**Highlights**

- Lower performance of existing controllers for inverters-interfaced renewable energy resources
- Requirement for extra electrical equipment for improving power quality leading to air pollution
- Implementing a full-order sliding mode controller via outer voltage and inner current loops
- Achieving the desired low harmonic output voltage under uncertainties for the inverters
- Mitigating the need for extra harmonic compensators resulting in less environmental effects
First Sliding Surface: Equation (4)

\[ k_1 v_L r_{rr} e_v + \frac{d}{dt} \]

Virtual Control Law

Second Sliding Surface: Equation (19)

\[ \sum_{u_2 u_{eq2}} \]

Original Voltage Control Law

Actual Control Law: Equation (20)

Standalone System: Equations (1) & (2)

PWM Signal

Control Step 1

Control Step 2

Graphical Abstract (for review)
A Step-by-Step Full-Order Sliding Mode Controller Design for Standalone Inverter-Interfaced Cleaner Renewable Energy Sources

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Abstract

Power electronic inverters are one of the most significant components for the integration of clean energy systems and the proper control of these power electronic interfaces helps to enhance power quality. This paper presents a robust step-by-step full-order sliding mode voltage control strategy for standalone single-phase inverters that can be interfaced with cleaner renewable energy sources. The proposed controller employs a cascaded method through outer- and inner-loops to facilitate tracking of the desired load voltage. In contrast to the most commonly used sliding mode techniques for interfacing these cleaner energy sources, which usually utilize the canonical-form of the system’s state-space model, the designed controller uses a block-controllable model that mitigates unwanted noises. To design the controller, the capacitor voltage and the inductor current of the LC filter are measured to be employed in the outer second-order sliding mode voltage control loop and the inner first-order sliding mode current control loop, respectively. The utilization of such full-order sliding surfaces in each step effectively reduces the chattering in the control signals as well as facilitates the transient and steady-state responses with less harmonic distortion and thereby, promoting the effective integration of renewable energy sources. To verify the performance of the proposed controller under various loading conditions as external disturbances, a 2.2 kW standalone single-phase inverter is simulated using MATLAB/Simulink platform along with a microcontroller-based processor-in-loop through the digital signal processing. In terms of internal disturbances, the proposed controller is also tested under different filter parameters’ variations over a wider range. The results indicate a better alignment for coupling with clean energy sources and the comparisons with other control techniques show better performance of the proposed controller.

Keywords: Full-order sliding mode controller, step-by-step controller, islanded operation, robustness, inverter control

1. Introduction

1.1. Background

Climate change is one of the most pressing issues of the current era, and carbon dioxide is one of the primary factors to the rise of Earth’s temperature. There are several reasons for the generation of these devastating gases, and many countries are working through various research prospects to accomplish the goal of de-carbonization (Cunha et al., 2022). For instance, Wimalaratna et al. (2022) have conducted a comprehensive feasibility study to determine ways of developing renewable energy in Australia, especially wave power, that helps in achieving net-zero emissions. In addition, the presented research by Coulibaly et al. (2022) demonstrates that...
the energy conversion from solar to hydrogen systems and the use of the generated heat by the fuel cell through the recovery process improves the energy efficiency up to 38.5%, resulting in a higher energy efficiency than when the heat is not recycled. Regarding this, the combustion of fossil fuels in traditional power plants releases considerable amounts of greenhouse gasses into the atmosphere. The production of energy in these centralized units, which are usually located far from the consumers, also results in an increase in transmission costs and power losses, coupled with the necessity for complex infrastructure. The use of gasoline-powered vehicles further increases global warming and other environmental problems while electric vehicles are considered more environmentally friendly alternatives to net-zero emission vehicles (Lavrador and de Sá Teles, 2022). The research communities are developing a number of solutions in order to deal with these challenges, and microgrids are regarded as one of the most promising solutions in order to address these issues (Archana et al., 2022).

1.2. Key challenges and open research questions

In spite of the tremendous benefits of microgrids, they pose a number of challenges to industries from a number of different perspectives. In recent years, the integration of distributed energy resources (DERs) based on solar, wind, etc. in power systems has led to the formation of microgrid systems. Microgrids have been widely used in different parts of power systems including islanded remote and rural areas (Khajehzadeh et al., 2022). In these applications, the use of renewable energy sources such as solar and wind is an inseparable part of microgrids, particularly in remote regions (Chen and Baran, 2022). Since solar and wind exhibit intermittent characteristics, it is essential to use power electronic converters, for example, inverters are very necessary in AC microgrids in order to deliver safe power with high quality to the grid or consumers (Nejabatkhah et al., 2019). The use of these DERs can be mainly used in two modes as the grid-connected operation mode (Safia et al., 2020) and standalone operation mode (Zheng et al., 2020). In grid-connected operation mode, it is attempted to deliver the active and reactive power to the main grid. However, in islanded mode, the main purpose is the control of consumers’ voltage toward a desired value with the less total harmonic distortion (THD). During the standalone operation of DERs, single-phase inverters are one of the mainly employed interfaces and the control of these converters plays a significant role to meet the requirements of the system and other relevant standards (EN 50160, 2010). From the first view, it might be thought that what is the relevancy of power electronic interfaces and their appropriate control to the global warming and clean energy industries? To address this critical question, two main reasons can be considered. Firstly, the clean energy production without power electronics is meaningless and infeasible. Secondly, the appropriate control of these power electronic devices is very important from a number of viewpoints (e.g., power quality (Kumar and Srikanth, 2023), effective integration with consumers’ load, etc.), which is the subject of this paper.

1.3. Literature Review

In order to appropriately control the voltage of standalone inverters, different control approaches comprising linear and nonlinear methods have been presented in literature in which some of these methods are proportional-integral (PI) controller in synchronous reference frame (SRF) integrating with the resonant harmonic compensator (Monfared et al., 2014), parallel operation of the inverters with the PI controllers (Rahman et al., 2015), improved PI-derivative (PID) controllers in motor drive applications (Shoujun and Weiguo, 2011), proportional-resonant (PR) controller (Bairagi et al., 2019), hybrid PID+PR with load current feedback (PID+R+LCF) controller (Dong et al., 2011), Posicast controller (Li, 2009), repetitive controller (Byen et al., 2015), deadbeat controller (Kim et al., 2016), hybrid deadbeat and repetitive controller (Lidozzi et al., 2017), optimized linear quadratic regulator (LQR) (Ufnalski et al., 2015), robust LQR associated with the integral action (Hossain et al., 2015), implicit model predictive controller (MPC) (Nauman and Hasan, 2016), energy-function based MPC (Guler and Komurcugil, 2022), fixed switching sliding mode controller (FS-SMC) for...
islanded single-phase inverters (Abrishamifar et al., 2012), FS-SMC for four-wire three-phase islanded inverter (Pichan and Rastegar, 2017), seeking the maximum power based on SMC for the application of islanded solar systems (Chaibi et al., 2019), reduced state feedback sliding mode current controller (RSFSMCC) (Gudey and Gupta, 2015), adaptive SMC (Chen et al., 2015), SMC+$H_2/H_\infty$ (Li et al., 2017), fractional-order SMC (FOSMC) (Delghavi et al., 2016), super-twisting SMC (ST-SMC) (Barzegar-Kalashani et al., 2022b), dynamic estimation of the power system incorporating with the higher-order SMC (HOSMC) (Rinaldi et al., 2020), and islanded operation of T-type inverter by using integral-HOSMC (Barzegar-Kalashani et al., 2020).

An effective multiloop control method associated with the SRF has been developed by Monfared et al. (2014) for single-phase inverters in which the inner loop uses the PI controller as well as the capacitor current of the output filter serves as an active damper, enhancing the transient and steady-state responses, while harmonics in the load voltage are indirectly minimized by a multiresonant harmonic compensator in the capacitor current loop. Multiresonant control loops within the controller can effectively eliminate harmonics; however, the higher gain in the controller may amplify noises, resulting in the possibility of resonating of the system in the presence of unknown different harmonic orders, which will require accurate parameter regulation. Rahman et al. (2015) present a communication-oriented strategy to control the voltage and current of the parallel solar inverters assisted with PI controllers for each inverter unit, in which the designed controller is tuned by Ziegler-Nichols giving satisfactory results for steady-state performance, while transient responses show oscillations under load disturbance conditions. For the purpose of overcoming such poor transient responses and facilitating controller response that is more critical for applications such as motor drives, a derivative action can be used along with a PI controller (Shoujun and Weiguo, 2011). An alternative method is the use of PR controllers, which yield virtually no steady-state error and operate more efficiently under load disturbances (Bairagi et al., 2019). However, it should be noted that the inclusion of a derivative or resonant loop might further augment the unknown noises. Dong et al. (2011) attempt to reduce the drawbacks of the conventional PID and PR controllers by using the LCF and improve the transient response of the single-phase inverter under load variations; nevertheless, the THD in the voltage is still high. A closed-loop voltage controller, named as Posicast controller, is suggested for three-phase inverters with $LC$ filters by Li (2009). With Posicast controller, some drawbacks of conventional controllers such as resonance phenomenon are restricted by employing a resonant damping on the $LC$ filter in order to stabilize the converter. However, the effects of resistive damping from the point of view of power losses are unaffordable (Peña-Alzola et al., 2013). All in all, these conventional and optimized linear controllers which are used for single- and three-phase inverters in various applications such as motor drives, islanded microgrids, and uninterruptible power supply systems under various loading conditions are characterized by advantages such as nearly zero steady-state error and structural simplicity. However, these linear controllers are highly sensitive to frequency variations and parametric uncertainties including large external disturbances namely nonlinear loads with higher harmonic components. Other advanced linear control strategies consisting of repetitive controller with a wider feedback bandwidth (Byen et al., 2015), high performance direct deadbeat voltage control integrated with the active damping (Kim et al., 2016), and their hybrid schemes (Lidozzi et al., 2017) are used to yield proper voltage waveforms during the islanded operation of voltage source inverters. The repetitive controller implemented to the islanded single-phase inverter by Byen et al. (2015) rejects disturbances well due to its better frequency bandwidth as long as the stabilization of the controller is achieved through an error transfer function; however, this controller has multiple zeroes which cause low transient responses. Kim et al. (2016) extend the deadbeat scheme assisted by active damping to put forward the closed-loop poles into the stable plane. Nevertheless, this controller has some failures in terms of the desired operation due to existing modeling inaccuracy and parameter uncertainties. The integration of the repetitive and deadbeat controllers can improve the overall performance of the controller for which Lidozzi et al. (2017) have implemented for a three-phase four-wire inverter in standalone operation mode. Lidozzi et al. (2017) respectively use the deadbeat and repetitive controllers to faster tracking of the desired output reference and harmonic compensation.
Recently, different optimal controllers based on the LQR concept are designed for the standalone operation of inverters. For example, Batiyah et al. (2018) carry out LQR approaches for PV converters in order to achieve the fast dynamic response with the low voltage tracking error. Another research based on LQR optimized by the partial swarm optimization (PSO) algorithm is employed by Ufnalski et al. (2015) for the islanded operation of three-phase inverters under linear and nonlinear loading conditions. Hossain et al. (2015) present a linear quadratic integral (LQI) control scheme for the secondary voltage control to share the active and reactive power under unequal line inductances as long as its performance is investigated with variations in loads while its parameters are tuned by the Ziegler-Nicholas strategy. However, the LQ control schemes require an analytical model of the system, especially linearization for nonlinear models. Furthermore, to design these controllers, it is required to assume that all states are measurable which will require so many sensors in real applications.

Another solution for controlling islanded inverters, is the use of implicit and energy-function based MPCs respectively for which a less computational one is presented by Nauman and Hasan (2016) for three-phase inverters where an optimization problem is solved to reduce the computational requirements of the designed explicit MPC approach by Mariethoz and Morari (2009) and reduces these computations five times compared to traditional ones. Furthermore, Guler and Komurcugil (2022) present another MPC method for LC-filtered split source single-phase inverters where the optimum switching function is generated by using the Lyapunov function. Contrary to reduce the computational complexities, the implicit MPC developed by Nauman and Hasan (2016) and energy-function based MPC by Guler and Komurcugil (2022) use meta-heuristic algorithms for which it will be difficult to define initial design parameters and there are chances of being trapped into local minimums, specially in large disturbances (Hassan et al., 2005). In order to address the problems of meta-heuristic methods, intelligent algorithms can be integrated with existing controllers for better performance and accuracy (Dey and Seok, 2022). Nevertheless, the inclusion of these algorithms brings some complexities to the controller, which might not result in faster responses to changes. For this reason, SMC schemes are used as an alternative solution to control power converters. Abrishamifar et al. (2012) have proposed a fixed-frequency SMC to control standalone single-phase inverters using a saturation function to eliminate high frequency components in the control signal. Pichan and Rastegar (2017) have applied similar strategy for four-leg three-phase inverters (formed using three single-phase full-bridge inverters for each phase) under different loading conditions. In addition, Chaibi et al. (2019) have developed the application of fixed-switching SMC to standalone photovoltaic (PV) system for the various environmental circumstances. Though the fixed-switching SMCs are robust against disturbances in the steady-state response, the transient response is poor and the load voltage experiences significant errors during load variations. The developed controller by Gudey and Gupta (2015) for a higher-order circuit (two back-to-back LC filters) uses two control loops consisting of inner current and outer voltage loops in order to regulate the output voltage of islanded single-phase inverters and ensures the desired tracking of the output voltage following its reference value and ensures the stability. However, the use of two back-to-back LC filters for reducing the harmonics increases the order of the system which complicate the selection of the sliding surface. An adaptive SMC is used by Chen et al. (2015) to limit the problem of unknown bounds on disturbances as well as to ensure better transient responses. However, it suffers from the chattering as it does not use integral terms or higher-order sliding surfaces. Another hybrid nonlinear controller for three-phase grid-forming inverters is discussed by Li et al. (2017) which uses two loops: inner current and outer voltage loops similar to the method employed by Gudey and Gupta (2015). Li et al. (2017) employ a SMC as the inner current loop and mixed $H_2/H_{\infty}$ as the outer voltage loop in order to achieve good performance and robustness. The main issue with the suggested controller by Li et al. (2017) is that it requires the infinity or uniform norms of the system which is hard to determine. Delghavi et al. (2016) have utilized a FOSMC to control the inverter-interfaced DERs for overcoming the problem of chattering effects that inherently appear in conventional SMCs. However, it has a problem of tracking a trajectory having an integer-order of the nonlinear system (Govea-Vargas et al., 2018). Backstepping control mechanisms developed by Roy and Mahmud (2022)
may overcome some of these issues and improve the dynamic stability of the system; however, this control system may not function in the presence of matched disturbances.

Other schemes of SMCs are also designed for standalone single- and three-phase inverters. Bagheri et al. (2021) have designed a terminal SMC (TSMC) incorporating with a meta-heuristic algorithm to control frequency of islanded microgrids-interfaced inverters. The TSMCs are robust against disturbances; however, the discrepancy between control efforts in the transient and tracking errors enforce the designers to use intelligent control strategies which will make the system more complex (Wang et al., 2009). Dynamic SMC presented by Mohammadhassani and Narm (2021) incorporating with PR controller is the other control method which uses an integral action on the sliding surface for limiting chattering and harmonics. Nevertheless, this dynamic SMC approach is more complex and its operation is failed for harmonic elimination without using PR controller. Robust backstepping control technique, such as the work by Shah et al. (2021), uses a virtual control signal in each step to find the final control action; however, this controller has to employ discontinuous switching function in each step to eliminate the disturbances which increase the undesired components on final control signal.

By considering all these concepts, higher-order SMCs are taken into consideration in recent years in order to restrict chattering problems on control signals without losing the robustness of the system (Mondal et al., 2017). An observer-based HOSMC is designed by Rinaldi et al. (2020) for microgrids to investigate dynamic behaviors of power grids in terms of controlling the voltage, frequency, and phase angle (i.e., power flow) of the system. Though the robustness of the controller is studied under some disturbances, it does not consider the complete model of the system containing the filter and associated parameters. Barzegar-Kalashani et al. (2022b) have designed two robust nonlinear controllers for on- and off- grid operation modes of single-phase inverters by employing super-twisting approaches. In addition, a robust integral-HOSMC is designed for three-phase T-type inverter under different loading conditions (Barzegar-Kalashani et al., 2020). The super-twisting SMC and integral-HOSMC approaches are robust against disturbances and parametric uncertainties; however, these controllers use canonical-form of the system which derivative actions to find this form of the system cause amplification of different noises.

1.4. Specific research gaps

Based on the existing literature review, the research gaps in this area can be specified as follows:

- Linear controllers for inverters with cleaner energy systems cannot cope with their intermittencies.
- Existing nonlinear controllers are mostly sensitive to system parameters and external disturbances and thus, these cannot ensure the robustness.
- Though the utilization of SMC ensures the robustness, the design of existing SMCs requires to represent the system as special form, called canonical form.
- Existing SMCs use derivative actions that amplify the noises and hence, adversely affect the robustness.

1.5. Research contributions and novelty

In power electronic interfaces for employing green energy resources, one of the significant issues is the generation of output power with higher quality in order to minimize the utilization of additional electrical equipment for enhancing power quality. As an example, in this paper, the output voltage of the single-phase inverter is produced with lower total harmonic distortion (THD), thereby, eliminating the need to use additional harmonic compensators in the system. As a result, this solution will become more important in large-scale applications by eliminating the requirements of extra compensators that would require more space. While these power electronic interfaces are widely used today, the proper control of them poses significant challenges for robust and suitable designs,
that is why, this paper offers a robust step-by-step higher-order sliding mode controller (SbS-HOSMC) to overcome some existing challenges for islanded mode operation of single-phase inverters as summarized in Table 1. Contrary to other common approaches, the proposed controller is designed with an outer voltage control loop and an inner current control loop, where the sliding surfaces of each loop are maintained based on the full-order of the main system’s dynamic in each step. The incorporation of such sliding surfaces in full order can inherently reduce undesirable harmonic components on both the virtual and control signals, which facilitates the operation of each control loop. In addition, low harmonic components on the original control signal result in low THD and a smaller number of sub-order harmonic components. The key innovative aspects can be summarized as:

- Despite the other control strategies specially SMCs which basically use canonical-form of the state-space model of the system that cause the amplification of system’s noises due to getting couple of derivatives form the dynamic equations, the proposed controller uses block controllable-form of the system.
- The uncertainties are considered in a wide range variation to show the robustness of the proposed controller.
- The proposed controller eliminates the effects of disturbances and uncertainties in each step which will strengthen the robustness of the controller.
- In addition to eliminate the high frequencies on the control signal, apart from employing the higher order sliding surfaces, an integral action is also utilized in each step to enhance the performance of the controller.

The structure of this article is organized as follows: Section 2 describes the model of the single-phase inverter as well as the controller design based on the step-by-step full-order SMC is explained in Section 3 along with the stability issues. In order to verify the performance of the proposed scheme, Section 4 illustrates simulation studies comprising processor-in-loop (PIL) results along with comparisons. Finally, concluding remarks including the scope for future works are provided in Section 5.

### 2. System modeling

The configuration of a standalone single-phase inverter is shown in Fig. 1 which is supplied by a DC source and connected to unknown loads through an LC filter. The symbols, $R_f$, $L_f$, and $C_f$ in Fig. 1 denote the filter resistor, filter inductor, and filter capacitor, respectively. The dynamics of the voltage source inverter can be written employing the Kirchhoff’s voltage low (KVL) and
Kirchhoff’s current low (KCL) on the configuration as shown in Fig. 1 and described through the following equations:

\[
\frac{dv_L}{dt} = \frac{1}{C_f} i_f - \frac{1}{C_f} i_L \tag{1}
\]

\[
\frac{di_f}{dt} = \frac{1}{L_f} v_t - \frac{1}{L_f} L - \frac{R_f}{L_f} i_f \tag{2}
\]

where \(v_L\), \(i_f\), \(i_L\), and \(v_t\) are the load voltage, filter current, load current, and inverter voltage, respectively. In this study, the main purpose is the load voltage control, i.e., \(v_L\) by producing a proper control signal (\(v_t\)) through robust controller. The load characteristics are usually unknown and it is extremely difficult to maintain the desired load voltage during the islanded operation of the single-phase inverter due to the absence of the grid. Therefore, the load current, i.e., \(i_L\) in equation (1) is considered as an external disturbance.

A robust SbS-HOSMC is designed to ensure the desired load voltage during the islanded operation of a single-phase inverter with unknown loads and the procedure of the proposed scheme is discussed in the following section.

3. Proposed step-by-step higher-order sliding mode controller design

In order to design the controller, some assumptions are taken into considerations by associating the physical model of the system. These assumptions are defined based on the physical system characteristics such as its limitations from the perspectives of rating values as well as bounded disturbances based on these rated values. All these can be defined as follows:

**Assumption 1.** The generator-side converter of the DER is responsible for regulating the input voltage of the inverter. It is therefore assumed that the dc-link voltage is constant.

**Assumption 2.** The behavior of existing loads in microgrids is uncertain, which can be considered an external disturbance. As each electrical component has a certain power rating, the upper-bound of this disturbance is limited with regard to the physical system’s perspectives.

**Assumption 3.** The parametric uncertainties are considered as internal disturbances that are bounded. In light of the physical model of the DER, this assumption is valid since the variations in the parameters, such as the filter capacitor and the filter resistor, are relatively bounded.

**Assumption 4.** Assumptions 2 and 3 also entails bounded algebraic expressions.
As mentioned earlier, the control objective for a standalone single-phase inverter as shown in Fig. 1 is the load voltage control. Since the characteristics of loads (i.e., the load current and load power) are unknown, the load current \( i_L \) is considered as an external disturbance satisfying Assumptions 2 – 4. By considering all these explanations, the outer- and inner loops of the proposed controller are designed as follows:

### 3.1. Step 1: Outer voltage control loop

As illustrated in Fig. 1, the outer control loop represents the voltage control loop which is the first step of the proposed controller. The outer control loop compares the actual load voltage with the reference voltage, and the resultant signal is then used as a virtual input reference for the second step. The following subsections provide details of the different parts of the outer loop:

#### 3.1.1. Second-order voltage sliding surface

If the load reference voltage is defined as \( v_{ref}^L \), the error signal corresponding to the load voltage can be defined through the following equation:

\[
e_v = v_L - v_{ref}^L \tag{3}\]

In this step, the first sliding surface, which has a second-order dynamic of the load voltage error signal, can be defined as follows (Mondal et al., 2017):

\[
S_1 = \dot{e}_v + k_1 |e_v|^\alpha_1 \text{sign}(e_v) + k_2 |\dot{e}_v|^\alpha_2 \text{sign}(\dot{e}_v) \tag{4}\]

where \( k_1 \) and \( k_2 \) are coefficients for the first sliding surface whose values are always positive and must satisfy the Hurwitz condition (Bhat and Bernstein, 2005). Furthermore, \( 0 < \alpha_1, \alpha_2 < 1 \) are another two coefficients for the proposed controller.

#### 3.1.2. Control signal of the outer loop

The proposed controller is based on a step-by-step mechanism that allows the inner- and outer loops to be designed in different order through sliding surfaces. For the studied system, since the output control signal of the outer loop serves as a reference signal for the inner loop, the filter current is selected as a virtual control signal, i.e., \( \phi_1 = i_f \). By considering the term \( -\frac{1}{C_f} i_L \) as external disturbance, the virtual control signal can be written as follows:

\[
\phi_1 = C_f (u_{eq1} + u_1) \tag{5}\]

where \( u_{eq1} \) and \( u_1 \) are the equivalent virtual control signal and virtual switching control signal, respectively which are determined by the following equations:

\[
u_{eq1} = -k_1 |e_v|^\alpha_1 \text{sign}(e_v) - k_2 |\dot{e}_v|^\alpha_2 \text{sign}(\dot{e}_v) + v_{ref}^L \tag{6}\]

\[
\dot{u}_1 + \tau_1 u_1 = -k_{uv} \text{sign}(S_1) \tag{7}\]

where \( \tau_1 \) is the inverse time constant of the switching signal. The equivalent and switching virtual signals in \( \phi_1 \) can be considered as continuous signals which facilitate the operation of the inner loop in the next step an the detailed evidence is provided as the following remark:

**Remark 1.** The equivalent control signal consists of the terms in the form of \( k_x \cdot |^{\alpha_x} \text{sign}(\cdot) \) within subscript \( x \) as 1 and 2, which is a continuous function (Bhat and Bernstein, 2005), resulting in less chattering on the virtual control input for the inner loop.
Remark 2. From equation (7), it can be seen that the virtual switching control signal appears as a derivative and the solution of equation (7) will require an integral action. In this regard, by taking a Laplace transform from equation (7), we have (Barzegar-Kalashani et al., 2022a):

$$U_1(s) = \frac{F_1(s)}{s + \tau_1} \quad (8)$$

where $U_1(s)$ is the Laplace transform of $u_1$ and $F_1(s) = \int_0^\infty -k_{q1} \text{sign}(S_1) dt$. From equation (8), it is evident that a portion of the existing high frequency component in the switching signum function is filtered by the term $\frac{1}{s + \tau_1}$ (Mondal et al., 2017), which leads to smooth virtual switching control signal as the input for the inner loop.

Remark 3. Remarks 1 and 2 approve smooth functionality of the $u_{eq1}$ and $u_1$ ensuing smooth virtual control signal (i.e. $\phi_1$) for the inner loop in the next step design of the controller.

3.1.3. Outer loop stability proof

To investigate the stability and convergence of the virtual control signal, the Lyapunov function can be defined in terms of the first sliding surface and written as follows:

$$W_1(t) = \frac{1}{2} S_1^2 \quad (9)$$

It is required that the derivative of $W_1(t)$, i.e., $\dot{W}_1(t)$ be negative semi-definite for the first sliding surface which contains the error signal of the load voltage and the negative semi-definiteness of $\dot{W}_1(t)$ defines the convergence of this error to zero. Therefore, it can be written as follows:

$$\dot{W}_1(t) = S_1 \dot{S}_1 \leq 0 \quad (10)$$

In order to prove (10), the sliding surface in equation (4) can be rewritten using equations (1) and (3) which can be expressed as follows:

$$S_1 = \frac{1}{C_f} i_f - \frac{1}{C_f} i_L - v_{ref}^L + k_1 |e_1|^{\alpha_1} \text{sign}(e_1) + k_2 |\dot{e}_1|^{\alpha_2} \text{sign}(\dot{e}_1) \quad (11)$$

By introducing the control signals from equation (5) into equation (11), it can be written as follows:

$$S_1 = u_{eq1} + u_1 - \frac{1}{C_f} i_L - v_{ref}^L + k_1 |e_1|^{\alpha_1} \text{sign}(e_1) + k_2 |\dot{e}_1|^{\alpha_2} \text{sign}(\dot{e}_1) \quad (12)$$

By substituting equation (6) into equation (12), it can be simplified as:

$$S_1 = u_1 - \frac{1}{C_f} i_L \quad (13)$$

Now, the derivative of $S_1$ can be obtained as:

$$\dot{S}_1 = \dot{u}_1 - \frac{1}{C_f} i_L \quad (14)$$

and by using equation (7), equation (14) can be written as follows:

$$\dot{S}_1 = -\tau_1 u_1 - k_{q1} \text{sign}(S_1) - \frac{1}{C_f} i_L \quad (15)$$
Using equation (15), equation (10) can be written as:

\[
\dot{W}_1(t) = S_1(-\tau_1 u_1 - k_1 \text{sign}(S_1) - \frac{1}{C_f} \dot{i}_L) = \\
- (\tau_1 u_1 S_1 + k_1 \text{sign}(S_1) S_1 + \frac{1}{C_f} \dot{i}_L S_1) \leq \\
- (\tau_1 |u_1| S_1 + k_1 |S_1| + \frac{1}{C_f} |\dot{i}_L| |S_1|) \leq \\
- (\tau_1 u_1 + k_1 + \frac{1}{C_f} |\dot{i}_L|) |S_1| \leq \\
- k_{u_1} |S_1| \leq -k_{u_1} \sqrt{2\dot{W}_1(t)} \leq 0
\]  

(16)

where the upper bound of the perturbation must satisfy the following inequality until equation (16) be negative semi-definite:

\[\left| \tau_1 u_1 + \frac{1}{C_f} \dot{i}_L \right| \leq k_{u_1} \]

(17)

Taking into account Assumptions 1 to 4, the perturbation on the left-hand side of inequality (17) consists of the derivative of current flow and capacitor capacitance $C_f$, which are bounded due to limited ratings.

**Remark 4.** The designed second-order sliding surface in the first step removes the disturbance effects under the practical assumptions and inequality (17), allowing the voltage control loop’s output to be mitigated by these disturbances and facilitating the operation of the inner current loop in the following step.

3.2. Step 2: Inner current control loop

In this last step, the virtual control signal generated by the outer voltage control loop is compared with the actual filter current and the resulting error signal is used through the inner loop to produce the control signal that will be utilized in the pulse width modulation (PWM) loop.

3.2.1. First-order current sliding surface

Since $\phi_1 = i_f$ and $\dot{\phi}_1$ is also a function of virtual equivalent and switching control signals as represented by equation (5), the error signal for the filter current can be written as follows:

\[e_i = i_f - \phi_1\]

(18)

The second sliding surface associated with this error signal can be defined as follows:

\[S_2 = \dot{e}_i + k_2 |e_i|^\alpha \text{sign}(e_i)\]

(19)

3.2.2. Control signal of the inner loop

The original control signal can be expressed as follows:

\[v_i = L_f (u_{eq2} + u_2)\]

(20)

where $u_{eq2}$ and $u_2$ are equivalent and switching terms of the control signal which can be written as:

\[u_{eq2} = \frac{1}{L_f} v_e + \frac{R_f}{L_f} i_f - k_2 |e_i|^\alpha \text{sign}(e_i) + \dot{\phi}_1\]

(21)

\[\dot{u}_2 + \tau_2 u_2 = -k_{u_2} \text{sign}(S_2)\]

(22)
**Remark 5.** As in Remark 1, the equivalent control signal $u_{eq2}$ is also based on a continuous function $k_2|\cdot|^{\alpha_1}\text{sign}(\cdot)$ that reduces certain high frequency components.

**Remark 6.** The discontinuous signum function in equation (22) will be under the integral term during the determination of $u_2$, limiting the high frequency components in the final control signal. Equation (22) is transformed into the Laplace form as follows (Barzegar-Kalashani et al., 2022a):

$$U_2(s) = \frac{F_2(s)}{s + \tau_2}$$  \hspace{1cm} (23)

With $U_2(s)$ as the Laplace transform of $u_2$ and $F_2(s) = \int_0^\infty -k_2\text{sign}(S_2)dt$, the term $\frac{1}{s + \tau_2}$ in equation (23) filters a certain amount of higher components of $F_2(s)$ (Mondal et al., 2017), yielding softer reference control signal for the PWM loop.

3.2.3. Inner loop stability proof

The stability can be analyzed by formulating the following the second Lyapunov function:

$$W_2(t) = \frac{1}{2}S_1^2 + \frac{1}{2}S_2^2$$  \hspace{1cm} (24)

For the stabilization of the controller against the changes, it is essential to make the derivative of second Lyapunov function as negative semi-definite for which the derivative of equation (24) can be written as:

$$\dot{W}_2(t) = \dot{W}_1(t) + S_2 \dot{u}_2 \leq 0$$  \hspace{1cm} (25)

Referring to equation (16), $\dot{W}_1(t)$ is semi-negative. So, the second term of (25) i.e. $S_2 \dot{u}_2$ is considered through stability issue. By using equations (2), (18), and (19), the second sliding surface is written as follows:

$$S_2 = \frac{1}{L_f}v_t - \frac{1}{L_f}v_L - \frac{R_f}{L_f}i_f - \dot{\phi}_1 + k_2|e_i|^{\alpha_1}\text{sign}(e_i)$$  \hspace{1cm} (26)

By substituting $v_t$ from equation (20) into equation (26), it can be written as:

$$S_2 = u_{eq2} + u_2 - \frac{1}{L_f}v_L - \frac{R_f}{L_f}i_f - \dot{\phi}_1 + k_2|e_i|^{\alpha_1}\text{sign}(e_i)$$  \hspace{1cm} (27)

Equation (27) can be more simplified by using equation (21) which can be written as:

$$S_2 = u_2$$  \hspace{1cm} (28)

By using the value of $S_2$ from equation (28), equation (25) can be written as follows:

$$\dot{W}_2(t) = \dot{W}_1(t) + S_2 \dot{u}_2$$  \hspace{1cm} (29)

As proved in equation (16), the derivative of $W_1(t)$ is equal to or less than $k_{u1}\sqrt{2W_1(t)}$, therefore, equation (29) can be written as:

$$\dot{W}_2(t) \leq -k_{u1}\sqrt{2W_1(t)} - \tau_2u_2S_2 - k_2\text{sign}(S_2)S_2 \leq -k_{u1}\sqrt{2W_1(t)} - \tau_2u_2|S_2| - k_2|S_2| \leq -k_{u1}\sqrt{2W_1(t)} - (\tau_2u_2 + k_2)|S_2| \leq -k_{u1}\sqrt{2W_1(t)} - k_2|S_2| \leq 0$$  \hspace{1cm} (30)
Equation (30) will be negative semi-definite under the following inequality:

$$|\tau_2 u_2| \leq k \nu_2$$  \hspace{1cm} (31)

Based on this design procedure, the closed-loop diagram for the voltage control of a standalone single-phase inverter using the proposed full-order SMC is illustrated in Fig. 1. The performance of the designed controller is studied in the next section through some simulation results along with real-time implementation and comparisons with other existing controllers.

4. Simulation and real-time implementation results

In this section, the simulation studies are investigated on a single-phase inverter as shown in Fig. 1 in which the power circuit is simulated in MATLAB 2020b/Simulink while the proposed controller is implemented through both Simulink blocks via MATLAB and PIL via DSP. Note that throughout all simulations and implementations, a discrete time simulation type is employed with a sampling time equal to 1 µs, utilizing variable-step for simulations as well as fixed-step for PIL using the ode45 (Dormand-Prince) solver. In this regard, Fig. 2 indicates the photography of the PIL implementation on PC (8 GB RAM, Intel(R) Core(TM) i5-7300U CPU @ 2.6 GHz) and F28335 controlCARD R2.2 DSP microprocessor. All simulations and PIL implementations have been run for 0.15 s in order to examine the performance of the designed controller. Within this time period and using the hardware configuration mentioned above, the simulation is completed within approximately 5 s due to the variable-step method used in the simulation. Nevertheless, this system in PIL simulations requires a longer runtime of about 10 minutes, due to the use of a fixed-step solving method and the need to transfer data from the PC to the DSP and receive the results from the DSP. In practice, transferring a large amount of data from the PC to the DSP via USB cable results in the system running for an extended period of time. The parameters of the system and controller are listed in Table 2 while keeping Assumptions 1 and 2. The desired load voltage is selected as $v_{ref}^L = 220 \sqrt{2} \sin(2\pi f_o t)$ in volts, and the controller must determine how to maintain the peak value of the load voltage at $220 \sqrt{2} \text{ V}$, while meeting grid code requirements for THD (IEEE Std 519, 2022). The robustness of the proposed controller needs to be examined under various external and internal disturbances, for which three different case studies are considered. From the standpoint of external disturbances, the inverter should be able to serve a variety of linear and nonlinear loads while maintaining the load voltage at its reference level. In light of Assumption 2, the rating values of these loads are determined by referring to the rating power of the DER. As should be noted, the load characteristics play a crucial role in the islanded operation of the single phase inverter, which is why the most severe case studies are taken into consideration by selecting the resistive and nonlinear loads for transient and steady-state analyses. Additionally, the effects of internal disturbances that arise from parametric uncertainties are also critical to the design of a more practical controller, and as a third case study, it is analysed how well the controller performs under large parametric uncertainties. There are three case studies presented below which provide more detailed analyses along with virtual and real-time simulations as well as appropriate comparisons of the results.

4.1. Case study 1: Performance analysis of the proposed ShS-HOSMC scheme with linear loads

A robustness assessment of the proposed controller is carried out in this case study in terms of both steady-state and transient responses from where Fig. 2 demonstrates an example of the implementation of the proposed ShS-HOSMC controller through Simulink and PIL. In Fig. 3, it can be seen that the virtual and real-time simulations are essentially identical with a little bit difference resulting from the delaying of the signal through PIL on the DSP. In this case study, a resistive full load with $R = 22 \Omega$ is considered from $t = 0$ s to $t = 0.05$ s for which the voltage and current will be in phase and can be clearly seen from Figs. 3(a) and (e). At $t = 0.05$ s, this load is disconnected from the inverter and for this reason, the current flowing through the load will be zero until $t = 0.07$ s
Table 2: The studied system and controller parameters

<table>
<thead>
<tr>
<th>Items</th>
<th>Coefficient</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>System</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$V_{dc}$</td>
<td></td>
<td>550 V (k1, k2)</td>
</tr>
<tr>
<td>$R_f$</td>
<td></td>
<td>0.2 $\Omega$ (\alpha_1, \alpha_2)</td>
</tr>
<tr>
<td>$L_f$</td>
<td></td>
<td>1.5 mH (k_{\upsilon 1}, k_{\upsilon 2})</td>
</tr>
<tr>
<td>$C_f$</td>
<td></td>
<td>80 $\mu$F (\tau_1, \tau_2)</td>
</tr>
<tr>
<td>Grid frequency ($f_o$)</td>
<td></td>
<td>50 Hz</td>
</tr>
<tr>
<td>Switching frequency ($f_{sw}$)</td>
<td></td>
<td>14 kHz</td>
</tr>
<tr>
<td>Sampling time ($T_s$)</td>
<td></td>
<td>1 $\mu$s</td>
</tr>
<tr>
<td>Controller</td>
<td></td>
<td></td>
</tr>
<tr>
<td>$k_1$, $k_2$</td>
<td></td>
<td>($3 \times 10^4$, 2)</td>
</tr>
<tr>
<td>$\alpha_1$, $\alpha_2$</td>
<td></td>
<td>(95, 8)$\times 10^2$</td>
</tr>
<tr>
<td>$k_{\upsilon 1}$, $k_{\upsilon 2}$</td>
<td></td>
<td>(9000, 1)$\times 10^4$</td>
</tr>
<tr>
<td>$\tau_1$, $\tau_2$</td>
<td></td>
<td>(1, 5)$\times 10^{-3}$ (1/s)</td>
</tr>
</tbody>
</table>

Figure 2: The photography of PIL implementation through PC and DSP
Figure 3: Load voltage, voltage error, THDs, and load current for Simulink and PIL implementations under linear loading conditions: (a) Load voltage with its snapshots under full load and no load conditions, (b) Instantaneous voltage error, (c) THD under the full load condition, (d) THD under the no load condition, and (e) Load current under full load and no load conditions.
though the voltage will remain the same, which can also be found from Fig. 3(a). Upon reconnection of the resistive full load to the system, the resistive load will again flow a current similar to the initial value while the voltage across the load will remain constant. This can be seen in Figs. 3(a) and (e). At $t = 0.07$ s, the resistive full load is reconnected to the system and in this circumstance, the current flowing through the load will be similar that of its initial value while the load voltage remains constant which can be found from Figs. 3(a) and (e). Fig. 3(a) with its close up images clearly show that the load voltage remains constant, approximately despite the loading conditions, i.e., whether the system is on no load or full load. Furthermore, according to the Fig. 3(e), the load current experiences transients at the instant of the disconnection and reconnection to the system. However, the designed controller manages these transients in a very good way. Fig. 3(b) indicates the instantaneous voltage errors for both Simulink and PIL implementations with a bit distinction between them where the percentage of the voltage error is calculated using the following equation (Pichan and Rastegar, 2017):

$$e_{v,p} = \frac{V_L - V_{Lref}}{V_L} \times 100 \quad (32)$$

where $V_L$ and $V_{Lref}$ are the root mean square (rms) values of the actual and reference output voltages, respectively. Considering equation (32), voltage tracking errors for full load and no load conditions are obtained as 0.017% and 0.047%, respectively. The THDs in the load voltage for full load and no load conditions are illustrated in Figs. 3(c) and (d), respectively. From these figures, it can be found that the THD for the full load is 0.31% while the peak value of the load voltage is 311.1 V. Similarly, the THD in the load voltage for the no load condition is 0.15% for which the peak value of this voltage is 311.2 V. Furthermore, the THD analyses from Figs. 3(c) and (d) indicate that the harmonic components in each order have been greatly improved. This is because the values for each order are very low, complying with the grid code standards (IEEE Std 519, 2022), which is crucial for sensitive loads.

4.2. Case study 2: Performance analysis of the proposed SbS-HOSMC scheme with nonlinear loads

In this case study, the robustness of the controller is investigated for large disturbances appearing from nonlinear loads. For this purpose, two standard nonlinear loads with different crest factors are connected at the output of the inverter and the schematic diagram of these loads are illustrated in Fig. 4 which include a parallel $RC$ load with a series resistor and a series $RL$ load as recommended by IEC 62040-3 (2021). According to the standard, a full-bridge rectifier is used to simulate highly harmonic loads in order to evaluate the performance of the controller, as shown in Fig. 4.

To explore the effectiveness of the proposed SbS-HOSMC approach, different loading scenarios are implemented at the output of the inverter by the use of a full-bridge inverter with $RL$ and $RC$ loads. At first, the $RC$ load is considered with a crest factor of 3.57 and this value is calculated using the approach as presented in (IEC 62040-3, 2021) by considering $R_s = 0.88 \ \Omega$, $R = 49.61 \ \Omega$, $C = 3000 \ \mu F$. The simulation results through Simulink and PIL implementations are obtained in terms of the load voltage and current responses along with the THDs corresponding to these responses. The designed full-order sliding mode controller provides robustness against load disturbances and the load voltage will remain the same which can be seen from Fig. 5(a) with acknowledging a bit discrepancy between Simulink and PIL consequences as can be seen in zoomed plots. At this instant, the current response is also depicted in Fig. 5(c) which shows that the proposed controller converges the current waveform to its steady-state value. Furthermore, the instantaneous load voltage error is shown in Fig. 5(b) where the percentage load voltage error is 0.98%. The THDs in both load voltage and current are shown in Fig. 5(d) and Fig. 5(e), respectively. Regarding this, it can be seen from Fig. 5(d) that the THD for the load voltage is 1.57% which is higher than that of the linear load and the main reason behind this is the connection of a highly nonlinear load; nevertheless, this value still meets the standard requirements. On the other hand, the THD for the load current with the $RC$ load is 101.96% which can be seen from Fig. 5(e). It is worth noting that the THD for the load current is too high and the reduction of the THD for the load current is not the main focus of this paper.
Similarly, the performance of the designed controller is evaluated with the nonlinear RL load by considering $R = 49.61$ Ω and $L = 180$ mH which correspond the crest factor as 1.32. Under this nonlinear loading condition, the load voltage and current responses are shown in Fig. 6(a) and (c), respectively. Furthermore, the percentage load voltage error and corresponding THDs for the load voltage and current are displayed in Fig. 6(b), Fig. 6(d), and Fig. 6(e), respectively. The percentage load voltage error is 0.017% while the values of THDs for the load voltage and current are 0.29% and 39.67%, respectively which can also be seen Fig. 6(d) and Fig. 6(e), respectively. The comparison among the results for the nonlinear load with RC and RL loads indicates that the values of THDs for the load voltage and current with the RL load are lower than that of the RC load.

4.3. Case study 3: Performance analysis of the proposed SbS-HOSMC scheme under parametric uncertainties

In this case study, the robustness of the proposed SbS-HOSMC strategy is investigated against internal disturbances. For this purpose, the values of the filter parameters, i.e., $C_f$ and $L_f$ are changed and the performance of the system is studied in terms of harmonic components and voltage tracking error from its reference value. The parameters of the system are similar to that as in Table 2 where the values of $L_f$ and $C_f$ are changed from -75% to +75% of from their nominal values with the step of 25%. It is also important to note that the worst load condition is considered for this case study (i.e., the nonlinear load in Fig. 4(a) as studied in the second case study). Fig. 7 indicates the percentage of relative voltage corresponding to each odd harmonic for different variations in the filter parameter. Fig. 7(a) illustrates the relative voltages as the percentage for each odd harmonic component compared with the EN 50160 standard (EN 50160, 2010) in the presence of variations in $C_f$ from its nominal value ($C_f = 80$ µF) while keeping constant $L_f$, i.e., $L_f = 1.5$ mH. As can be seen from the Fig. 7(a), with increasing $C_f$, the relative voltage for each harmonic component decreases. Furthermore, the percentages of this relative voltage corresponding to each odd harmonic component meet the standard in the minimum value of $C_f$ which is due to the robustness feature of the proposed controller as it efficiently eliminates these uncertainties. To investigate the variations in $L_f$, another study is performed by keeping $C_f$ as constant, i.e., $C_f = 80$ µF while varying $L_f$ from its nominal value (i.e., $L_f = 1.5$ mH). Fig. 7(b) depicts simulation results of relative voltages as the percentage comparing with the EN 50160 standard (EN 50160, 2010) for different harmonic components in the presence of uncertainties in $L_f$. As this figure shows, variations in $L_f$ do not affect the performance of the controller which also illustrates the robustness of the controller.

By comparing Figs. 7(a) and (b), it can be said that the controller’s performance is less sensitive to variations in $L_f$ than variations in $C_f$, since the capacitor acts as a low pass filter for the voltage. However, large variations in filter parameters cannot move the system towards the unstable region.

In order to further investigate the performance of the proposed controller in the presence of parametric uncertainties, Figs. 8(a) and (b) indicate the percentages of THDs and tracking errors in the load voltage, respectively. Fig. 8(a) shows that the THDs for different values of $C_f$ and $L_f$ meet the standard except for the THD corresponding to -75% variation in $C_f$ as it is a little higher than the standard limit (i.e., 5%) which is because of imposing the worst condition for the system when the system experiences the minimum value of $C_f$ in the presence of highly nonlinear harmonic loads. Fig. 8(b) shows the percentage of the load voltage tracking
Figure 5: Load voltage, voltage error, load current, and THDs for Simulink and PIL implementations under nonlinear RC loading conditions: (a) Load voltage, (b) Instantaneous voltage error, (c) Load current, (d) THD for the load voltage, and (e) THD for the load current.
Figure 6: Load voltage, voltage error, load current, and THDs for Simulink and PIL implementations under nonlinear RL loading conditions: (a) Load voltage, (b) Instantaneous voltage error, (c) Load current, (d) THD for the load voltage, and (e) THD for the load current.
Figure 7: The percentage of load voltage for different odd harmonic components in the presence of variations in filter parameters: (a) Constant $L_f = 1.5 \, \text{mH}$ and variable $C_f$, (b) Constant $C_f = 80 \, \mu\text{H}$ and variable $L_f$. 
error in the presence of variations in $C_f$ and $L_f$ variations in which the proposed controller has less tracking errors under most of these variations except for the case as indicated earlier on. All these can be compared with other controllers as indicated in Table 3.

The comparative studies between the designed controller and other existing methods are illustrated in Table 3. From this table, it can be seen that the designed controller performs better than other existing controllers in terms of maintaining reduced values of the THD and tracking error for the load voltage under both linear and nonlinear loading conditions. A distinct aspect of the proposed SbS-HOSMC over other controllers is its use of first-order dynamical models for each state variable in each loop, as opposed to other controllers which use the canonical-form of the state-space model. In this context, it may be advantageous to use a state-space model of the system in canonical form to obtain low measurements; nevertheless, derivative actions on variables can increase noise effects, which may not be appropriate for some systems, particularly those with sensitive loads.

5. Conclusions

A robust full-order SMC is designed for islanded inverter applications in order to ensure the voltage control. The performance of the designed controller is rigorously analyzed by considering different scenarios, mainly different loading conditions and variations in parameters. The designed controller ensures better performance comparing with all other methods which is clearly evidenced from its fast transient response for changing load conditions from full-load to no-load and vice-versa. In the worst case study for highly nonlinear loads, the load voltage at the inverter’s output tracks its reference with a minimum voltage tracking error of 0.45% and a lower voltage THD of 1.57%, which are somewhat less than the voltage errors with approaches like synchronous reference frame voltage control (Monfared et al., 2014) and optimized LQR (Dong et al., 2011) that are around 1.68%. However, the voltage

![Graph](image-url)
Table 3: Comparison of the designed controller with existing methods

<table>
<thead>
<tr>
<th>Category</th>
<th>(Monfared et al., 2014)</th>
<th>(Dou et al., 2015)</th>
<th>(Byen et al., 2015)</th>
<th>(Kim et al., 2016)</th>
<th>(Ufnalski et al., 2015)</th>
<th>Designed Controller</th>
</tr>
</thead>
<tbody>
<tr>
<td>THD, (Resistive load) (%)</td>
<td>0.20</td>
<td>NA</td>
<td>NA</td>
<td>NA</td>
<td>NA</td>
<td>0.31</td>
</tr>
<tr>
<td>Voltage Tracking Error, (Resistive load) (%)</td>
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<td>NA</td>
<td>NA</td>
<td>NA</td>
<td>0.01</td>
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<tr>
<td>THD, (Nonlinear load with RC load) (%)</td>
<td>1.68</td>
<td>2.95</td>
<td>2.00</td>
<td>10.06</td>
<td>1.67</td>
<td>1.57</td>
</tr>
<tr>
<td>Voltage Tracking Error, (Nonlinear load with RC load) (%)</td>
<td>3.00</td>
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<td>0.17</td>
<td>NA</td>
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<td>0.45</td>
</tr>
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<td>State-space model</td>
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<td>Canonical</td>
<td>Canonical</td>
<td>Canonical</td>
<td>Block-controllable</td>
</tr>
</tbody>
</table>

tracking error for the deadbeat control scheme presented by Kim et al. (2016) is around 10% which is much more inefficient than the proposed controller. Furthermore, the controller performance under large variations in parameters clearly demonstrate the robustness against parametric uncertainties, with variations in the filter parameters from -75% to +75% exhibiting approximately 3% error in voltage tracking and less than 5% THD in voltage, which demonstrate that the proposed SbS-HOMC approach still meets grid requirements. Hence, the designed controller ensures the desired voltage tracking performance under any loading conditions while varying parameters. Furthermore, the desired output voltage has low THD which is an important factor especially for sensitive loads such as computers’ consumers. In addition to the low value of the THD, the designed controller also has good harmonic components spectrum which meet the standard limitations. The proposed controller is designed based on the assumption that the input energy source is a dc-voltage source. The inclusion of renewable energy resources can add some additional dynamics which require redesigning of the existing controller by considering these nonlinear dynamics. As a future work, the authors intend to develop the proposed controller for the application of photovoltaic and wind energy integration into the system. Another future work will consider the design and implementation of the proposed scheme with multiple distributed energy resources connected through several parallel inverters with strong dynamic interactions.

References


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URL https://webstore.iec.ch/publication/60140


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Declaration of interests

☒ The authors declare that they have no known competing financial interests or personal relationships that could have appeared to influence the work reported in this paper.

☐ The authors declare the following financial interests/personal relationships which may be considered as potential competing interests: