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Flexible Transient Design-Oriented Model Predictive Control for Power Converters

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Abstract—Model predictive control (MPC) is one of the promising candidates for power converters with improved stability and faster response. Overshoot, a crucial transient precursor, is a custom design requirement based on the converters' applications having variant time-domain specifications that are subject to their safety and reliability. Apart from the generally accepted weight design process for improved settling time, when considering an MPC-controlled system, a theoretical framework to adjust the overshoot requirements has been overlooked so far. To bridge this gap, this article proposes a new method that integrates a flexible output voltage overshoot design method into MPC-controlled dc–dc boost converters, thereby formally ensuring that the controller always complies with the time-domain specifications under dynamic disturbances. First, this article provides the MPC controller's model by deriving the optimal duty cycle value, which is then used to derive the closed-loop transfer functions. Next, the cutoff frequency is obtained based on the transfer functions. The weighting factor ratio of the MPC, which is a tunable parameter is then adjusted to ensure an adequate system's cutoff frequency. Finally, the effectiveness and ruggedness of the proposed method are validated under experimental

conditions, allowing flexibility to extend its use for different power converter applications.

Index Terms—Cutoff frequency, dynamic performance, model predictive control (MPC), overshoot design, weighting factors.

I. INTRODUCTION

NOWADAYS, global ambitions of energy efficiency and carbon neutrality have put forward new design challenges in future power grids [1], [2]. In response, developments in electric vehicles (EVs), and renewable energy generation technologies have scaled up. To manage the power in such systems (EV charging stations and photovoltaic generation) more efficiently, the dc microgrid has gained a lot of academic and industrial attention [3], [4], [5], [6]. A typical structure of a dc microgrid involves several dc–dc converters connected via different topologies to a dc bus, with each converter supporting different output voltage levels for various loads [6]. To ensure a stable and smooth operation of the microgrid systems, control techniques play a crucial role. Model predictive control (MPC) is one of the promising candidates to tackle the control problems of such systems, offering benefits, such as improved stability, ease of implementation, explicit consideration of constraints, and other benefits [7], [8], [9].

Stability and dynamic performance are two vital performance indicators of power converters' operation. When looking into the stability performance, some of the typical stability issues, such as the instability of the system caused by the converters, which act as the constant power loads (CPL) have already been studied thoroughly in many previous works [10], [11], [12]. Considering the dynamic process, the authors in [13], [14], [15], and [16] proposed several MPC-based algorithms to improve the performance. More specifically, Song et al. [13] proposed a virtual power control and MPC combined algorithm to ensure an excellent dynamic response for isolated dc–dc converters. A double-loop control framework is proposed in [14] for microgrids. The inner loop with double vector MPC is utilized to track the output voltage derived from the outer loop, and this control strategy improves the control accuracy as well as the dynamic response. In [15], the solution to enhance the dynamic performance is to incorporate an additional item into the cost function. This additional item is composed of

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the output current, the output voltage reference as well as a weighting factor that balances the relative importance of the two items. However, a tradeoff should be considered between dynamic performance and steady-state error. In [16], artificial neural networks (ANN)-assisted MPC is proposed, which also selects weighting factors to enhance the dynamic response. The basic idea behind using ANN is to compare the response time with different combinations of weighting factors and then determine the optimal strategy. Although the said MPC-based control methods could enhance the dynamic performance, they cannot guarantee that the overshoot will be limited to a specific value, which affects their safety and reliability.

Dynamic overshoots exceeding the prespecified limits may have an impact on the performance and the lifetime of equipment connected to the dc microgrids and the converters themselves [17]. In addition, an overshoot in the output voltage may lead to electromagnetic interference problems, where the high-frequency components of the overshoot within every switching period can radiate and interfere with other electronic devices and circuits, leading to malfunction and performance degradation. Furthermore, if the system is not quickly damped, stability and transient response may also degrade, resulting in oscillations that can further negatively affect the overall performance. In practical applications, the overshoot of converters is always limited within the safety specifications. For instance, the design specification of a 12 V output dc–dc boost converter limits the output overshoot below 0.6 V when the load current changes with 1.5 A under specific circuit parameters [18]. In a battery charging specification for a universal serial bus (USB) application, the maximum overshoot of the charging port is 6 V, and the maximum undershoot voltage of the same port is 4.1 V with a 5 V rated output voltage [19]. Considering an EV charging station, the output voltage overshoot should not exceed 10% of the rated output voltage in a DS/EN 61851-23 standard [20]. In addition, the DS/EN 16603-20-20 standard specifies the allowable operational current in the electrical design and interface requirements for a power supply [21]. Moreover, the voltage deviation of pulse generators, according to ISO 7637 standard, is also illustrated with different levels from 5%–10% based on different load conditions [22]. Therefore, when designing a system, it is necessary to consider overshoot and incorporate it into the control loop design process.

The design of an overshoot in a system can be analyzed both with the time domain and frequency domain [17]. In the time domain, a concept of state overshoot for the systems is proposed based on the overshoot definition and norm. Then, the value of the overshoot can be expressed as the absolute value of the circuit variables [17], [23]. However, this approach may not fully express the overshoot based on the sum of the absolute value of each variable due to the inherent dependency between each variable. In the frequency domain, the Bode diagram is often used to qualitatively design overshoots in dc–dc converters [24]. By deriving the cutoff frequency, the overshoot can easily be determined [18]. However, these methodologies primarily focus on PI-controlled systems, with a little discussion of extension into MPC-controlled systems. Therefore, the main objective of this article is to achieve a specifically constrained overshoot/undershoot for the MPC-controlled system by

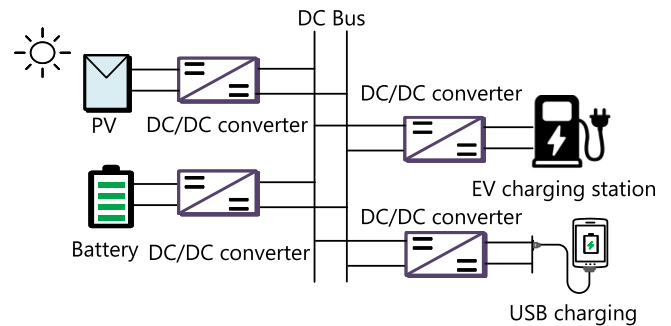


Fig. 1. Typical dc microgrid system with different types of loads and sources.

carefully designing the parameters in the controller. Notably, the study focuses on the continuous control set-MPC-(referred to as MPC for simplicity in the text)controlled system. This algorithm is chosen for its capability to generate a fixed switching frequency, effectively mitigating issues, such as high acoustic noises, operation mode uncertainty, aggressive ripples, and so on [9].

With a particular focus on flexibility induced into MPC to assist the transient design compliance, the key contributions of this article are as follows.

- 1) This article proposes a novel overshoot design method for MPC-controlled converters. The approach involves developing a small signal model to establish the relationship between the time domain specification and the system. Based on the system's cutoff frequency, the controller's parameter is determined to meet the desired overshoot specification effectively.
- 2) It introduces a new way for a flexible adjustment of the proposed method to satisfy the overshoot specifications. By conducting frequency domain analysis, the proposed method allows for achieving different overshoot requirements simply by modifying the ratio of weighting factors in the MPC controller. This efficient adjustment capability enables an effective control of overshoot under different operating conditions.

The rest of this article is organized as follows. Section II illustrates the basic model of the boost converter. Section III proposes the overshoot design method of the MPC-controlled system. Experiments are provided in Section IV. Finally, Section V concludes this article.

II. BASIC PRINCIPLE OF MPC-CONTROLLED DC–DC BOOST CONVERTER

Fig. 1 depicts a typical dc microgrid system with various types of loads. As previously mentioned, each load type has its unique overshoot requirement. Therefore, the objective is to design the MPC controller for each converter to meet their respective overshoot criteria, as illustrated in Fig. 2. The process entails the following steps. First, the parameters of each system and the design requirement (overshoot/undershoot under certain current changes) are determined. Next, the design method calculates the control parameters for each MPC controller. Finally, the MPC controller generates a switching signal to regulate the

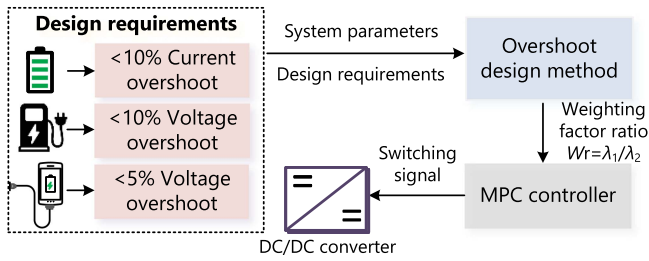


Fig. 2. Framework of the overshoot design method for MPC-controlled converters.

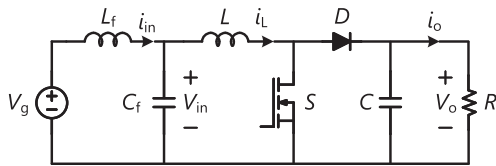


Fig. 3. DC-DC boost converter with the input LC filter.

converter. In this study, the voltage level in [18] is used as a demonstration.

Fig. 3 shows a single-line diagram of the considered system, which consists of an input LC filter and a boost converter. The inclusion of the input LC filter in the system serves multiple purposes. It helps attenuate high-frequency components, thereby ensuring a cleaner and more stable input voltage to the boost converter [25]. This is especially crucial in dc microgrids where sensitive loads or other power electronics devices may be connected to the same power source. Furthermore, due to the boost converter's CPL behavior, it is essential to avoid oscillation between the front-end system and the boost converter system. To achieve better control over the input voltage and prevent undesired interactions, an LC filter is equivalent to replacing the front-end converter [26]. V_g is the input voltage, V_{in} denotes the input voltage of the boost converter, i_{in} is the current across the inductor L_f and i_L is the inductor current of the boost converter, V_o represents the output voltage, i_o is the output current, and R is the load. Due to the following boost converter behaving as a CPL, whose incremental impedance is negative, it tends to destabilize the system to which it is connected. That means the input port V_{in} may oscillate without any control. Therefore, the goal of the controller is to ensure stable voltages at both the input and output ports. In this article, the difference equations are established using a single prediction horizon instead of long prediction horizons to reduce the computational burden. It should be noted that the inductor current is the control objective of the MPC-controlled boost converter to avoid nonminimum phase behavior [27], [28]. Therefore, in this article, V_{in} and i_L are selected as the control objectives. Based on the operating principle in continuous conduction mode, the following equations can be derived:

$$\frac{di_L}{dt} = \frac{1}{L}V_{in} - \frac{1-d}{L}V_o \quad (1)$$

$$\frac{dV_{in}}{dt} = \frac{1}{C_f}i_{in} - \frac{1}{C_f}i_L \quad (2)$$

where d is the duty cycle sampled at the k th instant. Assuming that the sampling frequency is relatively high, the state variables can be transformed into a discrete-time equation with the classical forward Euler approximation method. The difference equations containing two objectives (inductor current i_L and input port voltage V_{in}) are provided as

$$i_L(k+1) = i_L(k) + \frac{V_{in}(k) - V_o(k)}{L}T_s + \frac{dV_o(k)}{L}T_s \quad (3)$$

$$V_{in}(k+1) = V_{in}(k) + \frac{i_{in}(k) - i_L(k)}{C_f}T_s - \frac{V_{in}(k) - V_o(k) - dV_o(k)}{LC_f}T_s \quad (4)$$

where T_s denotes the switching cycle. The cost function can be derived as

$$J = \lambda_1(i_L(k+1) - i_L^*)^2 + \lambda_2(V_{in}(k+1) - V_{in}^*)^2 \quad (5)$$

where λ_1 and λ_2 are the weighting factors, i_L^* denotes the inductor current reference, and V_{in}^* represents the reference of the input voltage of the boost converter. In order to minimize the cost function, the derivative of the cost function to the duty cycle should satisfy

$$\frac{\partial J}{\partial d} = 0. \quad (6)$$

In this case, the optimal duty cycle value can be obtained according to the chain rule

$$d = \frac{(V_{in}(k) + \frac{i_{in}(k) - i_L(k) - \frac{V_{in}(k) - V_o(k)}{L}T_s}{C_f}T_s - V_{in}^*)}{\left(\frac{\lambda_1}{\lambda_2} + \frac{T_s^2}{C_f^2}\right) \frac{V_o(k)}{LC}} - \frac{\frac{\lambda_1}{\lambda_2}(i_L(k) + \frac{V_{in}(k) - V_o(k)}{L}T_s - i_L^*)}{\left(\frac{\lambda_1}{\lambda_2} + \frac{T_s^2}{C_f^2}\right) \frac{V_o(k)}{L}T_s}. \quad (7)$$

This optimal value of the duty cycle is modulated to generate a switching signal for the boost converter. To clarify whether the cost function J equals its minimum value when adopting the optimal variable d , the second derivative is utilized as follows:

$$\frac{\partial^2 J}{\partial d^2} = 2\lambda_1 \frac{V_o^2(k)T_s^2}{L^2} + 2\lambda_2 \frac{V_o^2(k)}{L^2 C_f^2} T_s^2 > 0. \quad (8)$$

The control variable d is in the specified range (0–1). Referring to (8), it is evident that the second-order derivative of the cost function to d is positive. Consequently, the cost function attains its minimum value by adopting the derived optimal value for d . By interconnecting the control loop and power stage loop using (7), the theoretical basis for the modeling process can be established in the subsequent step.

III. OVERSHOOT DESIGN OF THE MPC-CONTROLLED CONVERTER

A. Modeling of the MPC-Controlled Converter

To ensure a safe operation of a system, it is important to carefully consider the overshoot of the output voltage when the load changes. Overshoot can be predicted using the crossover frequency of the closed-loop system, which can reveal the system's adaptive ability during the dynamic process. In this study, the cutoff frequency is employed as an alternative to the crossover frequency. This choice is made because the cutoff frequency f_c signifies that the loop gain is below the unit value, thereby indicating an inadequate attenuation of disturbances [29]. In addition, the cutoff frequency provides information about the system's bandwidth, thereby aiding in subsequent design considerations [24].

Its relationship can be presented as [18]

$$\Delta V_o = \frac{\Delta I_o}{2\pi f_c C} \quad (9)$$

where ΔV_o is the output voltage overshoot, ΔI_o denotes the output current variation, and f_c is the cutoff frequency. Although predicting the overshoot can be complicated by the nonlinear behavior of the MPC-based systems, it provides a clear expression for the optimal duty cycle value based on the sampled value and system parameters in (7). By introducing a small perturbation and neglecting higher order perturbations, a small-signal equation can be obtained to analyze the system's response to changes in the load. This can help to predict and manage the overshoot in a more efficient and reliable manner.

By introducing the small-signal perturbation, (7) can be rewritten as

$$D + \hat{d} = \frac{I_{in} + \hat{i}_{in} - I_L - \hat{i}_L - \frac{V_{in} + \hat{v}_{in} - V_o - \hat{v}_o T_s}{L} T_s}{C_f} \quad (10)$$

$$- \frac{\left(\frac{\lambda_1}{\lambda_2} + \frac{T_s^2}{C_f^2} \right) \frac{V_o + \hat{v}_o}{LC}}{\left(\frac{\lambda_1}{\lambda_2} + \frac{T_s^2}{C_f^2} \right) \frac{V_o + \hat{v}_o}{LC}}$$

$$- \frac{\frac{\lambda_1}{\lambda_2} (I_L + \hat{i}_L + \frac{V_{in} + \hat{v}_{in} - V_o - \hat{v}_o T_s}{L} T_s - I_L^* - \hat{i}_L^*)}{\left(\frac{\lambda_1}{\lambda_2} + \frac{T_s^2}{C_f^2} \right) \frac{V_o + \hat{v}_o}{LC}}$$

$$- \frac{V_{in}^* + \hat{v}_{in}^*}{\left(\frac{\lambda_1}{\lambda_2} + \frac{T_s^2}{C_f^2} \right) \frac{V_o + \hat{v}_o}{LC}}.$$

Based on (10) and replacing the \hat{i}_L^* with the expression of \hat{v}_o^* , neglecting high-order perturbations, the following equation can be obtained:

$$\hat{d} = F_m (F_g \hat{v}_{in} + F_i \hat{i}_{in} + F_L \hat{i}_L + (F_v + F_o) \hat{v}_o + F_{inref} \hat{v}_{in}^* + F_{vref} \hat{v}_o^*) \quad (11)$$

TABLE I

OPEN LOOP TRANSFER FUNCTIONS OF THE BOOST CONVERTER

Open loop transfer functions	Values
Input current-to-inductor current G_{ii}	$\frac{1}{L_f C_f s^2 + 1}$
Input voltage-to-output voltage G_{vg}	$\frac{R(1-D)}{Ls(CRs+1)+R(1-D)}$
Duty ratio-to-output voltage G_{vd}	$\frac{RV_o(1-D) - \frac{V_o}{1-D} Ls}{Ls(CRs+1)+R(1-D)}$
Input voltage-to-inductor current G_{ig}	$\frac{CRs+1}{Ls(CRs+1)+R(1-D^2)}$
Duty ratio-to-inductor current G_{id}	$\frac{V_o(CRs+2)}{Ls(CRs+1)+R(1-D^2)}$

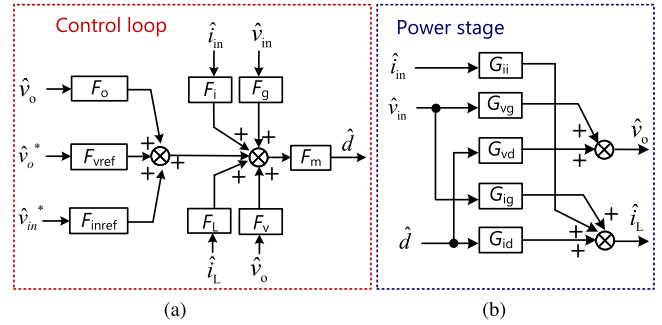


Fig. 4. Small signal diagram of the MPC controlled boost converter. (a) Control loop small signal diagram. (b) Power stage small signal diagram.

where $F_m = 1$, $F_i = kT_s^2/C_f^2$

$$F_g = k(1 - 1/(LC_f T_s^2)) T_s / C_f - \lambda_1 T_s / (\lambda_2 L)$$

$$F_L = -kT_s^2/C_f^2 - \lambda_1/\lambda_2$$

$$F_v = k(1 - D) T_s^3 / (LC_f^2) + \lambda_1 T_s / (\lambda_2 L)$$

$$F_o = kv_o^*/V_{in} R, \quad F_{vref} = k\lambda_1/\lambda_2$$

$$F_{inref} = kT_s / (C_f V_o V_{in} R)$$

$$k = 1/(V_o(k) \frac{T_s}{L} (\lambda_1/\lambda_2 + T_s^2 C_f^2)).$$

We define that G_{ii} represents the input current-to-inductor current transfer function, G_{vg} denotes the input voltage-to-output voltage transfer function, G_{vd} is the duty ratio-to-output voltage transfer function, G_{ig} stands for the input voltage-to-inductor current transfer function, and the duty ratio-to-inductor current transfer function is represented as G_{id} . The expressions of these open loop transfer functions are given in Table I. Using (11), the model of the MPC-controlled boost converter is shown in Fig. 4.

Typically, the cutoff frequency is obtained from the closed-loop transfer function. Therefore, it needs to derive the transfer function of the output voltage reference to the output voltage, as it forms a part of the control loop, as shown in Fig. 4. Setting the input perturbation \hat{v}_{in} , \hat{i}_{in} , and \hat{v}_{in}^* to zero, the closed-loop output voltage reference \hat{v}_o^* to output voltage \hat{v}_o transfer function G_{vr} can be obtained as

$$G_{vr} = F_m \frac{G_{vd} F_{vref}}{1 - (F_v + F_o) G_{vd} - F_L G_{id}}. \quad (12)$$

TABLE II
SYSTEM PARAMETERS OF THE DC-DC BOOST CONVERTER

Parameters	Symbols	Values
Input voltage	V_g	10 V
Output voltage	V_o	12 V
LC Inductance	L_f	0.8 mH
LC Capacitor	C_f	15 μ F
Inductance	L	1.5 mH
Capacitor	C	2000 μ F
Switching cycle	T_s	100 μ s
Output current	i_o	0.5A-2A
Output voltage reference	V_o^*	12 V
Inductor current reference	i_L^*	$V_o^* i_o(k) / V_g$
Input voltage reference	V_{in}^*	10 V

Based on (12), selecting the weighting as $\lambda_1 = 0.5$ and $\lambda_2 = 1$, the cutoff frequency f_c can be obtained using

$$20 \log_{10} G_{vrloop}(s = f_c) = -3\text{dB}. \quad (13)$$

Here, G_{vrloop} denotes the control loop of the G_{vr} . However, in practical and industrial environments, the cutoff frequency will vary with different system parameters, especially the inductance and capacitor. In this study, to simplify the control and analysis, a fixed inductor value and a fixed capacitor value are considered, which are given in Table II. It is also worth noting that the load resistance can also have an impact on the cutoff frequency. Typically, during the design phase, the rated load is used as a basis for this parameter, which can be obtained from specifications or through measurements [30], [31]. Furthermore, the maximum overshoot limitation is set as 1.5 V. Based on this assumption, it can be calculated that $f_c \approx 50$ Hz. Therefore, to ensure that the system meets the design specification, which stipulates that the voltage overshoot should be below 1.5 V when the load current changes by 1.5 A, it needs to determine the minimum cutoff frequency f_{cm} .

To reach the design compliance, the minimum cutoff frequency can be calculated as follows:

$$f_{cm} = \frac{1.5}{2\pi \times 1.5 \times 2000 \times 10^{-6}} \approx 80 \text{ Hz}. \quad (14)$$

It can be observed that the required minimum cutoff frequency f_{cm} is higher than the calculated value of f_c . This means that the system as currently designed will not meet the overshoot design requirement, which will exceed the maximum allowable voltage overshoot. Therefore, the system needs to adjust its parameters to increase the cutoff frequency.

B. Design of the Cutoff Frequency

In practical applications, some parameters of a system may be fixed and can be difficult to modify. However, adjusting the parameters in the control loop is possible, which can have a significant impact on the system's behavior. In (12), it can be seen that the transfer function G_{vr} can be modified by adjusting the weighting factors. Since the cost function only involves two weighting factors, the ratio of these factors can be employed to alter λ_1 . By setting λ_2 to unity and varying λ_1 , the weighting

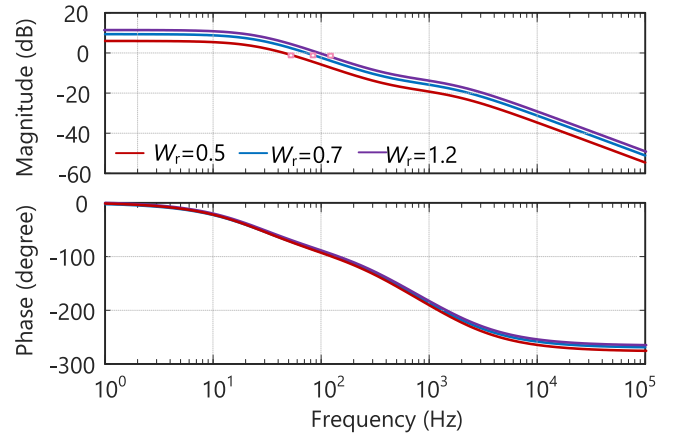


Fig. 5. Bode plots of the output voltage reference to output voltage transfer function G_{vr} with different weighting factor ratios—An increase in the weighting factor ratio results in a higher cutoff frequency.

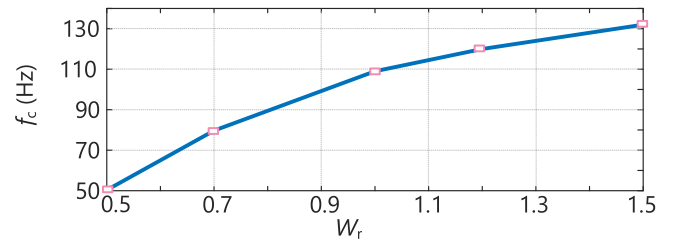


Fig. 6. Cutoff frequencies with different weighting factor ratios.

factor ratio $W_r = \lambda_1 / \lambda_2$ can be modified. Figs. 5 and 6 demonstrate the relationship between the weighting factor ratio and cutoff frequency, indicating that an increase in the weighting factor ratio results in a higher cutoff frequency.

Hence, a minimum weighting factor ratio W_{rm} can be determined as

$$20 \log_{10} G_{vrloop}(s = f_{cm}, \frac{\lambda_1}{\lambda_2} = W_{rm}) = -3 \text{ dB}. \quad (15)$$

The derived value of W_{rm} is approximately 0.7, implying that if the weighting factor ratio exceeds 0.7, the cutoff frequency f_c will exceed 80 Hz. Reflecting on the time domain, it means that the output voltage overshoot resulting from load changes of 1.5 A can be effectively limited to under 1.5 V.

Although increasing the weighting factor ratio can result in a higher cutoff frequency, there exist limitations to the control system's ability to reduce the overshoot even further. This is primarily due to the fact that the duty cycle of the system is always constrained within a certain range, such as the interval $d \in [0.1, 0.9]$ in this study. Consequently, the minimum achievable overshoot can be determined by analyzing the dynamic process of the system.

The system's performance during load changes is shown in Fig. 7. It can be observed that the overshoot reaches its peak value when the inductor current i_L arrives at its new steady-state value. The dynamic process can be described as follows: at the instant when the load changes, the output current changes,

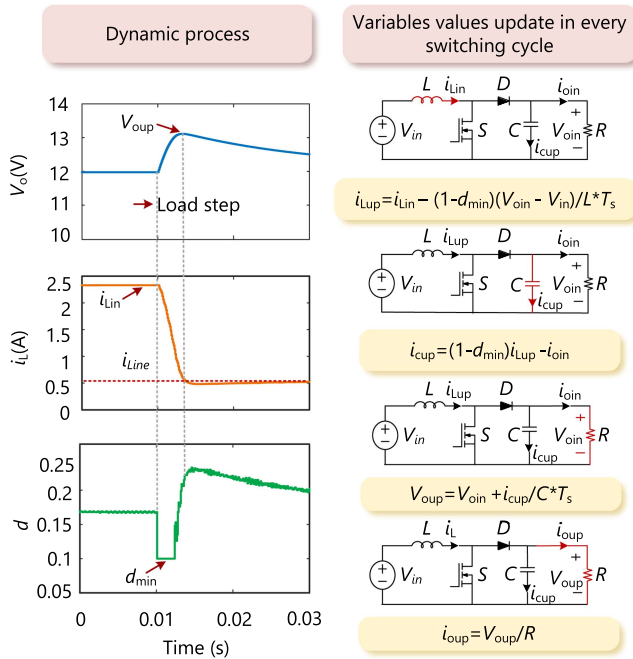


Fig. 7. Calculation of the minimum output voltage overshoot. (a) Dynamic response process. (b) Variables value updating process.

which is defined as i_{oin} . Next, the inductor current decreases with the limited minimum duty cycle of 0.1 in one switching cycle, and this is defined as i_{Lup} . Then, the capacitor current i_{cup} decreases, and the overshoot of the output voltage within one switching cycle can be derived based on capacitor current i_{cup} . Finally, the output voltage becomes the updated output voltage v_{oup} and the output current is updated as i_{oup} . The above iterations are repeated until the inductor i_L current reaches its new equilibrium value i_{Line} . The iteration process and related equations are expressed with a flowchart, as shown in Fig. 8. In the studied system, the minimum overshoot can be calculated as 0.9 V.

Based on this, the overall design process is complete, as shown in Fig. 8. First, the parameters and overshoot design specification ΔV_o are received. The next step is to obtain the minimum ΔV_{om} based on the minimum overshoot calculation process in Fig. 7. If ΔV_o is smaller than ΔV_{om} , the system needs to redesign the parameters. If ΔV_o is larger than ΔV_{om} , then the closed-loop transfer function is derived, and the minimum weighing factor ratio W_{rm} can be obtained. Finally, the weighing factor ratio W_r is selected to be larger than W_{rm} and updated in the MPC controller. In summary from Fig. 8, the proposed method initiates by establishing a clear relationship between the parameters of the MPC controller and the system's output overshoot. This approach offers effective and flexible design capabilities for the overshoot structurally, allowing the system to achieve specific target values. Furthermore, owing to the absence of supplementary constraints and conditions during the design process, this method exhibits versatility and applicability across diverse systems.

C. Validation of the Proposed Overshoot Design Method

To support the abovementioned analysis, Fig. 9 shows the simulations utilizing the parameters in Table II with different weighting factor ratios. As observed in the figure, when adopting a weighting factor ratio of $W_r = 0.6$, the overshoot of the output voltage is 1.7 V. Based on the analysis of the relationship between cutoff frequency and weighting factors ratio in Fig. 6, the overshoot should be $\Delta V_o = 1.5/(2\pi * 70 * 2000 * 10^{-6}) = 1.7$ V. As seen in the figure, when adopting a weighting factor ratio of $W_r = 0.7$, the overshoot of the output voltage is 1.6 V. Based on the analysis of the relationship between cutoff frequency and weighting factors ratio in Fig. 6, the overshoot should be $\Delta V_o = 1.5/(2\pi * 80 * 2000 * 10^{-6}) = 1.5$ V. When adopting the weighting factor ratio of $W_r = 1$, the overshoot of the output voltage is 1.1 V. From the bode plot analysis, the overshoot should be $\Delta V_o = 1.5/(2\pi * 110 * 2000 * 10^{-6}) = 1.2$ V. Next, when adopting the weighting factor ratio of $W_r = 1.2$, the overshoot is 0.95 V. From the bode plots analysis, the overshoot should be $\Delta V_o = 1.5/(2\pi * 120 * 2000 * 10^{-6}) = 1$ V. Finally, due to the limitations of the adjusting ability, when adopting the weighting factor ratio of $W_r = 2$, the output overshoot equals to that when adopting $W_r = 1.2$.

From the simulation results in Fig. 9, it can be concluded that the proposed method can be utilized to design the output voltage overshoot. Adjusting the weighting factor ratio can modify the system's behavior to meet specific design requirements. In addition, it is essential to consider the control system's adjusting ability limitations and the dynamic process characteristics.

IV. EXPERIMENTAL RESULTS

To validate the analysis, a dc–dc boost converter system has been built in the lab. The dc source used in the experiment is a Delta Elektronika SM 600–10 dc power supply. The MPC algorithm is implemented using the dSPACE DS1202 board. In addition, a PWM generation is used to generate a 10 kHz sawtooth for generating the desired PWM. The experimental prototype consists of an LC filter followed by the boost converter, as shown in Fig. 10. Table II gives the system's parameters and the control parameters used in the experiments.

A. Case Study 1: Input Voltage Stability Validation With MPC Algorithm

A comparative evaluation was conducted to demonstrate the necessity of controlling the input port voltage V_{in} in the system. In the first test, the MPC algorithm only controlled the inductor current of the LC filter connected boost converter. As predicted, the converter, which behaves as a CPL, led to oscillations in the system, as shown in Fig. 11(a).

In the second test, the MPC algorithm was used, which also controls the input voltage. As a result, the input voltage was stabilized, as shown in Fig. 11(b). These results clearly demonstrate that controlling both the input voltage and the inductor current is essential for achieving a stable performance in the studied system.

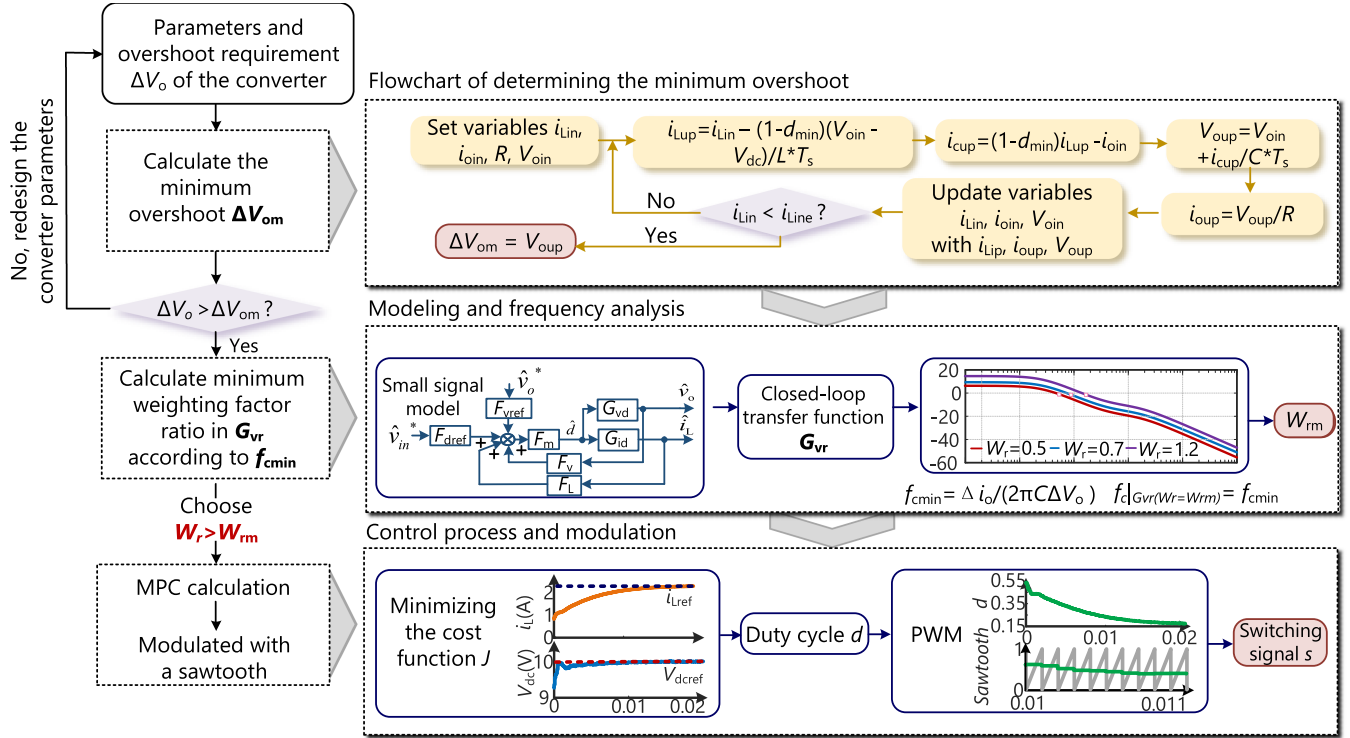
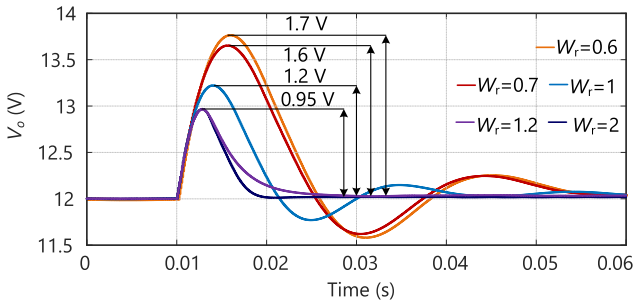


Fig. 8. Flowchart of the overshoot design process of the dc-dc boost converter.


 Fig. 9. Output voltage using different weighting factors ratio W_r during dynamic process with the parameters in Table II.

B. Case Study 2: Output Voltage Overshoot With Different Weighting Factor Ratios

This case study validates the proposed method when the load changes using different weighting factor ratios. As discussed in Section III, when the weighting factor ratio W_r is larger than 0.7, the overshoot can be limited to 1.5 V. First, the open loop dynamic process is presented in Fig. 12. As seen, the overshoot and the undershoot during the load step are approximately 2.1 and 2 V, respectively.

Fig. 13 shows the waveforms of the system with a weighting factor ratio $W_r = 0.6$. Compared with the open loop, the overshoot is reduced to 1.7 V. Fig. 14 provides the waveforms of the condition that $W_r = 0.7$, where the output overshoot equals 1.6 V approximately. From the analysis and simulation

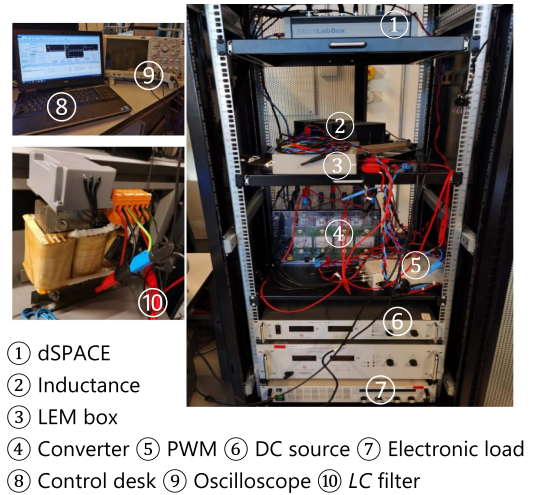


Fig. 10. Experimental setup for dc-dc converter.

results in Section III, the overshoot derived is 1.6 V in this case. Fig. 15 provides the waveforms of the condition that $W_r = 0.8$, where the output overshoot equals to 1.5 V approximately, and the result is approximately corresponding to the analysis in Section III. To obtain a smaller overshoot, W_r is further increased in Fig. 16 which equals 1. As seen, the overshoot is decreased to 1.2 V. Finally, when W_r is increased significantly to 2 in Fig. 17, the overshoot is decreased to 1 V. Because of the duty cycle limitation, the overshoot cannot be further decreased.

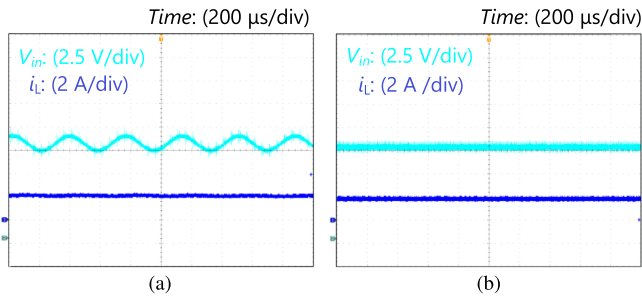


Fig. 11. Input voltage V_{in} and inductor current i_L of the boost converter. (a) Without controlling the input voltage. (b) With controlling the input voltage.

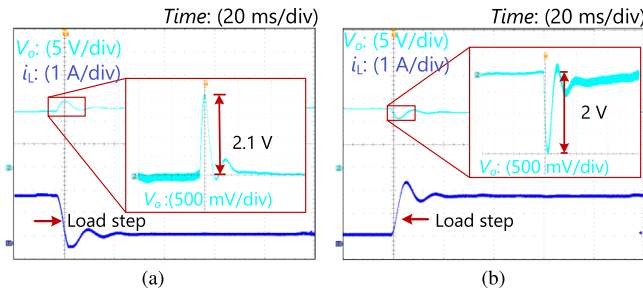


Fig. 12. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with open loop. (a) Overshoot of the output voltage when the load current steps down to 0.5 A. (b) Undershoot of the output voltage when the load current steps up to 2 A.

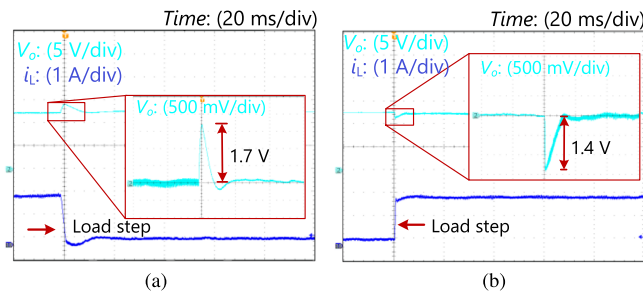


Fig. 13. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with $W_r = 0.6$. (a) Overshoot of the output voltage when the load current steps down to 0.5 A. (b) Undershoot of the output voltage when the load current steps up to 2 A.

C. Case Study 3: Output Voltage Overshoot (Undershoot) With Different Load Changes and Input Voltages

In real applications, the input voltage and load may vary, and it is important to ensure that the system performs well under all different conditions. The first case considers load changes from 0.6 to 1.5 A. The second case considers when the input voltage is 9 V, the load changes from 0.5 to 2 A.

Fig. 18 shows the results for load changes from 0.6 to 1.5 A and 1.5 to 0.6 A with the weighting factor ratio $W_r = 2$. As seen, the overshoot and undershoot remain below 1.5 V, with values of approximately 0.5 V, respectively.

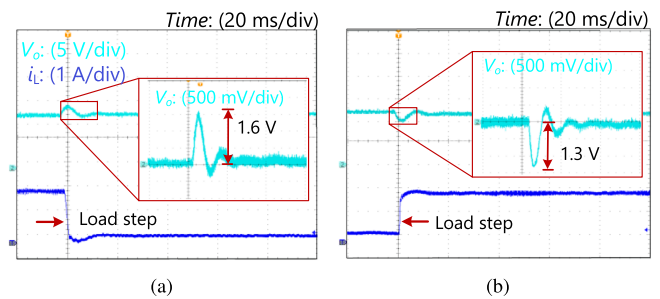


Fig. 14. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with $W_r = 0.7$. (a) Overshoot of the output voltage when the load current steps down to 0.5 A. (b) Undershoot of the output voltage when the load current steps up to 2 A.

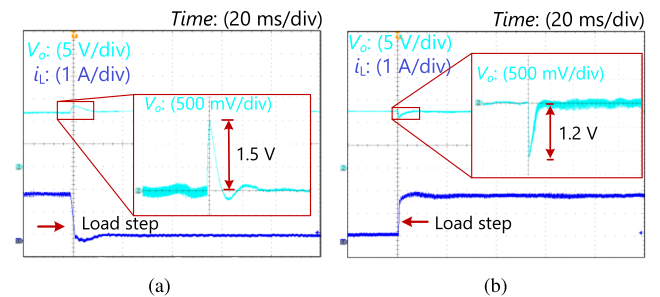


Fig. 15. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with $W_r = 0.8$. (a) Overshoot of the output voltage when the load current steps down to 0.5 A. (b) Undershoot of the output voltage when the load current steps up to 2 A.

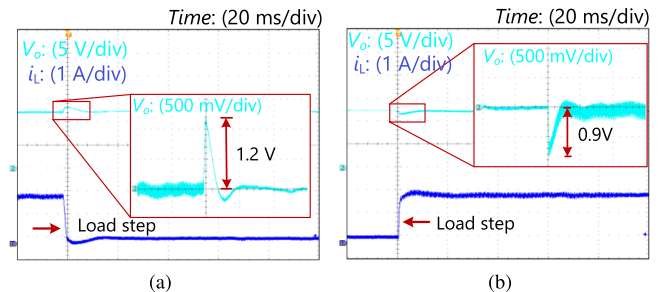


Fig. 16. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with $W_r = 1$. (a) Overshoot of the output voltage when the load current steps down to 0.5 A. (b) Undershoot of the output voltage when the load current steps up to 2 A.

Fig. 19 presents the results for an input voltage of 9 V with load changes from 0.5 to 2 A and a weighting factor ratio of $W_r = 2$. As seen, the overshoot is 1.2 V, which also satisfies the overshoot requirement.

In practice, the load characteristics may not always be purely resistance. To demonstrate the applicability of the proposed method in the studied system with different load characteristics, a validation test involving a CPL is conducted, as shown in Fig. 20. The power of the CPL is set to 12 W, and the output voltage undergoes a step change from 12 to 20 V and leads to

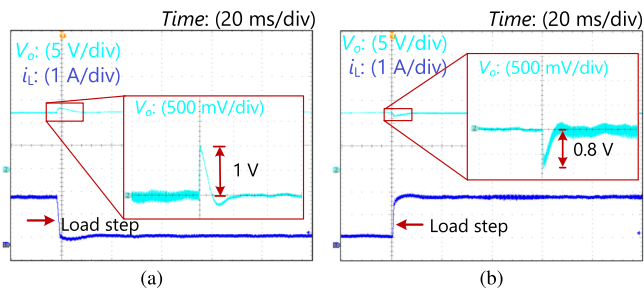


Fig. 17. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with $W_r = 2$. (a) Overshoot of the output voltage when the load current steps down to 0.5 A. (b) Undershoot of the output voltage when the load current steps up to 2 A.

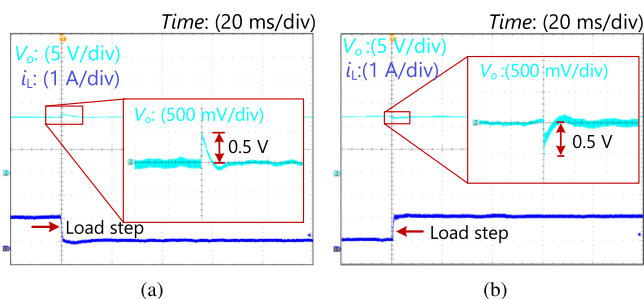


Fig. 18. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with $W_r = 2$. (a) Overshoot of the output voltage when the load current steps down to 0.6 A. (b) Undershoot of the output voltage when the load current steps up to 1.5 A.

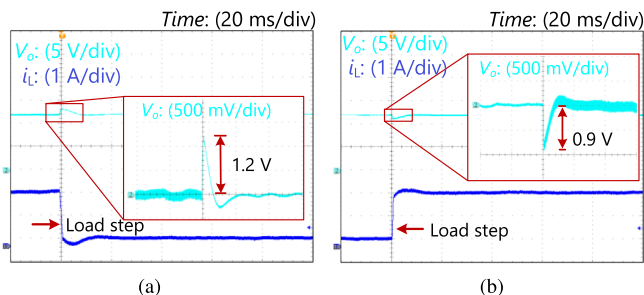


Fig. 19. Output voltage overshoot and undershoot ΔV_o and the inductor current i_L of the boost converter with $W_r = 2$ and the input voltage $V_{in} = 9$ V. (a) Overshoot of the output voltage when the load current steps down. (b) Undershoot of the output voltage when the load current steps up.

a 0.4 A load current variation. To meet the specified overshoot requirement of 1.5 V, a weighting factor $W_r = 1$ is selected.

Upon analysis, it is evident that during dynamic processes caused by the changing output reference, the overshoot (undershoot) is approximately 1.1 V, effectively controlled well within the desired 1.5 V threshold. This successful outcome demonstrates the efficacy of the proposed method in maintaining stable and accurate control of the system under varying load conditions.

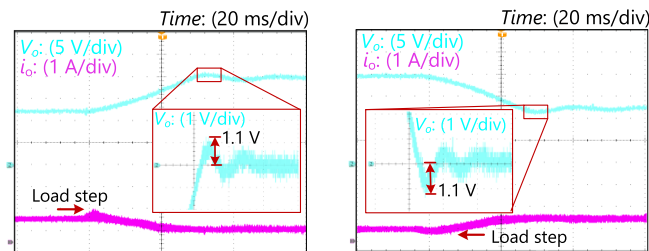


Fig. 20. Output voltage overshoot and undershoot ΔV_o and the output current i_o of the boost converter with $W_r = 1$ and the load power is 12 W. (a) Overshoot of the output voltage when the load current steps down to 0.6 A. (b) Undershoot of the output voltage when the load current steps up to 1 A.

In summary, the results validate that the proposed design method is capable of satisfying the dynamic overshoot requirement under various operating conditions. Specifically, the system exhibits good performance when the load changes, as well as when the input voltage varies to 9 V. In all cases where W_r exceeds 0.7, the overshoot remains consistently below 1.5 V, with values ranging from approximately 1 to 1.5 V, which are in strict compliance with the design requirement.

V. CONCLUSION

In conclusion, this article presented a novel overshoot design method for the MPC controller and validated it with a boost converter system. Based on the proposed method, a minimum weighting factor ratio could be derived for the MPC controller to satisfy the overshoot specification.

The experimental results demonstrated the effectiveness of the proposed design method with a boost converter system by limiting the overshoot during the load current changes with 1.5 A. It proves that as long as the weighting factor ratio W_r is greater than the minimum weighting factor ratio $W_{rm} = 0.7$, the overshoot was less than 1.5 V. Using a larger weighting factor ratio W_r , the overshoot could be further reduced within its adjusting ability. Moreover, the flexible adjustment of the overshoot was realized by simply changing the weighting factor ratio to meet different overshoot design requirements.

Compared with other methods for enhancing the dynamic performance of MPC-controlled systems, the proposed method was intrinsic, did not require any training, and provided flexibility for overshoot regulation, making it applicable to a wide range of MPC-controlled converters.

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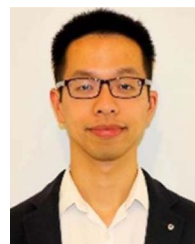
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