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Optimized Sliding Mode Control to Maximize Existence Region for Single-Phase Dynamic Voltage Restorers

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Abstract—This paper presents an optimized sliding mode control (SMC) strategy to maximize existence region for single-phase dynamic voltage restorers. It is shown analytically that there exists an optimum sliding coefficient which enlarges the existence region of the sliding mode to its maximum. Also, it is pointed out that the optimum sliding coefficient improves the dynamic response. In addition, a double-band hysteresis control which ensures the switching of a transistor in the voltage source inverter during a half-cycle while it remains either on or off in the other half cycle is used to mitigate the switching frequency. The theoretical considerations and analytical results are verified through computer simulations and experimental results. Simulation and experimental results show that the proposed SMC strategy not only compensates the undesired voltage disturbances and maintains the load voltage at desired level with low total harmonic distortion, but also exhibits fast dynamic response and operates at reasonably low switching frequency.

Index Terms—Dynamic voltage restorer, existence region, sliding mode control, voltage harmonic, voltage sag, voltage swell.

I. INTRODUCTION

VOLTAGE disturbances such as voltage sags, swells, and harmonics existing in the electrical grid have undesired effects on the sensitive loads such as computing devices, communication network, manufacturing process, and adjustable speed drives [1]. Specially, the voltage sags lead to interruption of the sensitive loads employed in the industry which cause significant costs due to loss of production. Generally, power electronics converters such as shunt active power filters (SAPF) and dynamic voltage restorers (DVR) are used to improve the quality of power in such sensitive loads and reduce the total harmonic distortion (THD) of the current and voltage [2], [3]. A DVR which is a series custom power device offers a prominent solution for protecting the sensitive load from these disturbances.

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The objective of a DVR is to inject a voltage (known as the compensation voltage) in series with the grid voltage to maintain the load voltage at desired level for all times.

This objective can be achieved if the DVR is controlled by an appropriate control strategy. Generally, the control strategy devised for a DVR should be able to offer a fast dynamic response, high robustness to parameter variations, sinusoidal load voltage with low THD, and small steady-state errors. In addition, it should be able to generate sinusoidal reference for the compensation voltage under distorted grid voltage conditions.

Many control strategies for single- and three-phase DVRs have been devised in the literature for achieving the aforementioned performance requirements [3]–[20]. The feedback control introduced in [3] and [4] involves intensive computation to generate the compensation voltage reference. The H-infinity control strategy proposed in [5] offers robustness in controlling the compensation voltage at the expense of complexity in the practical implementation. The control method presented in [6] is based on fuzzy logic which models qualitative aspects of human knowledge through linguistic IF-THEN rules. However, the number of IF-THEN rules, shape of membership functions, and input-output scaling gains are determined by trial-and-error method. The repetitive control strategy copes with the periodic disturbances successfully and offers a good steady-state performance, but it suffers from slow dynamics and poor performance to nonperiodic disturbances [7]. As a remedy to the variable switching frequency problem existing in the conventional hysteresis control, an alternative hysteresis control operating with constant switching frequency has been proposed in [8]. However, the high-switching frequency still exists. In [9], different voltage injection methods are proposed with emphasis on the reduced rating of the voltage source inverter (VSI). The method proposed in [10] prevents the saturation of series transformer. On the other hand, the voltage sag detection methods are also proposed in [11] and [12].

Recently, new DVR topologies are proposed to achieve the aforementioned objectives. The interline DVR consists of several DVRs and shares a common dc link connecting independent feeders [13]. In [14] and [15], the DVR is realized by employing a multilevel inverter. The control strategies for the transformerless DVR [16], [17] and the storageless DVR [18]–[20] are also studied.

The abovementioned control strategies yield various advantages and disadvantages related to dynamic response, steady-state error in the load voltage, control circuit complexity, robustness, and switching frequency. The sliding mode control

(SMC) strategy is a good candidate for offering important advantages like fast dynamic response, strong robustness to parameter variations, and simplicity in practical implementation [21]–[24]. The SMC involving higher order sliding-modes is also proposed by Levant [25]. Recently, the adaptive version of the higher order SMC is proposed for fault reconstruction for a class of nonlinear uncertain systems where the uncertain parameters are estimated [26]. The adaptive law was then utilized to reconstruct the existing faults in the system. The SMC strategy with the aforementioned advantages is applied to the control of three-phase ac/dc converter [27], dc–dc converters [28], interleaved converters [29], and single- and three-phase DVRs [30]–[32]. In [30] and [31], the sliding surface function is formed by using the inductor current and capacitor voltage errors where the measurements of inductor current and capacitor voltage are essential. Also, the generation of inductor current and capacitor voltage references are required in this approach. Furthermore, the necessity of having two sliding coefficients in the sliding surface function, not only increases the complexity of this approach, but also makes the determination of existence conditions difficult. Although the SMC approach proposed in [32] eliminates the need of using inductor current error in the sliding surface function, it still requires two sliding coefficients, which lead to a difficulty in determining the existence condition of the sliding mode.

The SMC approach having only the capacitor voltage error with one sliding coefficient is successfully applied to the control of single-phase UPS inverters [33], [34]. In [33], it is pointed out that the sliding coefficient plays an important role in determining the dynamic response and existence region of the sliding mode. While small-valued sliding coefficient leads to slow dynamic response, large valued sliding coefficient may cause undesirable overshoots in the inverter's output voltage. The rotating sliding-line-based SMC, which employs a time-varying sliding coefficient, offers an improvement in the dynamic response without paying attention to the existence region of the sliding mode [34]. The value of sliding coefficient is increased or decreased according to the values of state variables. However, the relationship between the dynamic response and existence region of the sliding mode is not investigated.

In this paper, an optimized SMC strategy is proposed for single-phase DVRs which guarantees a maximum existence region for the sliding mode and achieves fast dynamic response without having a risk of overshoots in the compensation voltage. The optimum value of the sliding coefficient that enlarges the existence region to its maximum is derived analytically. In addition to the prominent advantages of SMC, a double-band hysteresis control (DBHC) is used to mitigate the switching frequency with the result that a switching device is only switched during a half-cycle, while it remains either on or off in the other half cycle. The theoretical considerations and analytical results are verified through computer simulations and experimental results.

II. MODELING AND OPERATION PRINCIPLE

Fig. 1 shows the basic circuit diagram of a single-phase DVR. The DVR consists of an H-bridge VSI with four switching

Fig. 1. Basic circuit diagram of a single-phase DVR.

devices which are operated in such a way that the DVR injects the required compensation voltage in series with the grid voltage ($v_g = V_g \sin(\omega t)$) through a series transformer. The transformer also provides an electrical isolation between the VSI and grid. The grid impedance is denoted by Z_g . The equations describing the operation of DVR are

$$L_f \frac{di_f}{dt} = v_i - v_c \quad (1)$$

$$C_f \frac{dv_c}{dt} = i_f - i_g \quad (2)$$

where i_g is the grid current, i_f is the inductor current, v_c is the compensation voltage, $v_i = uV_{dc}$ is the output voltage of VSI, u is the control input, V_{dc} is the chargeable dc power supply, L_f is the filter inductance, and C_f is the filter capacitance. The protection of sensitive load from the voltage disturbances existing in the grid can be achieved if the voltage on the series transformer (compensation voltage) is controlled. The relationship between the injected, grid, and load voltages can be written as

$$v_c(t) = v_g(t) - v_L(t). \quad (3)$$

It is clear from (3) that when there is no disturbance in the grid voltage (i.e., no voltage sags, swells, and harmonic distortion), the load voltage, which is required to be a sine wave with desired amplitude and frequency, is equal to the grid voltage. In this case, the compensation voltage is zero. However, in the case of voltage disturbances in the grid voltage, the DVR should inject a compensation voltage so that the load voltage is not affected from the disturbances occurring in the grid. In order to achieve this, a compensation voltage reference v_c^* generation (described in Section III-C) is essential.

Now, let us define the compensation voltage error x_1 and its derivative as

$$x_1 = v_c - v_c^* \quad (4)$$

$$x_2 = \dot{x}_1 = \dot{v}_c - \dot{v}_c^* \quad (5)$$

where $\dot{x}_1$ denotes the derivative of x_1 , $\dot{v}_c$ denotes the derivative of v_c . The behavior of the DVR in terms of x_1 and x_2 can be

written as

$$\dot{x}_1 = x_2 \quad (6)$$

$$\dot{x}_2 = \omega_0^2 (-x_1 + uV_{dc} + D(t)V_{dc}) \quad (7)$$

where $\omega_0^2 = 1/L_f C_f$ and $D(t)$ is the disturbance term defined as

$$D(t) = -\frac{L_f}{V_{dc}} \frac{di_g}{dt} - \frac{1}{V_{dc}} v_c^* - \frac{L_f C_f}{V_{dc}} \frac{d\dot{v}_c^*}{dt}. \quad (8)$$

It should be noted that $D(t)$ is a time-dependent function.

III. SMC AND VOLTAGE SAG/SWELL DETECTION

A. SMC With Maximum Existence Region

The sliding surface function is defined as

$$S = \lambda x_1 + x_2 \quad (9)$$

where λ is a positive sliding coefficient. The synthesis of S requires the calculation of x_1 and x_2 , while the calculation of x_1 requires measurement of v_c , the calculation of x_2 is possible either by measuring the capacitor current and multiplying it by $1/C_f$ or differentiating x_1 . In fact, the proposed controller was resulting in satisfactory performance when x_2 is calculated in both ways. Clearly, the value of C_f is needed if the former is preferred in calculating x_2 . This means that the robustness and matched uncertainty are not always guaranteed in a SMC-based approach. It is shown in [33] that any uncertainty in C_f (or a deviation from its nominal value) used in the controller does not have a considerable effect on the robustness of the system. However, it is worth to note that synthesizing S employing the capacitor current requires an additional current sensor. On the other hand, if the differentiation of x_1 is employed in calculating x_2 , there is no need to use additional sensor. Although, the differentiation amplifies the noise in the input signal, it resulted in a satisfactory performance in the experimental studies (see Section IV).

When the system enters into the sliding mode ($S = 0 \Rightarrow x_2 = -\lambda x_1$), the state variables are forced to move on the sliding surface toward the origin ($x_1 = 0$ and $x_2 = 0$). The sliding mode is described by the following first-order equation:

$$\dot{x}_1 = x_2 = -\lambda x_1 \quad (10)$$

Equation (10) represents a line passing through the origin with a slope equal to $-\lambda$. The time derivative of (9) can be written as

$$\dot{S} = \lambda \dot{x}_1 + \dot{x}_2. \quad (11)$$

Substituting (6) and (7) into (11) gives

$$\dot{S} = -\omega_0^2 x_1 + \lambda x_2 + \omega_0^2 V_{dc} (u + D(t)) \quad (12)$$

Since, $S = 0$ and $\dot{S} = 0$ in the sliding mode, then from (11) the following equation can be obtained:

$$\dot{x}_2 = -\lambda \dot{x}_1. \quad (13)$$

Substituting $\dot{x}_1 = x_2$ into (13) yields

$$\ddot{x}_1 + \lambda \dot{x}_1 = 0. \quad (14)$$

Fig. 2. Existence regions of the sliding mode.

The solution of (14) is given by

$$x_1(t) = x_1(0)e^{s_1 t} \quad (15)$$

where s_1 equals to $-\lambda$ and denotes the pole of sliding-mode dynamics. Clearly, the sliding mode does not depend on the filter parameters (ω_0^2) and disturbance $D(t)$. Despite the voltage sags or swells existing in the grid voltage, the compensation voltage error ($x_1(t)$) converges to zero in the steady state. This can be interpreted as the strong robustness of the SMC strategy against voltage sag and swells. Hence, the SMC requires only the selection of λ . However, selection of λ should be done such that the existence of sliding mode is guaranteed. Once the existence of sliding mode is guaranteed, the system trajectories are directed to the sliding line and slide along the sliding line toward the origin. To ensure the existence of sliding mode, the sliding surface function S and its derivative $\dot{S}$ should have opposite signs and satisfy the following condition [23], [35]:

$$S\dot{S} < 0. \quad (16)$$

If the control input u is defined as

$$u = -\text{sign}(S) \quad (17)$$

then, the existence conditions of the sliding mode can be derived as follows:

Case I: $S < 0 \Rightarrow u = 1$:

$$\dot{S} > 0 \Rightarrow l_1(t) = \dot{S} = -\omega_0^2 x_1 + \lambda x_2 + d_1(t) > 0. \quad (18)$$

Case II: $S > 0 \Rightarrow u = -1$:

$$\dot{S} < 0 \Rightarrow l_2(t) = \dot{S} = -\omega_0^2 x_1 + \lambda x_2 + d_2(t) < 0. \quad (19)$$

In (18) and (19), $d_1(t)$ and $d_2(t)$ are defined as

$$d_1(t) = \omega_0^2 V_{dc} + D(t) \quad (20)$$

$$d_2(t) = -\omega_0^2 V_{dc} + D(t). \quad (21)$$

Equations (18) and (19) denote two parallel lines which determine the boundaries of the existence regions in the (x_1, x_2) -plane as shown in Fig. 2. It is clear that the sliding line ($S = 0$) divides the (x_1, x_2) -plane into two regions. The sliding mode occurs on the segment of the sliding line that covers both regions (red line between S_1 and S_2) where the state trajectory is directed toward the sliding line by an appropriate switching action

($u = 1$ if $S < 0$ or $u = -1$ if $S > 0$) in each region. The existence region is determined by three lines $l_1(t) = 0$, $l_2(t) = 0$, and $S = 0$. The sliding-mode operation exists and starts immediately when the state trajectory hits the sliding line between S_1 and S_2 . Then, the system states are forced to slide along $S = 0$ toward the origin. Hence, the maximum existence region occurs when the distance between S_1 and S_2 is a maximum.

For stability reasons, the existence region should be kept as large as possible. Line $l_1(t)$ intersects x_1 and x_2 axes at points $L'_1 = d_1(t)/\omega_0^2$ and $L''_1 = -d_1(t)/\lambda$. Similarly, line $l_2(t)$ intersects x_1 and x_2 axes at points $L'_2 = d_2(t)/\omega_0^2$ and $L''_2 = -d_2(t)/\lambda$. Among these intersection points, only L''_1 and L''_2 are dependent on λ . The x_1 and x_2 intercepts of points S_1 and S_2 can be determined by solving (18) and (19) during the sliding mode ($S = 0$). Setting (9), (18), and (19) to zero and solving for x_1 and x_2 yields

$$S_1 = (x_1, x_2) = \left(\frac{\omega_0^2 V_{dc} + D(t)}{\omega_0^2 + \lambda^2}, \frac{-\lambda (\omega_0^2 V_{dc} + D(t))}{\omega_0^2 + \lambda^2} \right) \quad (22)$$

$$S_2 = (x_1, x_2) = \left(\frac{-\omega_0^2 V_{dc} + D(t)}{\omega_0^2 + \lambda^2}, \frac{\lambda (\omega_0^2 V_{dc} - D(t))}{\omega_0^2 + \lambda^2} \right). \quad (23)$$

Equations (22) and (23) imply that the points S_1 and S_2 are also dependent on λ . Hence, when there is a change in λ , the points L''_1 , L''_2 , S_1 , and S_2 are also changed accordingly, which clearly shows that the existence region is controlled with the selection of λ . It is clear from Fig. 2 that the maximum existence region occurs when the distance between S_1 and S_2 is maximum. After doing some algebraic operations, the distance between S_1 and S_2 ($\overline{S_1 S_2}$) can easily be obtained as

$$\overline{S_1 S_2} = \frac{2\omega_0^2 V_{dc} \sqrt{\lambda^2 + 1}}{\omega_0^2 + \lambda^2}. \quad (24)$$

It is obvious from (24) that $\overline{S_1 S_2}$ is time invariant. Now, let us investigate the effect of changing λ on $\overline{S_1 S_2}$. The plot of (24) versus λ obtained with the parameters used in the simulation and experimental system (see Table II) is depicted in Fig. 3. From Fig. 3(a), it can be seen that $\overline{S_1 S_2}$ increases and then decreases as λ is increased. It is clear from (24) that the value of $\overline{S_1 S_2}$ depends on V_{dc} as well. For the typical values of λ and V_{dc} , the maximum value of $\overline{S_1 S_2}$ is usually a large value. Taking derivative of (24) with respect to λ , setting the resulting equation to zero and solving for λ gives

$$\lambda_m = \sqrt{\omega_0^2 - 2} \quad (25)$$

where λ_m is the optimum sliding coefficient. Equation (25) gives the value of λ that corresponds to the maximum value of $\overline{S_1 S_2}$. Hence, the maximum existence region of the sliding mode can be guaranteed when λ is set to λ_m . Also, it can be observed from Fig. 3(b) that $\overline{S_1 S_2}$ increases if λ is increased within a range of $0 < \lambda \leq \lambda_m$ and vice versa. Therefore, selecting $\lambda = \lambda_m$ not only enlarges the existence region to its maximum, but also makes the dynamic response of the compensation voltage faster [see (15)]. However, it is worth to note that, using $\lambda < \lambda_m$ (or $\lambda > \lambda_m$) would slow down (or speed up) the dynamic response and may cause stability problems due to the reduced size of the existence region for the selected λ values as shown in Fig. 3.

Fig. 3. Plot of $\overline{S_1 S_2}$ versus λ . (a) $0 < \lambda < 5 \times 10^4$. (b) Enlarged view for $0 < \lambda < 10^4$.

However, when λ is set to λ_m , the optimized SMC becomes dependent on ω_0^2 , which affects the position of pole s_1 and $\overline{S_1 S_2}$ in case of uncertainty in ω_0^2 . Let us investigate the effect of variations in L_f and C_f on the desired λ and $\overline{S_1 S_2}$. The values of $\overline{S_1 S_2}$ and λ_m that correspond to the $\pm 10\%$ variations in L_f and C_f are computed from (24) and (25), respectively, and are recorded into Table I together with the percent changes in λ_m and $\overline{S_1 S_2}$. Clearly, the effect of parameter variations on λ_m and $\overline{S_1 S_2}$ is small when the variation occurs only in one component. In the case of simultaneous variations in both components, while λ_m deviates 9.09%/11.11% from its optimum value, the deviation in $\overline{S_1 S_2}$ is computed to be 0.46%/0.55%, which is totally negligible. This clearly shows that while $\pm 10\%$ variation in L_f and C_f changes the position of the closed-loop pole s_1 by 9.09% toward the origin (or 11.11% away from the origin), its effect on the size of the existence region is very small.

B. Double-Band Hysteresis Control

Due to the finite-time delay of the control computations, the direct implementation of (17) causes the VSI to operate with infinite switching frequency, which is not possible in practice. Since it is impossible to switch the control at infinite rate, chattering (high-frequency oscillations) exists at the origin. Chattering involves high control activity which may excite the unmodeled dynamics. Although the reduction and analysis of chattering have

TABLE I
EFFECT OF PARAMETER VARIATIONS ON λ_m AND $\overline{S_1 S_2}$

	L_f (mH)	C_f (μ F)	λ		$\overline{S_1 S_2}$	
			λ_m	Change in λ_m (%)	$\times 10^3$	Change in $\overline{S_1 S_2}$ (%)
No variation in L_f and C_f	0.7	50	5345.2	0	3207.1	0
$\pm 10\%$ variation in C_f	0.7	55/45	5096.5/5634.4	4.65/5.41	3203.5/3202.7	0.12/0.14
$\pm 10\%$ variation in L_f	0.77/0.63	50	5096.5/5634.4	4.65/5.41	3203.5/3202.7	0.12/0.14
$\pm 10\%$ variation in L_f and C_f	0.77/0.63	55/45	4859.3/5939.1	9.09/11.11	3192.6/3189.4	0.46/0.55

TABLE II
SYSTEM PARAMETERS

Description and Symbol	Value
Grid voltage amplitude, V_g	$230\sqrt{2}$ V
Grid impedance, Z_L	1 m Ω and 0.1 mH
Grid frequency, f_g	50 Hz
Series transformer turns ratio	1:1
DC voltage, V_{dc}	600 V
Filter inductance, L_f	0.7 mH
Filter capacitance, C_f	50 μ F
Sensitive load (series RL)	54 Ω and 30 mH
Sampling time, T_s	35 μ s

Fig. 4. Double-band hysteresis switching in one-cycle.

been proposed recently in [36] and [37], the popular approach to eliminate the chattering effect is to introduce a boundary layer around the sliding line. There are different types of boundary layer approaches such as the ideal saturation control, the practical relay control, and the practical saturation control [35]. In the ideal saturation control, although the system is driven to the boundary layer, the sliding mode does not exist since the trajectory is not forced to remain on the sliding line $S = 0$. In the case of practical saturation control, the hysteresis characteristic exists, but the sliding mode does not exist. Also, the analysis of the system under consideration with the practical saturation control is complicated. In this study, the practical relay control with hysteresis characteristic is adopted [35]. The hysteresis characteristic implies that the switching occurs on the hysteresis band widths ($S = -h$ and $S = +h$), which causes the existence of a nonideal sliding mode. An important consequence of using the practical relay control is that the solution of the system

under in the sliding mode can be determined. Generally, there are two types of hysteresis control methods: single- and DBHC. In [33], the switching frequency expressions for single- and DBHC methods are derived analytically, where the switching frequency plots are also presented for comparison. From this comparison, it can be noticed that the DBHC leads to a much smaller switching frequency than that of the single-band hysteresis control. Smaller switching frequency is the result of having a three-level voltage at the output of VSI (v_i). Therefore, in this study, the DBHC is employed to control the VSI and reduce the chattering effect.

Hence, the control input in (17) is replaced by the following function:

$$u = \begin{cases} \left. \begin{aligned} u_1 = 1 & \quad u_2 = 0 \quad u_3 = 0 \quad u_4 = 1 \text{ if } S \text{ hits } -h \\ u_1 = 0 & \quad u_2 = 1 \quad u_3 = 0 \quad u_4 = 1 \text{ if } S \text{ hits } 0 \end{aligned} \right\} & \text{during -ve cycle of } S \\ \left. \begin{aligned} u_1 = 1 & \quad u_2 = 0 \quad u_3 = 0 \quad u_4 = 1 \text{ if } S \text{ hits } +h \\ u_1 = 0 & \quad u_2 = 1 \quad u_3 = 0 \quad u_4 = 1 \text{ if } S \text{ hits } 0 \end{aligned} \right\} & \text{during +ve cycle of } S \end{cases} \quad (26)$$

where h is the hysteresis band width in S , u_1 , u_2 , u_3 , and u_4 denote the driving signals of the switching devices T_1 , T_2 , T_3 , and T_4 , respectively. As a consequence of replacing the sign function in (17) with (26), the sliding-surface function is forced to remain in the defined hysteresis bands and from this point of view, the DBHC approach is equivalent to the pseudo-SMC.

Fig. 4 shows a typical double-band hysteresis switching in one cycle [38]. When $S < 0$ in the half-cycle, the switching occurs for T_1 and T_2 on one leg of the inverter. Similarly, when $S > 0$ in the other half-cycle, the switching is done between T_3 and T_4 on the second leg of the inverter. At the start of T_{on} , T_1 is turned ON and T_2 is turned OFF when S hits the hysteresis boundary at $-h$. As soon as T_1 is turned ON, S changes its direction and moves toward zero. At the start of T_{off} , T_1 is turned OFF and T_2 is turned ON when S hits the hysteresis boundary at $S = 0$. The switching of T_3 and T_4 occurs in the similar way for $S > 0$ in the other half-cycle.

The idea of replacing the sign function by a hysteresis band plays an important role in controlling the switching frequency of the inverter. However, the relationship between the switching frequency and the hysteresis band together with other parameters is very crucial and should be investigated for a practical application. Here, the switching frequency computation is based on the assumption that the state variables x_1 and x_2 are negligibly small in the steady state. Taking this assumption into

consideration, (12) can be approximated as

$$\dot{S} \cong \omega_0^2 V_{dc} (u + D(t)). \quad (27)$$

Assuming that the sensitive load draws a sinusoidal current $i_L = I_L \sin(\omega t - \phi)$, the instantaneous switching frequency during half-cycle ($0 \leq \theta = \omega t \leq \pi$) can be derived as [33]

$$f_{sw} = \frac{1}{T_{on} + T_{off}} = \frac{\omega_0^2 V_{dc}}{h} (M \sin \theta - M^2 \sin^2 \theta) \quad (28)$$

where

$$M = \sqrt{M_1^2 + M_2^2 + 2M_1 M_2 \sin \phi} \quad (29)$$

$$M_1 = \frac{V_c^*}{V_{dc}} \left(1 - \frac{\omega^2}{\omega_0^2} \right) \quad (30)$$

$$M_2 = \frac{\omega L_f I_L}{V_{dc}} \quad (31)$$

where V_c^* is the amplitude of compensation voltage reference. In the derivation of (28), the disturbance term was taken as $D(t) = M \cos(\omega t - \pi/2)$. It should be noted that $T_{on} + T_{off}$ is the switching period for T_1 and T_2 in the half-cycle, where T_3 is OFF and T_4 is continuously ON. The overall switching period of the inverter in one-cycle is twice of $T_{on} + T_{off}$, which in turn results in half of f_{sw} . Hence, the average switching frequency in one-cycle can easily be obtained as [33]

$$f_{sw,av} = \frac{\omega_0^2 V_{dc} M}{2h} \left(\frac{2}{\pi} - \frac{M}{2} \right). \quad (32)$$

Equation (32) implies that the average switching frequency is dependent on many parameters such as h , ω_0 , V_{dc} , and M .

C. Determination of Voltage Sag/Swell and Harmonic Voltage

The associated control system is required to generate the reference load voltage (or reference grid voltage) waveform which will be processed to determine the amount of voltage sags/swells that occur on the grid [39]. As a first step, a sine wave template synchronized with the grid voltage can be obtained by dividing the measured grid voltage with the amplitude of the grid voltage determined by the peak detector as follows:

$$\sin(\omega t) = \frac{v_g(t)}{V_g}. \quad (33)$$

Then, the desired reference load voltage can easily be generated by multiplying the target amplitude $230\sqrt{2}$ with the generated sine wave template as follows:

$$v_L^*(t) = 230\sqrt{2} \sin(\omega t). \quad (34)$$

Thereafter, the voltage differences on the grid voltage can be determined by subtracting the reference load voltage from the measured grid voltage as

$$\Delta v_g(t) = v_g(t) - v_L^*(t). \quad (35)$$

Equation (35) determines the amount of voltage sag or voltage swell on the grid.

On the other hand, it is well known that the grid voltage is usually distorted by the voltage harmonics. In such a case, the control strategy should also be able to determine these harmonics and inject a compensation voltage with the same amplitude and 180° phase difference to those of the grid voltage harmonics so as to maintain a sinusoidal voltage at the load terminals. The determination of voltage harmonics is usually achieved by processing the grid voltage using special algorithms at the expense of increased complexity. In this study, a self-tuning filter (STF) is utilized for detecting the harmonics on the grid. A lot of work has been reported on the control of single- and three-phase SAPFs based on STF. However, the STF has only been applied to three-phase DVR system to reduce the effects of distorted grid voltages [40]. This method is obtained from the integration of the synchronous reference [41], and is defined as

$$H(s) = \frac{V_{xy}(s)}{U_{xy}(s)} = \frac{s + j\omega}{s^2 + \omega^2} \quad (36)$$

where $V_{xy}(s)$ and $U_{xy}(s)$ are the input and output signals of the STF algorithm, respectively. Using the inverse Laplace transform, it can be shown that the impulse response of the STF is given by

$$h(t) = \cos(\omega t) + j \sin(\omega t) = e^{j\omega t}. \quad (37)$$

It is clear that the STF is a filter with complex coefficients, describing a complex sinusoid. Given a cosine function on one of the inputs and a sine function on the other input, it adaptively seeks to generate a complex sinusoid. For the three-phase power system, the in-phase and quadrature-phase components (v_α and v_β) are made available via Clark (or Park) transform [42]. In the studied single-phase system, there is only one sinusoidal input. Therefore, there is a requirement for a second signal, which is 90° out of phase with the single-phase input signal. Biricik *et al.* [43] applied this approach to a single-phase SAPF. The main idea is to consider the sensed grid voltage as an in-phase component (sine function), $v_\alpha(t) = v_g(t)$. Then, the quadrature-phase component can be synthesised by phase-shifting $v_\alpha(t)$ as

$$v_\beta(t) = \sin(\omega t + 90^\circ) = \cos(\omega t). \quad (38)$$

In order to have unity magnitude, i.e., $|H(s)| = 0$ dB, a constant K is incorporated into (36) as

$$H(s) = \frac{V_{xy}(s)}{U_{xy}(s)} = K \frac{(s + K)j\omega}{(s + K)^2 + \omega^2}. \quad (39)$$

The undistorted in-phase and quadrature-phase voltage components ($\bar{v}_\alpha$ and $\bar{v}_\beta$) can be obtained by processing their distorted components (v_α and v_β) with the STF algorithm resulting in

$$\bar{v}_\alpha(s) = \frac{K}{s} (v_\alpha(s) - \bar{v}_\alpha(s)) - \frac{\omega}{s} \bar{v}_\beta(s) \quad (40)$$

$$\bar{v}_\beta(s) = \frac{K}{s} (v_\beta(s) - \bar{v}_\beta(s)) + \frac{\omega}{s} \bar{v}_\alpha(s). \quad (41)$$

The undistorted quadrature-phase voltage can be considered as the filtered grid voltage

$$\bar{v}_g(t) = \bar{v}_\alpha(t). \quad (42)$$

Fig. 5. Real-time laboratory set up with OPAL-RT.

Thereafter, the voltage harmonic components of the grid can be obtained by

$$\tilde{v}_g(t) = v_g(t) - \bar{v}_a(t). \quad (43)$$

Finally, the total compensation voltage reference can be determined by adding $\tilde{v}_g(t)$ to the voltage differences on the grid defined in (35) as

$$v_c^*(t) = \tilde{v}_g(t) + \Delta v_g(t). \quad (44)$$

Hence, the proposed control strategy injects the compensation voltage into the grid which tracks the reference given in (44) in the case of voltage sags/swells and distortions on the grid.

IV. SIMULATION AND EXPERIMENTAL RESULTS

The effectiveness and feasibility of the proposed SMC strategy have been verified by simulations and experiments. Simulations were carried out by MATLAB/Simulink. The experimental performance of the proposed SMC was observed in a real-time environment by using OPAL-RT real-time platform and its associated tools. This platform allows designers to test prototypes in a realistic environment by using the hardware-in-the loop approach [44]–[46]. The proposed SMC strategy modeled in Simulink is embedded into the OPAL-RT by using the RT-LAB software which communicates among the sensing boards, OP5142- reconfigurable board, and host computer. The field programmable gate array (FPGA) technologies such as Xilinx Spartan-3 FPGA can be easily incorporated into RT-LAB by using OP5142. The results are observed in a digital storage oscilloscope connected to OPAL-RT via mini-BNC probes. Fig. 5 shows the real-time laboratory setup using the OPAL-RT platform. The system parameters used in the simulation and experimental studies are given in Table II.

The sliding coefficient was set to $\lambda_m = \sqrt{\omega_0^2 - 2} = 5345.2$ in accordance with (25). The other control parameters were selected as $K = 100$ and $h = 25 \times 10^4$. The block diagram of the proposed SMC is shown in Fig. 6. The measured voltages (v_g , v_c , and v_L) are scaled down by a factor of 1/100. In the experimental system, the dc voltage V_{dc} is obtained from a battery group.

Fig. 6. Block diagram of the proposed SMC strategy.

Fig. 7. Simulated and experimental responses of v_g , v_c , and v_L for a voltage sag in the grid voltage from 230–120 V. (a) Simulation. (b) Experiment (CH1 (v_g): 200 V/div, CH2 (v_c): 200 V/div, CH3 (v_L): 200 V/div).

Fig. 8. Simulated and experimental responses of v_g , v_c , and v_L for a voltage sag in the grid voltage from 120–230 V. (a) Simulation. (b) Experiment (CH1 (v_g): 200 V/div, CH2 (v_c): 200 V/div, CH3 (v_L): 200 V/div).

Fig. 7 shows the simulated and experimental dynamic responses of the grid voltage v_g , compensation voltage v_c , and load voltage v_L before and during a voltage sag on the grid voltage from 230 to 120 V (rms). As can be seen in Fig. 7(a) and (b), the compensation voltage on the secondary side of transformer is quite close to zero before the occurrence of the voltage sag. When the voltage sag occurs, the compensation voltage is successfully generated by the proposed DVR and injected into the system through the series transformer. As a result of this, the voltage on the load terminals is not affected from the grid fluctuation.

Fig. 8 shows the simulated and experimental dynamic responses of the grid, inverter, and load voltages when the grid voltage is increased from 120 V (rms) to its nominal (desired) value 230 V (rms). From Figs. 7 and 8, it is evident that during the voltage fluctuations, the load voltage is dynamically

Fig. 9. Simulated and experimental responses of v_g , v_c , and v_L for a voltage swell in the grid voltage (270 V). (a) Simulation. (b) Experiment (CH1 (v_g): 200 V/div, CH2 (v_c): 200 V/div, CH3 (v_L): 200 V/div).

maintained at desired level without having any interruption on the load voltage and stability problem in the sliding mode. This shows the effectiveness of the proposed optimized SMC with maximum existence region and optimum sliding coefficient.

Fig. 9 shows the simulated and experimental steady-state response of v_g , v_c , and v_L for a voltage swell in the grid voltage (270 V). Despite the voltage swell in the grid voltage, the load voltage is maintained at the desired level which is 230 V (rms) and the sensitive load is protected from this voltage increment in the grid voltage. The compensation of this voltage swell is achieved successfully with the help of DVR by injecting a compensation voltage having 180°-phase difference to those of the grid voltage. Simulation and experimental results show a close agreement.

Fig. 10 shows the simulated and experimental steady-state response of v_g , v_c , and v_L under highly distorted grid voltage.

Fig. 10. Simulated and experimental responses of v_g , v_c , and v_L under highly distorted grid voltage. (a) Simulation. (b) Experiment (CH1 (v_g): 200 V/div, CH2 (v_c): 200 V/div, CH3 (v_L): 200 V/div).

It is obvious that the DVR injects the necessary compensation voltage to protect the sensitive load from the distorted grid voltage. Despite the distorted grid voltage, the load voltage is sinusoidal and maintained at desired level. THDs of the grid and load voltages were measured as 17.53% and 1.8%, respectively.

The spectrums of the grid and load voltages obtained experimentally are depicted in Fig. 11. It is clear from Fig. 11(a) that the distorted grid voltage contains 3rd-, 5th-, and 7th-order harmonics. However, these harmonics are suppressed in the load voltage as shown in Fig. 11(b).

Fig. 12 shows the simulated and experimental control signals (u_1 , u_2 , u_3 , and u_4) for the switching devices obtained with the DBHC for the case shown in Fig. 7. Clearly, no switching occurs when there is no voltage sag in the grid voltage. However, when a voltage sag occurs, the control signals are generated in such a way that the switching devices T_1 and T_2 are switched ON and OFF, while T_4 is always ON and T_3 is

Fig. 11. Experimental spectrums of v_g and v_L (50 Hz/div). (a) v_g . (b) v_L .

always OFF during the positive half-cycle. Similarly, T_3 and T_4 are switched ON and OFF, while T_2 is always ON and T_1 is always OFF during the negative half-cycle. It is obvious that such switching leads to a lower switching frequency than that obtained by the conventional single-band hysteresis scheme. Using the parameters listed in Table I, the average switching frequency is computed as 4.49 kHz, which is quite reasonable for such application.

Fig. 13 shows the simulated and experimental responses of sliding surface function in the steady state. It can be seen that the sliding surface function has three levels as described in Section III-B, verifying the correct operation of the proposed control strategy. Also, the simulation and experimental results are seen to be in good agreement.

Fig. 14 shows the experimental responses of grid, inverter, and load voltages for a step change in V_{dc} from 600 to 550 V and back to 600 V again during a voltage sag existing in the grid voltage. It is obvious that the inverter voltage is not affected from the variations in V_{dc} .

Fig. 12. Simulated and experimental driving signals for the switching devices T_1 , T_2 , T_3 , and T_4 for the case shown in Fig. 6. (a) u_1 and u_4 . (b) u_2 and u_3 . (c) u_1 and u_4 . (d) u_2 and u_3 .

Fig. 13. Simulated and experimental responses of S in the steady state. (a) Simulation. (b) Experiment.

Fig. 14. Experimental responses of v_g , v_c , and v_L for a step change in V_{dc} from 600 V to 550 V and back to 600 V again during a voltage sag in the grid. (CH1 (v_g): 200 V/div, CH2 (v_c): 200 V/div, CH3 (v_L): 200 V/div).

V. CONCLUSION

In this paper, the SMC of single-phase DVR with guaranteed maximum existence region is proposed. The proposed SMC guarantees a maximum existence region by employing an optimum sliding coefficient in the sliding surface function. The value of optimum sliding coefficient is determined analytically by analyzing the expressions which bound the existence region of the sliding mode. It is pointed out that the use of optimum sliding coefficient not only maximizes the existence region, but also achieves faster dynamic response. In addition, a DBHC scheme which ensures the switching of a transistor in the VSI during a half-cycle while it remains either ON or OFF in the other cycle is used to mitigate the switching frequency. The theoretical considerations and analytical results are verified through computer simulations and experimental results. Simulation and experimental results show that the proposed SMC strategy is quite successful in maintaining the desired sinusoidal voltage at load terminals in case of voltage sags, swells, and distortions in the grid. An important consequence of using the proposed SMC is that the load voltage is maintained at desired level with low THD, the existence region is guaranteed which leads to a faster dynamic response, the practical implementation is quite easy and low switching frequency operation is possible.

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