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Observer-Based Backstepping Sliding Mode Control Design for Microgrids Feeding a Constant Power Load

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***Abstract*— This paper deals with the problem of controller design for DC microgrids that feed constant power loads. To design the proposed controller, first by the use of the exact feedback linearization approach, the linear model of Brunovsky's canonical representation of the system has been obtained to address the nonlinearity problem of the system. Then, the desired control technique is developed by a combination of sliding mode and backstepping control approaches in which a nonlinear disturbance observer is utilized to estimate the disturbance. The overall stability of the system is analyzed based on the Lyapunov approach. A suitable and practical sliding surface is one of the controller strengths that allow the bus voltage to track the reference voltage with high accuracy and fast transient response. Finally, to prove the mentioned claims, an experimental setup has been constructed and the proposed controller is implemented. The experimental results have been analyzed and error analysis is performed. The results confirm the superiority of the proposed controller compared to state-of-the-art controllers.**

Index Terms— Backstepping sliding mode control, Nonlinear disturbance observer, Constant power load, Boost converter.

I. INTRODUCTION

Environmental problems and restrictions on the supply of fossil fuels have led to the growth and development of microgrids in modern power systems. Advances in renewable energy technologies, the development of innovations in power electronics, and government support for renewable energy rates have led electrical companies to develop and expand distributed generation. Moreover, proximity to the place of consumption due to the development of microgrids has been expressed as a suitable solution to mitigate the problems of traditional power networks [1-3].

Some loads are inherently constant power load (CPL), however in some cases, due to the type of operation and control of converters, these devices will behave like a CPL [4]. For example, using a multi-converter in a cascade structure and applying a tight control to the load converter, even when the actual network loads are inductive or resistive, the multi-converter system will tend to act as a CPL. This behavior makes a negative incremental impedance in the input terminal, which negatively affects the stability and voltage quality of the system. In other words, in these conditions, the system represents a non-minimum phase characteristic that is difficult to control. Accordingly, the primary cause of instability is the negative impedance phenomenon caused by the CPLs while the nonlinearity of the system complicates the design of the controller. Therefore, the issues of voltage regulation and stability of microgrids in the presence of CPLs are essential challenges of future power networks [5]. In case of negligence or failure to address these issues, instability and blackouts have spread to a higher level, where the possibility of the collapse of the whole system is not far from the mind and imagination [6].

Moreover, the presence of disturbance in industrial environments is a matter that should be considered along with sudden changes in load and input voltage of the power source as factors affecting the stability of the system [7]. Ensuring the system stability, improving the dynamic response, and limiting its overshoot in the presence of a disturbance, sudden load changes, and input voltage changes for the boost converter in the DC microgrid is the main goal of the controller design.

So far, different methods have been developed to address the effect of negative impedance of CPLs, each of which has advantages and disadvantages. Eliminating this effect through passive damping is a practical and straightforward way in which the damping of the system is increased by adding passive elements [8]. This method is known as a simple method that stabilizes the system without the need to change or redesign the control loop [9]. Active damping is another

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method that is performed by adding equipment, which injects a compensatory current into the circuit and by modifying the control loop [10].

The pole placement control technique is utilized to move the system's pole to the left side of the imaginary axis. The new location of the pole should be chosen so that the overall closed-loop system be stable [11].

Moreover, the pulse adjustment method, which is based on providing digital control through pulse control and PWM guidance, is used to overcome this issue [12]. It should be noted that these techniques may adversely affect the dynamic response of the system. In addition, since the performance of these methods is based on the small-signal model, in practice, these methods can only guarantee the stability of the system to a limited extent and near the points of operation [13]. Therefore, these control methods are valid only locally and in case of significant disturbance, the control method is ineffective, and the system will be unstable. A DC microgrid in the presence of a CPL is a nonlinear switching system, therefore, the linear controller faces many limitations in dealing with this condition [14]. Accordingly, the design of a nonlinear controller is essential to stabilize these systems.

Recently, the Exact Feedback Linearization (EFL) method has been successfully used to turn nonlinear systems into linear ones. Compared to other linearization methods, the non-elimination of non-linear high-order terms of the system's linearization process is one of the main advantages of this method [15]. As a result, the EFL technique has been widely used in the design of power conversion control systems [16,17]. In [14], AC/DC three-phase converter voltage control is implemented using the EFL technique. However, many of these controllers are designed by the assumption that the nonlinear model of the system is accurate while uncertainties or external disturbances are not considered. Moreover, this technique cannot be used directly for systems with unstable zero dynamics [18]. Therefore, when the system does not meet the exact linearization conditions, the zero dynamic stability must be considered and checked. Otherwise, the system may represent a non-minimum phase characteristic and exhibit harmful behavior [19]. This negatively affects the dynamic quality of the system and increases transient dynamics. In [18], based on EFL theory, the system is converted into a non-minimum phase system with the help of small-signal inductor current injection at the output and a suitable controller is designed.

In [20], a controller has been designed for the buck converter system with CPL to improve the dynamic performance of the system by using EFL with a combination of feedback and feedforward control methods. In [21], a robust control system has been designed using Kalman filter and feedback control. In [22, 23] a model predictive control method has been used to stabilize the boost converter in the presence of CPL. It should be noted that the high computational burden of this controller limits the practical implementation.

If the zero dynamics of the system are unstable or the mathematical model of the system is not accurate, to ensure the proper performance of the controllers in different conditions, it is necessary to use robust controller design methods. To this aim, sliding mode control is used to deal with external perturbations, unmodeled dynamics, and parameter uncertainty [24-25].

Robust sliding mode control of bandwidth modulation ensures the stability of the systems by utilizing a new sliding surface [26]. Although the adverse effects of uncertainty of system parameters are limited and controlled by sliding model control, fixed sliding mode coefficients cause a strong chattering phenomenon in the control signal and cause severe fluctuations. In [27], a sliding mode controller for a buck converter system with a CPL at a constant switching frequency is proposed. By changing the input voltage and changing the level of load consumption, different operating points are created. However, the stability of the system in other conditions is well guaranteed by this controller. Nevertheless, the requirement to measure different values causes limitations in its applications.

The adaptive backstepping technique is one of the most effective methods for designing systematic nonlinear control in stabilization problems which has been widely employed for the purposes of converter stabilization [28]. However, the load is supposed to be resistive and the stability analysis of the converters in the presence of a CPL is neglected.

An effective technique to estimate uncertainties and disturbances in nonlinear systems is the nonlinear disturbance observer (NDO) [29,30]. Recently, a backstepping sliding control technique is used to deal with disturbances and possible changes in parameters [31]. Large-signal stability and step-by-step design are some advantages of this method.

In this paper, in order to focus on the subject of controller design at the primary level, a simplified model of microgrids in the islanded mode in the form of a source, a converter, and power loads has been employed. In the following, it has been tried to solve the problem of microgrid instability due to the CPL by combining two control techniques, i.e., sliding mode control (SMC) and backstepping mode control (BMC), while an NDO is employed to estimate the uncertainties and disturbances. The Lyapunov-based approach is utilized to derive the stability condition. Therefore, the contributions of this paper can be outlined as:

1) By using a combination of two types of controllers, the advantages of both are adopted and their weaknesses are covered.
2) By using the NDO, it is possible to neutralize the adverse effects of uncertain CPLs.
3) The need to change the output of the state equations to implement the EFL technique has caused the output voltage to be indirectly controlled and faced with inertia. Using a combination of two controllers with an NDO and using a suitable sliding surface has solved the problem of indirect control and inertia in the system.
4) Due to the use of a suitable sliding surface and an increase in the degree of freedom in the design coefficients, the construction of the controller has become more accessible and possible.

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The rest of the article is organized as follows: Section 2 describes the basic principles and problem statement. In Section 3, the proposed controller is designed and proven to be stable. In Section 4, experimental results and comparative graphs are presented. Finally, the conclusion is mentioned in Section 5.

II. Basic Principles and Problem Statement

Boost and buck converters are used to change the voltage level using components such as inductors, capacitors, diodes, and controlled switches. Changing the connection method and component values can reduce or increase the voltage and create other properties of converters.

A simplified scheme of the DC power distribution system is shown in Fig. 1. As shown, a boost converter transfers power from the voltage source to the CPL and the resistive loads. To model the CPL, a dependent current source at the output is used. However, due to the changes in current and instantaneous voltage of the CPL, we are faced with a negative impedance. Negative impedance is the leading cause of instability in such systems. If the output voltage decreases due to disturbance, the negative impedance characteristic further reduces the output voltage, and if a proper controller is not designed, this will eventually make the system unstable and cause the voltage to collapse.

Consider the following averaged model for the continuous-conduction case of the boost converter shown in Fig 1:

$$\begin{cases} L\dfrac{di_l}{dt} = V_{in} - (1-u)v_c \\ C\dfrac{dv_c}{dt} = (1-u)i_L - \dfrac{v_c}{R} - \dfrac{P_{CPL}}{v_c} \end{cases} \tag{1}$$

where the common notations have been used. The goal is to regulate the output voltage v_C to its setpoint v_{Cref}. Based on the location of v_C, the system is nonlinear. However, by considering another output, the system becomes completely linear, which will be discussed in the next section.

III. Controller Design

To prepare the form of state equations, especially for the implementation of the backstepping sliding mode control (BSMC), a form of equations called EFL must be obtained.

An appropriate output is selected to overcome the non-minimum phase characteristic of the system, which results in indirect control of the capacitor voltage. Accordingly, the overshoot of the voltage is increased. Then, by using the Brunovsky transform, the canonical form of the system is obtained, and by applying feedback, the linear form of the system is obtained. Finally, microgrid instability in the presence of a CPL is addressed by employing a combination of sliding mode control and backstepping mode control. To achieve the goal, s suitable Lyapunov candidate function is employed to guarantee the stability of the system. Taking the advantage of these two types of controllers and the use of NDO will allow the final controller to operate in industrial environments.

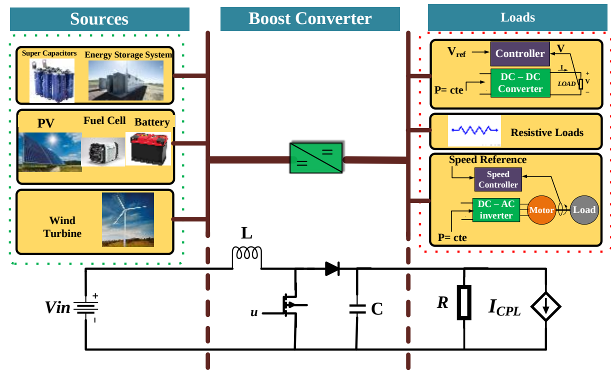

Fig. 1: Typical DC power distribution system and its equivalent circuit.

By examining the possibility of complete linearization through the Lee Bracket derivative criterion, it is determined that if the capacitor voltage is selected as the output, the system is partially linear. However, by selecting the system energy as the output, the system becomes completely linear. It should be noted that in the new situation, the capacitor voltage control is indirectly possible, and adverse effects such as overshoot and slow control speed occur.

Assume γ_1 is the sum of the energy stored in the system:

$$\gamma_1 = \frac{1}{2}Li_l^2 + \frac{1}{2}Cv_c^2 \tag{2}$$

taking derivation from it, yields:

$$\dot{\gamma}_1 = V_{in}i_l - \frac{v_c^2}{R} - P_{CPL} \tag{3}$$

Consider the second state variable γ_2 as:

$$\gamma_2 = V_{in}i_l - \frac{v_c^2}{R_0} \tag{4}$$

where R_0 is the nominal resistance of the resistive load. Taking derivation from γ_2, yields:

$$\dot{\gamma}_2 = \frac{{V_{in}}^2}{L} + \frac{2v_c^2}{R_0^2C} - \left(\frac{V_{in}v_c}{L} + \frac{2i_lv_c}{R_0C}\right)(1-u) + \frac{2}{R_0C}\left(P_{CPL} - \frac{v_c^2}{R_0} + \frac{v_c^2}{R}\right) \tag{5}$$

Now, define the auxiliary variable d, disturbances h_1 and h_2 as follows:

$$d = \frac{{V_{in}}^2}{L} + \frac{2v_c^2}{R_0^2C} - \left(\frac{V_{in}v_c}{L} + \frac{2i_lv_c}{R_0C}\right)(1-u) \tag{6}$$

$$h_1 = -P_{CPL} + \frac{v_c^2}{R_0} - \frac{v_c^2}{R} \tag{7}$$

$$h_2 = \frac{2}{R_0C}\left(P_{CPL} - \frac{v_c^2}{R_0} + \frac{v_c^2}{R}\right) \tag{8}$$

Then, the canonical form of the system is achieved by using the following definitions:

$$\begin{cases} \dot{\gamma}_1 = \gamma_2 + h_1 \\ \dot{\gamma}_2 = d + h_2 \end{cases} \tag{9}$$

To find out the reference values of the new state variables, define:

$$\gamma_1^* = \frac{1}{2}Li_{ref}^2 + \frac{1}{2}Cv_{cref}^2 \tag{10}$$

$$P_{ref} = \frac{v_{cref}^2}{R_0} - \hat{h}_1 \tag{11}$$

where $\hat{h}_1$ is the estimate of h_1 and is estimated using the following disturbance observers:

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$$\begin{cases}\hat{h}_1 = b_1(\gamma_1 - q_1) \\ \dot{q}_1 = \gamma_2 + \hat{h}_1\end{cases} \tag{12}$$

and

$$\begin{cases}\hat{h}_2 = b_2(\gamma_2 - q_2) \\ \dot{q}_2 = \theta + \hat{h}_2\end{cases} \tag{13}$$

where q_1 and q_2 is are auxiliary variables, and b_1 and b_2 are positive design coefficients of the observers, and $\theta = -k_2\varphi_2 - \hat{h}_2 + \ddot{\gamma}_1$.

Achieving a robust and adaptive controller is the result of combining two types of controllers. The initial steps are taken through BMC, and in the final step, SMC is used.

Now, define the new error variables φ_1 and φ_2, as follows:

$$\begin{cases}\varphi_1 = \gamma_1 - \gamma_1^* \\ \varphi_2 = \gamma_2 - \gamma_2^*\end{cases} \tag{14}$$

The goal is to die out the error between γ_1 and γ_1^*.

$$\dot{\varphi}_1 = \varphi + \gamma_2^* + h_1 - \dot{\gamma}_1^* \tag{15}$$

Now, by selecting the Lyapunov function as follows:

$$V_1 = \frac{1}{2}\varphi_1^2 + \frac{1}{2}\tilde{h}_1{}^2 \tag{16}$$

where $\tilde{h}_1 = h_1 - \hat{h}_1$, now, taking derivation from it, yields:

$$\dot{V}_1 = \varphi_1\dot{\varphi}_1 + \tilde{h}_1\dot{\tilde{h}}_1 \tag{17}$$

By selecting γ_2^*, we have:

$$\gamma_2^* = -k_1\varphi_1 - h_1 + \dot{\gamma}_1^* \tag{18}$$

According to the definition of the disturbance observer:

$$\dot{V}_1 = -k_1\varphi_1^2 + \varphi_1\varphi_2 + \varphi_1\tilde{h}_1 + h_1\dot{h}_1 - b_1\tilde{h}_1^2 \tag{19}$$

Now, taking derivation from (13), yields:

$$\dot{\varphi}_2 = \dot{\gamma}_2 - \dot{\gamma}_2^* \tag{20}$$

In this stage, consider the following Lyapunov function:

$$V_2 = V_1 + \frac{1}{2}\delta^2 + \frac{1}{2}\tilde{h}_2^2 \tag{21}$$

where $\tilde{h}_2 = h_2 - \hat{h}_2$. The reaching phase is achieved in the last stage for the proposed BSMC, now by selecting a proper sliding surface the sliding phase is accomplished.

Using the error dynamic is a common way to define the sliding surface. In order to improve the controller efficiency and increase the degree of freedom for the designer, a combination of the existing errors is suggested. It is suggested to select a dynamic sliding surface based on the linear combination of errors, φ_1 and φ_2, as follows [31]:

$$\delta = M\varphi_1 + \varphi_2 \tag{22}$$

Taking derivation from this sliding surface, yields:

$$\dot{\delta} = -\alpha\delta - \eta sgn(\delta) \tag{23}$$

Taking derivation from (21), yields:

$$\dot{V}_2 = \dot{V}_1 + \delta\dot{\delta} + h_2\dot{h}_2 \tag{24}$$

where
$$\delta\dot{\delta} = \delta[(M + k_1)(\varphi_2 - k_1\varphi_1) + (m + k_1)\tilde{h}_1 + d + h_2 + \dot{\tilde{h}}_1 - \ddot{\gamma}_1^*] \tag{25}$$

and

$$d = -(M + k_1)(\zeta_2 - k_1\zeta_1) - \dot{\hat{h}}_2 + \ddot{\gamma}_1^* - \alpha\delta - \eta sgn(\delta) \tag{26}$$

According to the (6) and (26), we have:

$$u = 1 - \frac{\frac{V_{in}{}^2}{L} + \frac{2v_C^2}{R_0^2C} - d}{\frac{V_{in}v_C}{L} + \frac{2i_lv_C}{R_0C}} \tag{27}$$

Substituting the NDO relations and (19) into (25), yields:

$$\dot{V}_2 = -k_1\varphi_1^2 + \varphi_1\varphi_2 - \alpha\delta^2 + \varphi_1\tilde{h}_1 + \tilde{h}_1\dot{h}_1 - b_1\tilde{h}_1^2 - b_2\tilde{h}_2^2 + \tilde{h}_2\dot{\tilde{h}}_2 + \delta\big[(M + k_1 + b_1)\tilde{h}_1 + h_2\big] - \eta|\delta| \tag{28}$$

Assuming that the disturbances are limited, yields:

$$\begin{cases}|\tilde{h}_1| \le |\bar{h}_1| \\ |\tilde{h}_2| \le |\bar{h}_2|\end{cases} \tag{29}$$

$$\begin{cases}(M + k_1 + b_1)\tilde{h}_1 + \tilde{h}_2 \le (M + k_1 + b_1)\bar{h}_1 + \bar{h}_2 \\ (M + k_1 + b_1)\bar{h}_1 + \bar{h}_2 = \bar{\Omega}\end{cases} \tag{30}$$

$$(M + k_1 + b_1)\tilde{h}_1 + h_2 \le \bar{\Omega} \tag{31}$$

By defining:

$$\begin{cases}\varphi^T = [\varphi_1 \quad \varphi_2] \\ \mu = \begin{bmatrix}(k_1 - \frac{1}{2}) + \alpha M^2 & \alpha M - \frac{1}{2} \\ \alpha M - \frac{1}{2} & \alpha\end{bmatrix}\end{cases} \tag{32}$$

We have:

$$\varphi^T\mu\varphi = \left((k_1 - \frac{1}{2}) + \alpha M^2\right)\varphi_1^2 + (2\alpha M - 1)\varphi_1\varphi_2 + \alpha\varphi_2^2 \tag{33}$$

Using relations (29) to (33), we have:

$$\dot{V}_2 \le -\varphi^T\mu\varphi - \frac{1}{2}\varphi_1^2 + \varphi_1\tilde{h}_1 + \tilde{h}_1\dot{h}_1 + \tilde{h}_2\dot{h}_2 - b_1\tilde{h}_1^2 - b_2\tilde{h}_2^2 - (\eta - \bar{\Omega})|\delta| \tag{34}$$

Lemma 1: Assume that the coefficients m and n are positive, then, for any positive number ζ, the following inequality is held:

$$|\mu|^m|\omega|^n \le \frac{m}{m+n}\zeta|\mu|^{m+n} + \frac{n}{m+n}\zeta^{-\frac{m}{n}}|\omega|^{m+n} \tag{35}$$

Based on this Lemma one can conclude that:

$$\begin{cases}\varphi_1\tilde{h}_1 \le \frac{1}{2}\varphi_1^2 + \frac{1}{2}\tilde{h}_1^2 \\ \tilde{h}_1\dot{h}_1 \le \frac{1}{2}\tilde{h}_1^2 + \frac{1}{2}\dot{h}_1^2 \\ \tilde{h}_2\dot{h}_2 \le \frac{1}{2}\tilde{h}_2^2 + \frac{1}{2}\dot{h}_2^2\end{cases} \tag{36}$$

Combining relations (34) and (36), yields:

$$\dot{V}_2 \le -\varphi^T\mu\varphi + \frac{1}{2}\dot{h}_1^2 + \frac{1}{2}\dot{h}_2^2 - (b_1 - 1)\tilde{h}_1^2 - (b_2 - .5)\tilde{h}_2^2 - (\eta - \bar{\Omega})|\delta| \tag{37}$$

Choosing the proper coefficients as follows:

$$\begin{cases}b_1 > 2.5 \\ b_2 > 2\end{cases} \tag{38}$$

$$\begin{cases}\|h_1\| \ge \|\dot{h}_1\| \\ \|\tilde{h}_2\| \ge \|\dot{h}_2\|\end{cases} \tag{39}$$

By combining relations (37), (38), and (39), we have:

$$\dot{V}_2 \le -\varphi^T\mu\varphi - \tilde{h}_1^2 - \tilde{h}_2^2 - (\eta - \bar{\Omega})|\delta| \tag{40}$$

Now, select the following conditions:

$$\eta \ge \bar{\Omega} \tag{41}$$

$$\mu \ge 0 \rightarrow \begin{cases}(k_1 - \frac{1}{2}) + \alpha M^2 > 0 \\ ((k_1 - \frac{1}{2}) + \alpha M^2)\alpha - (\alpha M - \frac{1}{2})^2 > 0\end{cases} \tag{42}$$

Finally, it is straightforward to verify the derivative of the Lyapunov candidate function is negative, i.e., $\dot{V}_2 \le 0$.

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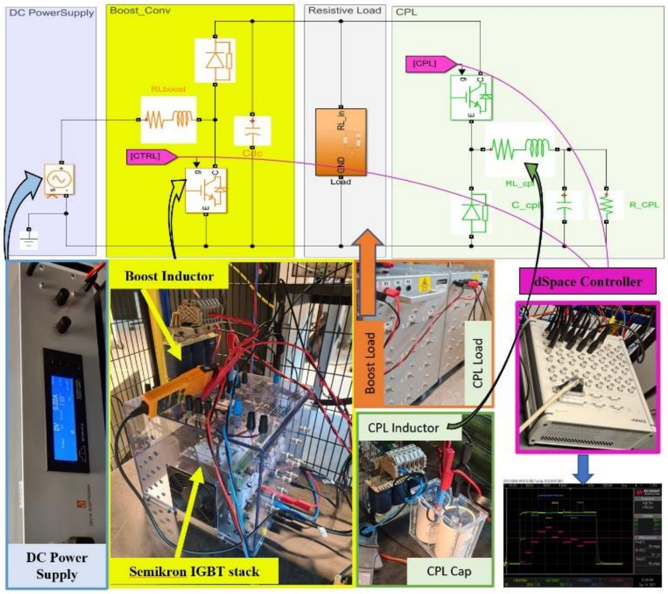

Fig. 2: The overall scheme of the experimental platform using switches of boost and buck converter (two legs of a two-level converter) to implement a CPL.

TABLE I: SYSTEM AND CONTROLLER PARAMETERS

Variables	Description	Value
V_{ref}	DC bus voltage reference	96 V
V_{in}	Converter input voltage	48 V
f_s	Switching frequency	10 kHz
L	Inductance	850 µH
C	Capacitance	1100 µF
R	Resistance	40 Ω
b_1, b_2	NDO gains	2000,2000
M, k_1	controller gains	1200, 300
α, η	controller gains	2000, 1500

IV. EXPERIMENTAL RESULTS

The main aim of this section is to evaluate the efficiency of the proposed method in practice. To this end, an experimental testbed is established as shown in Fig. 2, and the ability of the proposed controller is evaluated in different conditions.

As illustrated in Fig. 2, the experimental setup of a boost converter with resistive load and CPL comprised of a buck converter with resistive load is implemented using Semikron IGBT stack (SEMITEACH B6U+E1CIF+B6CI) and passive components. The dSPACE MicroLabBox with DS1202 PowerPC DualCore 2-GHz processor board is adopted to control the duty cycle of the switch and generate associated PWM for both boost and CPL converters such that the desired output voltage is tracked. The electrical parameters for a typical boost converter are listed in Table I.

In this study, three types of controllers have been used to compare and evaluate the ability of the proposed controller.

a) The first controller is a Backstepping Mode Control with NDO (BMC_NDO).
b) The second controller is a Backstepping Sliding Mode Control (BSMC).
c) The third controller, which is developed in the current work, is a Backstepping Sliding Mode Control with Nonlinear Disturbance Observer (BSMC_NDO).

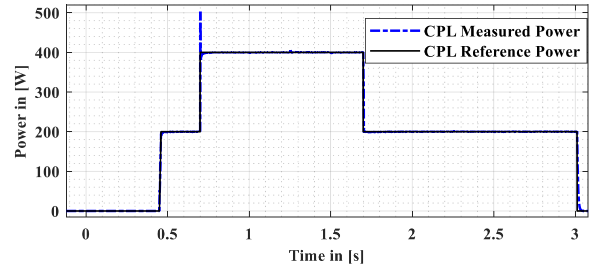

Fig. 3: The output profile of constant power load changes.

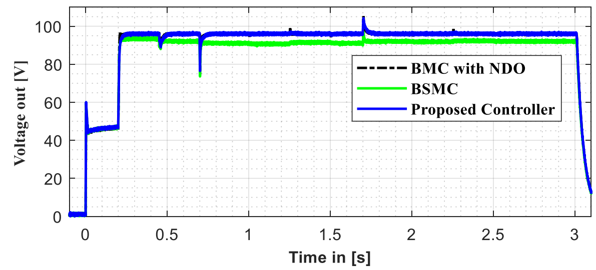

Fig. 4: The output voltage of the controller.

Based on the proof presented in Section III, the method of calculating the constraints governing the positive control coefficients $b_1 > 2.5$, $b_2 > 2$, $\alpha > 0$, $\eta > 0$, $k_1 > 0$, and $M > 0$ is shown. The values of M, α and k_1 must be such that satisfy (41). The values presented in Table I have been obtained by repeating and modifying the coefficients in order to achieve the best control response.

A. Investigation of the effect of load power changes and zero power (load resistance) on output voltage:

The primary purpose of the controller design is to regulate the output voltage at the reference value of 96 volts. To investigate the performance of the controller in contradiction of CPL changes, load changes are assumed as shown in Fig. 3 in a time interval of 3 seconds. Although the controller is designed for a DC microgrid system that feeds a CPL, different reasons cause the CPL to change and represent a resistive behavior. Therefore, the controller is supposed to demonstrate good robustness against these changes. If the constant power load becomes zero due to the parallelism of resistor R, the final power is considered as an entirely resistive load. As shown in Fig. 3, the output load is initially set at 0 watts (buck converter is off) and will be 200 Watts at 0.45 seconds. It then changes to 400 watts at 0.7 seconds and 200 watts at 1.7 seconds. As shown in Fig. 4, by evaluating the output voltage diagram, the robustness of all three controllers to power changes is determined. Fig. 5 shows the output voltage (zoom) at power change moments. As shown in Fig. 6, the difference between the output voltage and the reference value at 0.7 seconds for the proposed controller is less than other controllers. Fig. 7 shows the zoomed plot of the output voltage around 1.7 seconds when a sudden decrease occurs in the power. In this figure, the superiority of the proposed controller with less settling time and acceptable overshoot compared to other controllers is well demonstrated.

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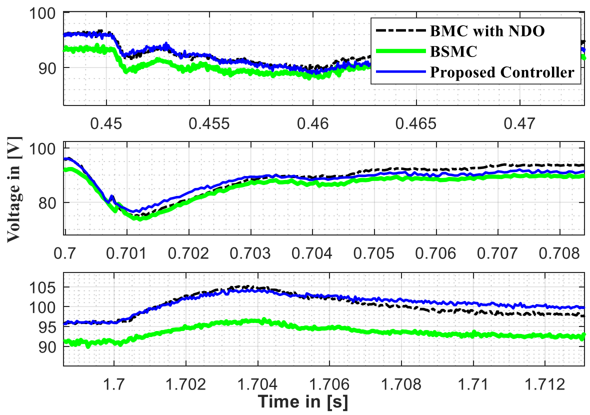

Fig. 5: The zoomed plot of the output voltage while the load changes.

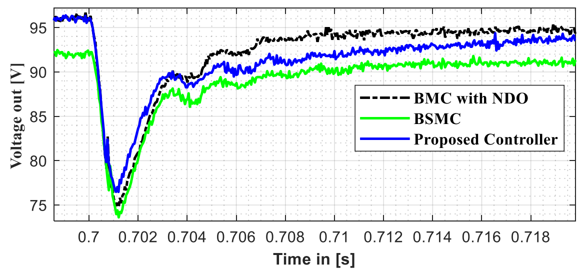

Fig. 6: The zoomed plot of the output voltage around 0.7 seconds.

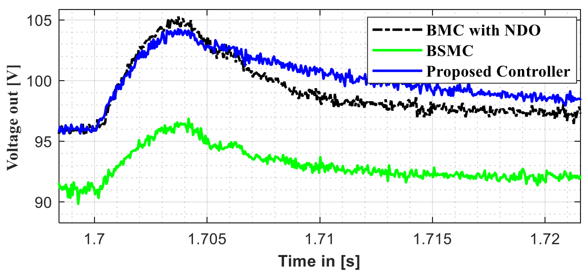

Fig. 7: The zoomed plot of the output voltage around 1.7 seconds.

The second controller fluctuates by several volts when the power changes and is practically unable to control the voltage on the reference value of 96 volts. Checking the final value of the output voltage shows that the second controller is not able to follow the reference voltage and has a residual value.

However, the proposed controller has a good performance against small power changes and resists high power changes, and the system remains stable. Moreover, the proposed controller exhibits good performance against a resistance load.

B. Investigation of the effect of input voltage change on output voltage:

The input voltage at the early step is set to 47 volts, the next 0.25 seconds is set to 42 volts, the next 0.25 seconds is set to 56 volts and this pattern is repeated once again. As illustrated in Fig. 8 when the input voltage changes, the output voltage is perfectly regulated. Fig. 9 well demonstrates the ability of the BSMC with the NDO to regulate the output voltage variations while the input voltage changes, and the grid feeds a CPL. This approves that the output voltage is well regulated and stabilized with the least acceptable fluctuations.

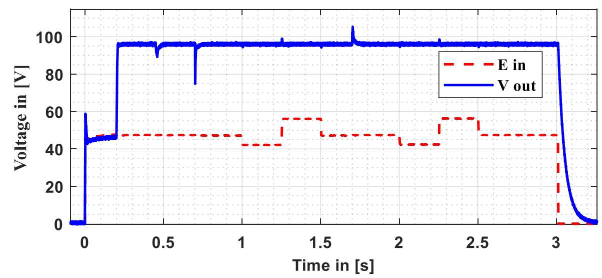

Fig. 8: The output voltage of the controller while the input voltage changes.

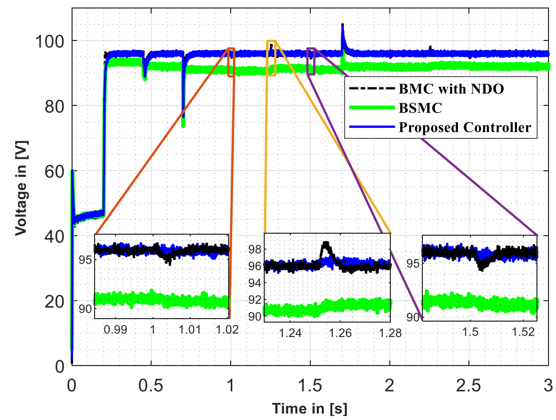

Fig. 9: The zoomed plot of the output voltage controller while the input voltage changes.

Although BSMC does not oscillate in the presence of disturbance, it is not possible to adjust the voltage to the reference value. As shown, it differs practically a few volts from the reference value. On the other hand, for BMC with NDO, although it offers acceptable voltage regulation, however, at moments of disturbance, it has more overshoot than the proposed controller. The proposed controller is more capable than other controllers in terms of performance characteristics.

C. Investigation of the effect of sliding surface type on output voltage control:

The main aim of this section is to examine the effectiveness of selecting a more complex sliding surface on voltage control. Indeed, the effectiveness of this choice on the possibility of building the controller is also a serious and undeniable challenge.

1) *The first choice:*

The sliding surface is selected as $\delta = \varphi_2$. Moreover, in this case, $\dot{\delta} = -K_1 Sgn(\delta) - K_2\delta$.

In the BSMC, with this sliding surface, it is possible to control the output voltage at an acceptable level. In this controller, the coefficients of variables and the degree of freedom are coefficients K_1 and K_2. Proper selection of these coefficients provides the possibility of appropriate control; however, three fundamental problems occur in this case.

- The ability to fine-tune the output voltage to the desired value.
- The output range of voltage when changing power.
- The ability to select the appropriate coefficient for all different operating conditions.

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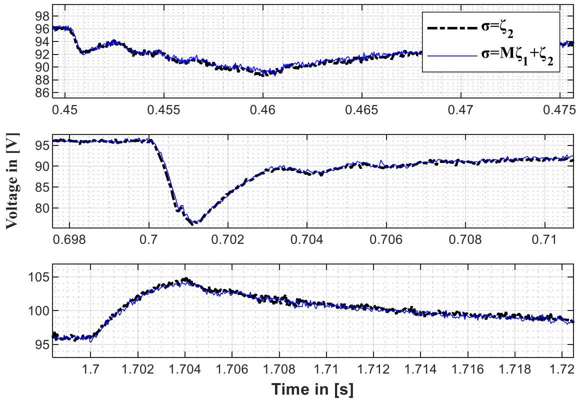

Fig. 10: Output voltage with different sliding surface for the proposed method.

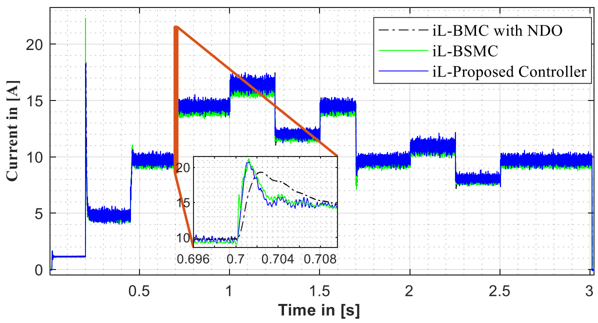

Fig. 11: Inductor current under different controls when the CPL load and input voltage change.

2) *Second choice:*

The sliding surface is selected as $\delta = M\varphi_1 + \varphi_2$, which is more complex than the simple sliding surface of the previous model, and with three degrees of freedom, including M, K_1 and K_2 adds a lot of capabilities to the controller. K_1 is the factor that helps to deal with disturbances. The coefficients M and K_2 effectively affect the output voltage fluctuations and allow tight control of the output voltage.

As shown in Fig. 10, although the use of a complex sliding surface does not appear to be effective in better controlling the output voltage, it should be noted that providing a degree of freedom in building the project has been very helpful. Using the M coefficient, it is possible to select other coefficients to an acceptable level. Moreover, if M is not used, other coefficients must be chosen so large to stabilize the system, which makes project construction almost impossible. In summary, to build a controller it is necessary to use a complex sliding surface with a further degree of freedom.

D. *Investigation of the effect of load power and voltage in changes and on inductor current:*

In order to design an appropriate control scheme, it is necessary to consider the limit of the inductor current. In Fig. 11, the inductor current when the constant power load and input voltage change is shown. As can be seen, the proposed controller compared to other controllers represents an acceptable performance. It should be noted that the advantages of the proposed controller in dealing with power and voltage changes outperform the other controllers.

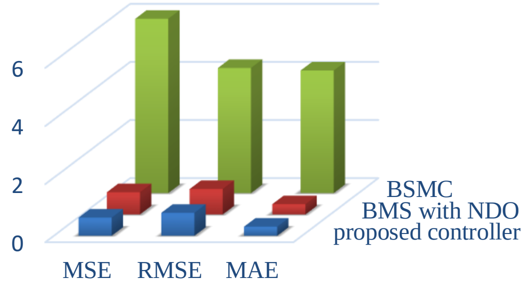

Fig. 12: Quantitative comparison of error values of the controllers.

TABLE II: ERROR ANALYSIS INDICES

	MSE	RMSE	MAE
SMC with NDO (proposed controller)	0.6271	0.7919	0.3135
BMS with NDO	0.7649	0.8746	0.3526
BSMC	18.58	4.31	4.22

E. *Error analysis*

To evaluate the ability of the proposed controller compared to state-of-the-art controllers, error analysis is performed. The data shown in Fig. 4 are analyzed using the Mean Squared Error (MSE), Root Mean Square Error (RMSE), and Mean Absolute Error (MAE) [32]. These indices are computed, and the results are plotted in Fig. 12. The method is that by changing the constant power load and input voltage, the output voltage is measured and the error indicators for each controller are analyzed separately. The results show a significant advantage of the proposed controller compared to other controllers.

V. CONCLUSION

In this paper, a novel control technique that uses a nonlinear disturbance observer is developed to deal with the effect of constant power load instability in DC microgrids. This control technique employs two different control approaches and tries to take advantage of them and cover the weaknesses of the approaches simultaneously. By comparing the output of different controllers, the effect of nonlinear disturbance observer in fine-tuning of the output voltage is quite apparent. Moreover, by choosing an appropriate sliding surface, it is possible to increase the controller adaptability level in other conditions. The experimental results have proved the idea of using several different control techniques simultaneously to achieve an efficient and straightforward method.

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