, 72.0%, and 52.2%, respectively, in the case of sudden increases, and there were reductions of 64.6%, 64.4%, and 59.4%, respectively, in the case of sudden decreases) and basically reaches zero back-flow power. In the case of sudden changes in the output load, the current stress (there were reductions of 10.5%, 73.4%, and 69.3%, respectively, in the case of sudden increases, and there were reductions of 4.6%, 56.8%, and 39.0%, respectively,

in the case of sudden decreases) and dynamic performance under this strategy are also superior to other methods.

Therefore, it can be concluded that the proposed method not only solves the problem of strong dependence on traditional MPC parameters but also effectively reduces the current stress and back-flow power of the converter, improves the steady and dynamic performance of the converter, and realizes the collaborative optimization of the comprehensive performance of the DAB converter. However, the drawback of this paper lies in that it only conducted simulation verification of the proposed strategies; it did not conduct tests on physical hardware. In future research, experimental tests will be carried out to further verify the effectiveness of the algorithms presented in this paper.

Author Contributions: Conceptualization, L.W. and D.L.; methodology, L.W. and D.L.; software, L.W. and G.Q.; validation, L.W. and B.C.; writing—original draft preparation, L.W.; writing—review and editing, D.L. and Y.C. All authors have read and agreed to the published version of the manuscript.

Funding: This research was funded by The Graduate Innovation Program Project of Beihua University, grant number 2025055, and the Science and Technology Research Project of the Education Department of Jilin Province (JJKH20261143KJ).

Data Availability Statement: Data are contained within the article.

Conflicts of Interest: The authors declare no conflicts of interest.

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