2017 American Control Conference
Sheraton Seattle Hotel
May 24–26, 2017, Seattle, USA

Filter-Based Control of an H-Bridge Inverter with Output *LC* Filter

Mohammad Mohebbi, *Student Member, IEEE*, Joseph Latham, Michael L. McIntyre, *Senior Member, IEEE,* and Pablo Rivera

***Abstract*—This paper presents a filter-based control scheme for an H-Bridge inverter with output *LC* filter. This approach relies only on a single output voltage measurement to reduce the system cost as well as measurement noise and disturbance injected by output current and/or inductor current measurements. To reduce the controller sensitivity to the system parameters, the proposed controller is developed for unknown system parameters. A Lyapunov stability analysis is utilized to demonstrate system stability. Experimental results demonstrate excellent voltage regulation, insensitivity to load variations, and low output voltage distortion as well as the stability of the system under both linear and nonlinear loads.**

I. Introduction

As inverter-based Distributed Generation (DG) [1]-[4], Vehicle to Grid (V2G) [5], Battery Energy Storage System (BESS) [6], [7] and Uninterruptable Power Supplies (UPS) [8]-[10] are more widely adopted, pulse width modulated (PWM) inverters have become more broadly utilized for DC to AC voltage conversion.

In stand-alone mode the local load is supplied by the inverter. Therefore generation of a high quality output voltage with low distortion and excellent voltage regulation as well as disturbance rejection are the essential requirements of the associated control system. Good transient response and insensitivity to the load and system parameter variations are other metrics in the performance evaluation of inverters, which also necessitates the use of high performance controllers.

Many control techniques such as proportional–resonant (PR) [11], [12], multiloop feedback control [13]-[15], dead-beat control [16] and repetitive control [17] have been proposed to control a single phase voltage-source inverter (VSI) in standalone mode. Although a single output voltage measurement is sufficient for the control of the inverter, to the best knowledge of the authors, the majority of the existing control approaches also require an inductor current measurement. Using two measurements gives these controllers improved system stability and dynamic performance through both output voltage and inductor current regulation. For example, a simple multiloop control technique utilizes two traditional Proportional, Integral, and Derivative (PID) controllers to regulate both output voltage and inductor current in the voltage and current control loops, respectively. Finite loop gain of the PID controller at the fundamental frequency and its sensitivity to the load variations have motivated combining other techniques such as frame transformation [18]-[20] and Load Current Feedback (LCF) [14],[21],[22] to the multiloop control scheme. These combinations alleviate shortfalls of the multiloop control scheme at the cost of more computational complexity resulting from signal transformations between frames and an extra current sensor for an output current measurement.

Nonlinear control techniques such as backstepping controller [23] and sliding mode controller have been shown to demonstrate good tracking performance. Discrete-time sliding mode control technique has been used in multi-loop feedback systems due to its overshoot-free tracking capability [24]. However, the dependency of these controllers to the knowledge of the system parameters limits their practical application.

In the majority of the control schemes presented for the control of VSI with output LC filter, at least two sensors are used to measure the output voltage and the inductor current. In practice this inductor current measurement has a significant amount of ripple and measurement noise resulting from the switching scheme. This noise and ripple are then propagated into the control algorithm adding noise and disturbance to the system. Some control schemes use capacitor current measurement instead of the inductor current measurement [14], [25]-[27] where the same problem remains. Also some works use an output current sensor in addition to the other two sensors [14], [19], [23], [25] to reduce the effect of the high frequency noise and ripple resulting from switching, utilization of a low-pass filter (LPF) is suggested. Addition of LPF introduces phase delays, which can have an adverse effect on the control schemes, which can limit any performance improvement.

In [28] and [29] a filter-based discontinuous tracking controller for a general class of nonlinear, multi-input/multi-output (MIMO) mechanical systems with no disturbances is presented. In the present work a modified filter-based control scheme is proposed which utilizes the known system structure of a second order linear system and compensates for unknown disturbances. This scheme removes the need for parameter knowledge by utilizing a robust algorithm comprising a nonlinear sliding term which compensates for parameter uncertainties.

In this paper a filter-based controller with only single output voltage measurement is presented to eliminate the need for costly current sensors to measure the inductor and/or output currents. The elimination of the sensor along

Authors are with the Electrical and Computer Engineering Dept. of the University of Louisville, Louisville, KY 40292 (corresponding email: m0mohe01@louisville.edu).

with the removal of current ripple and noise from the control algorithm provides an advantage over previous methods. The high frequency noise resulting from PWM switching is inherently filtered out of the output voltage measurement by the LC filter of the inverter. Also, to reduce the control sensitivity to the system parameters and compensate for parameter variation, the control scheme is developed for unknown system parameters. A Lyapunov stability analysis proves that the sinusoidal voltage tracking objective is achieved by the controller with all signals remaining bounded. Experimental results further validate this approach.

The remainder of this paper is organized as follows. The system model is presented in Section II. Section III presents the filter-based control development. Experimental results are presented in Section IV. Conclusions of this paper are given in the last section.

II. System Model

An H-Bridge inverter with an output LC filter as seen in Fig. 1 is used for DC to AC power conversion. Applying the state averaging method and unipolar PWM switching scheme, the average model for an H-Bridge inverter can be written as follows [30]:

$$L\dot{I}_L = V_{in}(D + d_o) - RI_L - V_o \tag{1}$$

$$C\dot{V}_o = I_L - I_o \tag{2}$$

where L, C, R are the inductance, capacitance and series resistance of the inductance, respectively. V_{in} is the input supply voltage, $D(t)$ is the PWM duty ratio, d_o is a constant unknown disturbance and $I_L(t)$ is the inductor current. $V_o(t)$, and $I_o(t)$ are the output voltage and output current, respectively. The objective of the control scheme is to design $D(t)$ such that $V_o(t) \to V_d(t)$ as $t \to \infty$, where $V_d(t)$ is the desired sinusoidal output voltage trajectory defined by amplitude, frequency and phase. Taking derivative of (2) and substituting for $\dot{I}_L(t)$ from (1) the following second order equation is obtained to represent the system dynamics of the inverter.

$$m\ddot{V}_o + a\dot{V}_o + V_o = V_{in}D + u_o \tag{3}$$

Where $m \triangleq LC$, $a \triangleq RC$, and the lumped disturbance u_o is defined as follows:

$$u_o \triangleq V_{in}d_o - RI_o - L\dot{I}_o. \tag{4}$$

III. Filter-Based Controller for Unknown System Parameters

For the control development, a general case in which the inverter parameters including L, C, R are unknown is considered. This is a practical approach as parameter values change over the life of operation. Also parameter tolerance can be a performance issue. To facilitate the control development, the following set of assumptions are made.

Assumption 1: L, C, R, are unknown and time varying, but limited in a specific range such that:

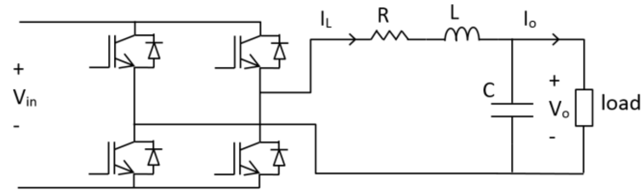

Figure 1. H-Bridge inverter with output LC filter.

$$\underline{m} < m(t) < \overline{m} \tag{5}$$

$$\underline{a} < a(t) < \overline{a} \tag{6}$$

Assumption 2: The rate of change of m with time is limited such that:

$$\dot{m}(t) < \overline{M}. \tag{7}$$

Assumption 3: The output voltage $V_o(t)$ is measurable.

Assumption 4: The input voltage V_{in} is known and constant.

Assumption 5: The load current and disturbance have the following properties: $I_o, \dot{I}_o(t), d_o \epsilon \mathcal{L}_\infty$.

Assumption 6: $V_d(t), \dot{V}_d(t), \ddot{V}_d(t), \dddot{V}_d(\mathrm{t}) \in \mathcal{L}_\infty$.

To facilitate the controller development and characterize its performance, the tracking error signal $e(\mathrm{t})$ and filtered error signals, $r_f(t), e_f(t)$ are defined as follows:

$$e \triangleq V_d - V_o \tag{8}$$

$$\dot{p} \triangleq -K_1 r_f + (K_2 + \alpha)(\alpha e - r_f) - e - e_f \tag{9}$$

$$r_f \triangleq p + (K_2 + \alpha)e \tag{10}$$

where K_1, K_2, α are positive gains, $p(t)$ is an auxiliary variable defined for filter implementation and $e_f(t)$ is defined with the following differential equation.

$$\dot{e_f} \triangleq r_f - \alpha e_f \tag{11}$$

To further the control development the following error signal is also defined:

$$\eta \triangleq \dot{e} + \alpha e - r_f \tag{12}$$

Taking derivative of (10) and using (9) and (12) results in:

$$\dot{r_f} = -K_1 r_f + (K_2 + \alpha)\eta - e - e_f \tag{13}$$

Taking derivative of (12) and utilizing (12), (13) and the second derivative of (8) after some mathematical simplifications results in:

$$\dot{\eta} = \ddot{V}_d - \ddot{V}_o + (K_1 + \alpha)r_f - K_2\eta - \alpha^2 e + e + e_f. \tag{14}$$

Multiplying both sides of (14) by m and substituting for $m\ddot{V}_o$ from (3), we get (15) after utilizing (8) and (12):

$$\begin{aligned} m\dot{\eta} = {} & m\ddot{V}_d + m(K_1 + \alpha)r_f - mK_2\eta - m\alpha^2 e \\ & + m(e + e_f) + a\dot{V}_d + a\alpha e - ar_f - a\eta \\ & + V_o - V_{in}D - u_o \end{aligned} \tag{15}$$

From (15) and motivated by the subsequent stability analysis the duty ratio control signal is defined as:

$$D \triangleq \frac{1}{V_{in}}[V_d + (K_2+\alpha)r_f - \hat{u}_0 + K_3 sgn(e-e_f)] \quad (16)$$

where K_3 is a positive constant gain, $sgn(.)$ is the standard signum function and $\hat{u}_0$ is the estimated disturbance with the following estimation error:

$$\tilde{u}_0 \triangleq u_o - \hat{u}_0. \quad (17)$$

Finally, the closed loop error system is obtained by substituting (16) in (15) as follows.

$$\begin{aligned} m\dot{\eta} &= N_d + \tilde{N} - mK_2\eta - a\eta - \frac{1}{2}\dot{m}\eta \\ &\quad -\tilde{u}_0 - e - (K_2+\alpha)r_f - K_3 sgn(e-e_f) \end{aligned} \quad (18)$$

with $N_d \triangleq m\ddot{V}_d + a\dot{V}_d$ and $\tilde{N}$ defined as:

$$\begin{aligned} \tilde{N} &= m(K_1+\alpha)r_f - m\alpha^2 e + m(e+e_f) + a\alpha e \\ &\quad -ar_f + \frac{1}{2}\dot{m}\eta \end{aligned} \quad (19)$$

where $\frac{1}{2}\dot{m}\eta$ is added to and subtracted from the right hand side of (19) and (18), respectively.

As the load is unknown and the load current is not measured we cannot directly account for the corresponding terms in u_o. However, we can make some simplifying assumptions to develop an appropriate observer based on the control implementation. In a PWM-VSI the switching and sampling frequency are typically orders of magnitude higher than the fundamental frequency. Therefore, in comparison with the sampling and switching frequencies, the output current and its derivative are changing very slowly, so that it can be approximated as a constant [31]. Using this fact, u_o can be approximated as:

$$\dot{\hat{u}}_0 \approx -\dot{\tilde{u}}_0. \quad (20)$$

Motivated by the subsequent stability analysis the update law for $\hat{u}_0$is defined as:

$$\dot{\hat{u}}_0 \triangleq -K_4\eta \quad (21)$$

where K_4 is a positive constant gain. As can be inferred from (12), η is not a measurable signal. But by taking the integral of (21) and substituting for η from (12) the update law for $\hat{u}_0$ becomes realizable as:

$$\hat{u}_0(t) = -K_4\left[\int_0^t \left(\alpha e(\tau) - r_f(\tau)\right)d\tau + e(t) - e(0)\right] \quad (22)$$

A. Stability Analysis

Before stating the main theorem, the following lemma is presented to be invoked later.

Lemma 1: Let the auxiliary function $L(t)$ be defined as follows:

$$L(t) \triangleq \eta(N_d - K_3 sgn(e-e_f)) \quad (23)$$

If K_3 is selected to meet the following gain condition:

$$K_3 > (|N_d| + \frac{1}{\alpha}|\dot{N}_d|) \quad (24)$$

then

$$\int_0^t L(\tau)d\tau \leq \zeta \quad (25)$$

where the positive constant ζ is defined as:

$$\zeta \triangleq |e(0)N_d(0)| + K_3|e(0)|. \quad (26)$$

Proof: See Appendix A of [29].

Theorem 2: Using the closed loop error system equation found in (18) the error signals defined in (8), (10), (11) and (12) are regulated as follows:

$e(t), e_f(t), r_f(t), \eta(t) \to 0$ as $t \to \infty$.

Proof: A non-negative Lyapunov function $S(t) \in \mathbb{R}$ is defined as

$$\begin{aligned} S &\triangleq \frac{1}{2}e^2 + \frac{1}{2}e_f^2 + \frac{1}{2}r_f^2 + \frac{1}{2}m\eta^2 \\ &\quad + \frac{1}{2K_4}\tilde{u}_0^2 + \zeta - \int_0^t L(\tau)d\tau \end{aligned} \quad (27)$$

Taking the derivative of (27) with respect to time and substituting $\dot{e_f}(t), \dot{e}(t), \dot{r_f}$(t) and $m\dot{\eta}(t)$ from (11), (12), (13) and (18), respectively, after some mathematical simplifications the expression in (28) is obtained where (20), (21) and (23) are also utilized.

$$\begin{aligned} \dot{S} &= -\alpha e^2 - \alpha e_f^2 - K_1 r_f^2 - K_2 m\eta^2 - a\eta^2 \\ &\quad +\tilde{N}\eta \end{aligned} \quad (28)$$

To proceed we first need to find an upper bound for $|\tilde{N}|$

$$\begin{aligned} |\tilde{N}| &\leq [\overline{m}(K_1+\alpha) + \overline{a}]|r_f| + \overline{m}|e_f| \\ &\quad +[\overline{m}\alpha^2 + \overline{m} + \overline{a}\alpha]|e| + \frac{1}{2}\overline{M}|\eta| \\ &= b_1|r_f| + b_2|e_f| + b_3|e| + b_4|\eta| \end{aligned} \quad (29)$$

where $b_1 \triangleq [\overline{m}(K_1+\alpha)+\overline{a}]$, $b_2 \triangleq \overline{m}$, $b_3 \triangleq [\overline{m}\alpha^2 + \overline{m} + \overline{a}\alpha]$ and $b_4 \triangleq \frac{1}{2}\overline{M}$.

Assuming $K_2 = \frac{1}{\underline{m}}(K_{21}+K_{22}+K_{23}+K_{24})$where K_{21}, K_{22}, K_{23}, K_{24} are all positive and using (29), $\dot{S}(t)$ can be upper bounded as:

$$\begin{aligned} \dot{S} &\leq -\alpha e^2 - \alpha e_f^2 - K_1 r_f^2 - a\eta^2 \\ &\quad +[b_1|\eta||r_f| - K_{21}\eta^2] + [b_2|\eta||e_f| - K_{22}\eta^2] \\ &\quad +[b_3|\eta||e| - K_{23}\eta^2] + (b_4 - K_{24})\eta^2. \end{aligned} \quad (30)$$

The three bracketed terms in (30), each represents nonlinear damping pairs that can be upper bounded as (31)-(33) [33].

$$b_1|\eta||r_f| - K_{21}\eta^2 \leq \frac{b_1^2 {r_f}^2}{K_{21}} \quad (31)$$

$$b_2|\eta||e_f| - K_{22}\eta^2 \leq \frac{b_2^2 {e_f}^2}{K_{22}} \quad (32)$$

$$b_3|\eta||e| - K_{23}\eta^2 \leq \frac{b_3^2 e^2}{K_{23}} \tag{33}$$

Assuming the gain conditions (34)-(37) are meet, $\dot{S}(t)$ can be upper bounded as (38).

$$K_{21} > \frac{b_1^2}{K_1} \tag{34}$$

$$K_{22} > \frac{b_2^2}{\alpha} \tag{35}$$

$$K_{23} > \frac{b_3^2}{\alpha} \tag{36}$$

$$K_{24} > b_4 \tag{37}$$

$$\dot{S} \leq -\beta_1 r_f^2 - \beta_2 e_f^2 - \beta_3 e^2 - \beta_4 \eta^2 - \alpha\eta^2 \tag{38}$$

Where $\beta_1 \triangleq K_1 - \frac{b_1^2}{K_{21}}$, $\beta_2 \triangleq \alpha - \frac{b_2^2}{K_{22}}$, $\beta_3 \triangleq \alpha - \frac{b_3^2}{K_{23}}$ and $\beta_4 \triangleq K_{24} - b_4$ are positive constants.

From (27) and (38) it is clear that $e(t), e_f(t), r_f(t), \eta(t) \in \mathcal{L}_2 \cap \mathcal{L}_\infty$ and $\tilde{u}_0 \in \mathcal{L}_\infty$. From (8) and the fact that $V_d(t)$ $\mathcal{L}_\infty$, therefore $V_o(t) \in \mathcal{L}_\infty$. From (4) and (17) along with Assumption 5 it is clear that $\hat{u}_0 \in \mathcal{L}_\infty$. Now from (16) along with $V_d(t) \in \mathcal{L}_\infty$, we can see that all the signals contributed in the definition of $D(t)$ are bounded, therefore $D(t) \in \mathcal{L}_\infty$. From (11)-(13) along with the previously stated bounding statements it is clear that $\dot{e}(t)$, $\dot{e_f}(t), \dot{r_f}(t) \in \mathcal{L}_\infty$. With $\dot{e}(t) \in \mathcal{L}_\infty$, (8) can be used to deduce $\dot{V}_o(t) \in \mathcal{L}_\infty$. From (3) it can be inferred that $\ddot{V}_o(t) \in \mathcal{L}_\infty$. Hence it is clear that all signals in the closed loop are bounded. From (14) it is clear that $\dot{\eta}(t) \in \mathcal{L}_\infty$. Since $e(t), e_f(t), r_f(t), \eta(t) \in \mathcal{L}_\infty \cap \mathcal{L}_2$ and $\dot{e}(t)$, $\dot{e_f}(t), \dot{r_f}(t), \dot{\eta}(t) \in \mathcal{L}_\infty$, Barbalat's Lemma [32] can be utilized to prove that $e(t), e_f(t), r_f(t), \eta(t) \to 0 \ as\ t \to \infty$. Thus completing the proof of the theorem.

IV. Experimental Results

The test rig used for verifying the performance of the proposed controller and observers is shown in Fig. 2. The NI CompactRIO, cRIO-9022, with cRIO-9113 chassis and the commercial software LabVIEW are used for implementation of the controller algorithm. The control algorithm is first developed using LabVIEW software on a personal computer and then downloaded to the onboard Virtex-5 LX50 FPGA of the cRIO-9113. The real-time experimental results were sent back to the personal computer through real time controller cRIO-9022 for monitoring and data logging. The dc link is fed by a single phase voltage doubler rectifier. The control algorithm requires only one voltage sensor to measure the output voltage, $V_o(t)$. For the purposes of data logging and visualization a current sensor is also utilized for the output current measurement. Table I summarizes the system parameters used for experimental test.

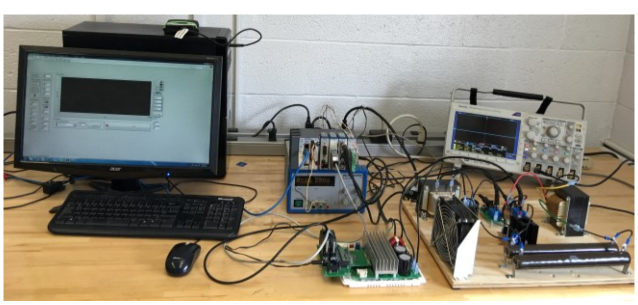

Figure 2. Experimental setup of the H-Bridge inverter.

TABLE I. System Parameters

	Parameter	*Value*	*Units*
Inverter	L	10	mH
	C	100	µF
	R	0.1	Ω
	V_o (peak-to-peak)	200	V
	V_{in}	350	V
	f	60	Hz
	Switching Frequency (f_{sw})	5	KHz
.Resistor-Inductor Load	L_{load1}	32	mH
	R_{load1}	37.5	Ω
Nonlinear Load	C_{load2}	220	µF
	R_{load2}	250	Ω
Controller Gain	K_1	20	-
	K_2	0.5	-
	K_3	10	-
	K_4	15	-
	α	0.5	-

In the first study, the steady state performance of the proposed control scheme under linear resistive-inductive load is investigated. Fig. 3 shows the desired, $V_d(t)$, and actual output voltage, $V_o(t)$, the tracking error, $e(t)$, and the output current as well as the control signal, $D(t)$. As can be seen in this figure, the excellent reference tracking with the steady-state peak error less than 1.45%, is achieved for the proposed control scheme.

A second study evaluates the performance of the proposed control scheme under a worst case operation scenario where a highly distorting load is used consisting of a full wave rectifier bridge feeding a 250 [Ω] resistor in parallel with a 220 [µF] capacitor. The results under nonlinear rectifier load are illustrated in Fig. 4. Despite highly distorted load current, the output voltage regulation with the steady-state peak error less than 2.15%, is achieved for the proposed control scheme.

A third study evaluated performance under no load operation of the inverter. Table II summarizes the results in terms of total harmonic distortion (THD) and steady-state error between the output voltage and its reference for

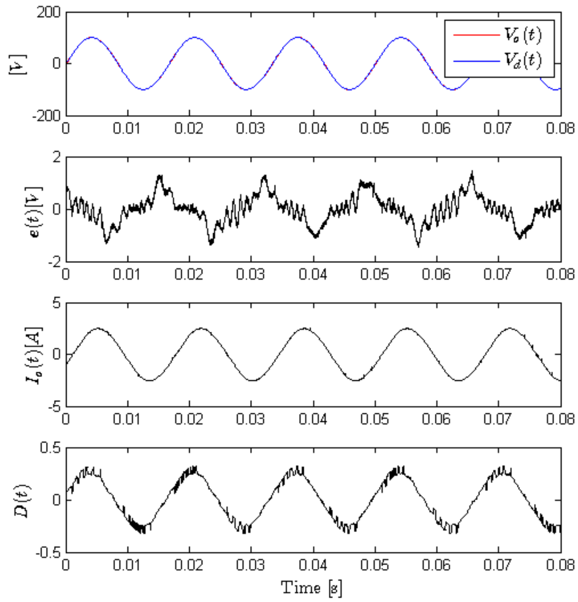

Figure 3. Steady-state results under *RL* load.

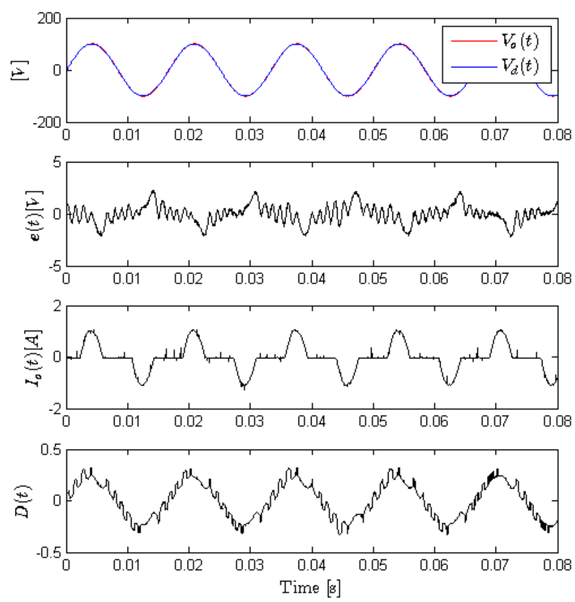

Figure 4. Steady-state results under highly distorting nonlinear load.

TABLE II. PERFORMANCE COMPARISION

Load Type	*Peak Error (%)*	*THD (%)*
No load	1.44	0.38
RL load	1.45	0.38
Highly nonlinear load	2.15	0.76

different test cases. As it can be seen in this table voltage THD is limited within 0.76% which fulfills IEEE 519 and EN 50160 standards for US and European power systems, respectively.

In a final study, the transient response to a -50% step change in amplitude of reference voltage, $V_d(t)$, under nominal 37.5 [Ω] resistive load is considered. As it can be

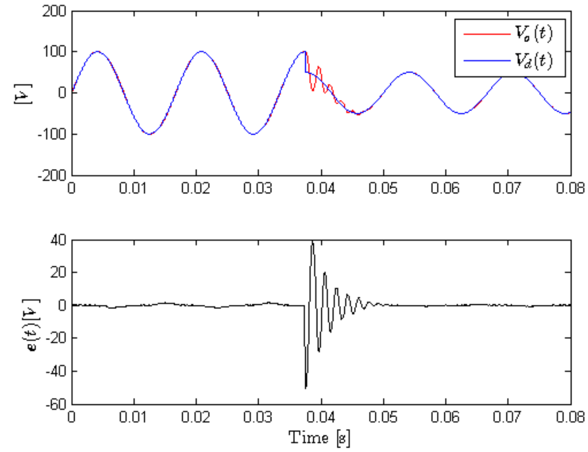

Figure 5. Transient results in response to -50% step change in amplitute of the reference voltage under resistive load.

seen in Fig. 5, to represent the worst case operation, the reference command is changed when the output voltage is at its peak value. Due to the excellent transient performance of the proposed control schemes, the output voltage recovers in less than half of a cycle.

V. CONCLUSION

In this paper a new filter-based control scheme relying on only a single output voltage measurement is proposed to regulate the instantaneous voltage of single-phase inverter in stand-alone mode. The performance of the control scheme is confirmed through experimental results in terms of steady-state tracking error, THD, stability as well as transient response. In addition to the lower cost resulting from removing current measurement sensors, the proposed control scheme has demonstrated its effectiveness in terms of low THD, excellent voltage regulation and insensitivity to load variation, even under a nonlinear load. The development of the control scheme for unknown system parameters makes it more attractive for its robustness against parameter variations in practical systems as the system parameters are subject to change during long term operation of the system.

REFERENCES

[1] R. J. Wai, C. Y. Lin, Y. C. Huang and Y. R. Chang, "Design of High-Performance Stand-Alone and Grid-Connected Inverter for Distributed Generation Applications," in *IEEE Transactions on Industrial Electronics*, vol. 60, no. 4, pp. 1542-1555, April 2013.

[2] M. E. Raoufat; A. Khayatian; A. Mojallal, "Performance Recovery of Voltage Source Converters With Application to Grid-Connected Fuel Cell DGs," in *IEEE Transactions on Smart Grid* , vol.PP, no.99, pp.1-8.

[3] E. Afshari, R. Rahimi, B. Farhangi and S. Farhangi, "Analysis and modification of the single phase transformerless FB-DCB inverter modulation for injecting reactive power," *2015 IEEE Conference on Energy Conversion (CENCON)*, Johor Bahru, 2015, pp. 413-418.

[4] S. Akhlaghi, A. A. Ghadimi and A. Akhlaghi, "A novel hybrid islanding detection method combination of SMS and Q-f for islanding detection of inverter- based DG," *2014 Power and Energy Conference at Illinois (PECI)*, Champaign, IL, 2014, pp. 1-8.

[5] M. Mohebbi, M. L. McIntyre and J. Latham, "Vehicle to grid utilizing a backstepping controller for bidirectional full-bridge converter and five level active neutral point inverter," *2015 IEEE 16th Workshop on*

Control and Modeling for Power Electronics (COMPEL), Vancouver, BC, 2015, pp.

[6] D. Velasco de la Fuente, C. L. Trujillo Rodríguez, G. Garcerá, E. Figueres and R. Ortega Gonzalez, "Photovoltaic Power System With Battery Backup With Grid-Connection and Islanded Operation Capabilities," in *IEEE Transactions on Industrial Electronics*, vol. 60, no. 4, pp. 1571-1581, April 2013.

[7] I. N. Moghaddam and B. Chowdhury, "Optimal sizing of Hybrid Energy Storage Systems to mitigate wind power fluctuations," *2016 IEEE Power and Energy Society General Meeting (PESGM)*, Boston, MA, 2016, pp. 1-5.

[8] R. Razi and M. Monfared, "Simple control scheme for single-phase uninterruptible power supply inverters with Kalman filter-based estimation of the output voltage," in *IET Power Electronics*, vol. 8, no. 9, pp. 1817-1824, 9 2015.

[9] H. Komurcugil, "Improved passivity-based control method and its robustness analysis for single-phase uninterruptible power supply inverters," in *IET Power Electronics*, vol. 8, no. 8, pp. 1558-1570, 8 2015.

[10] G. Willmann, D. F. Coutinho, L. F. A. Pereira and F. B. Libano, "Multiple-Loop H-Infinity Control Design for Uninterruptible Power Supplies," in *IEEE Transactions on Industrial Electronics*, vol. 54, no. 3, pp. 1591-1602, June 2007.

[11] G. Bonan, O. Mano, L. F. A. Pereira and D. F. Coutinho, "Robust control design of multiple resonant controllers for sinusoidal tracking and harmonic rejection in Uninterruptible Power Supplies," *2010 IEEE International Symposium on Industrial Electronics*, Bari, 2010, pp. 303-308.

[12] Keliang Zhou and Yunhu Yang, "Phase compensation multiple resonant control of single-phase PWM inverter," *Industrial Electronics (ISIE), 2012 IEEE International Symposium on*, Hangzhou, 2012, pp. 453-457.

[13] S. Buso, S. Fasolo and P. Mattavelli, "Uninterruptible power supply multiloop control employing digital predictive voltage and current regulators," in *IEEE Transactions on Industry Applications*, vol. 37, no. 6, pp. 1846-1854, Nov/Dec 2001.

[14] Poh Chiang Loh, M. J. Newman, D. N. Zmood and D. G. Holmes, "A comparative analysis of multiloop voltage regulation strategies for single and three-phase UPS systems," in *IEEE Transactions on Power Electronics*, vol. 18, no. 5, pp. 1176-1185, Sept. 2003.

[15] Poh Chiang Loh and D. G. Holmes, "Analysis of multiloop control strategies for LC/CL/LCL-filtered voltage-source and current-source inverters," in *IEEE Transactions on Industry Applications*, vol. 41, no. 2, pp. 644-654, March-April 2005.

[16] P. Mattavelli, "An improved deadbeat control for UPS using disturbance observers," in *IEEE Transactions on Industrial Electronics*, vol. 52, no. 1, pp. 206-212, Feb. 2005.

[17] Keliang Zhou, Kay-Soon Low, D. Wang, Fang-Lin Luo, Bin Zhang and Yigang Wang, "Zero-phase odd-harmonic repetitive controller for a single-phase PWM inverter," in *IEEE Transactions on Power Electronics*, vol. 21, no. 1, pp. 193-201, Jan. 2006.

[18] A. Roshan, R. Burgos, A. C. Baisden, F. Wang and D. Boroyevich, "A D-Q Frame Controller for a Full-Bridge Single Phase Inverter Used in Small Distributed Power Generation Systems," *APEC 07 - Twenty-Second Annual IEEE Applied Power Electronics Conference and Exposition*, Anaheim, CA, USA, 2007, pp. 641-647.

[19] D. Dong, T. Thacker, R. Burgos, F. Wang and D. Boroyevich, "On Zero Steady-State Error Voltage Control of Single-Phase PWM Inverters With Different Load Types," in *IEEE Transactions on Power Electronics*, vol. 26, no. 11, pp. 3285-3297, Nov. 2011.

[20] D. N. Zmood and D. G. Holmes, "Stationary frame current regulation of PWM inverters with zero steady-state error," in *IEEE Transactions on Power Electronics*, vol. 18, no. 3, pp. 814-822, May 2003.

[21] A. Kawamura and T. Yokoyama, "Comparison of five different approaches for real time digital feedback control of PWM inverters," *Industry Applications Society Annual Meeting, 1990., Conference Record of the 1990 IEEE*, Seattle, WA, USA, 1990, pp. 1005-1011 vol.2.

[22] Shih-Liang Jung, Meng-Yueh Chang, Jin-Yi Jyang, Li-Chia Yeh and Ying-Yu Tzou, "Design and implementation of an FPGA-based control IC for AC-voltage regulation," in *IEEE Transactions on Power Electronics*, vol. 14, no. 3, pp. 522-532, May 1999.

[23] M. Mohebbi, M. L. McIntyre and J. Latham, "Energy efficient DC to AC power conversion using advanced controllers and novel voltage trajectories," *Control and Modeling for Power Electronics (COMPEL), 2015 IEEE 16th Workshop on*, Vancouver, BC, 2015, pp. 1-8.

[24] Tsang-Li Tai and Jian-Shiang Chen, "UPS inverter design using discrete-time sliding-mode control scheme," in *IEEE Transactions on Industrial Electronics*, vol. 49, no. 1, pp. 67-75, Feb 2002.

[25] M. J. Ryan, W. E. Brumsickle and R. D. Lorenz, "Control topology options for single-phase UPS inverters," in *IEEE Transactions on Industry Applications*, vol. 33, no. 2, pp. 493-501, Mar/Apr 1997.

[26] N. M. Abdel-Rahim and J. E. Quaicoe, "Analysis and design of a multiple feedback loop control strategy for single-phase voltage-source UPS inverters," in *IEEE Transactions on Power Electronics*, vol. 11, no. 4, pp. 532-541, Jul 1996.

[27] M. J. Ryan and R. D. Lorenz, "A high performance sine wave inverter controller with capacitor current feedback and “back-EMF” decoupling," *Power Electronics Specialists Conference, 1995. PESC '95 Record., 26th Annual IEEE*, Atlanta, GA, 1995, pp. 507-513 vol.1.

[28] B. Xian, M. S. Queiroz, D. M. Dawson, and M. L. McIntyre, "A discontinuous output feedback controller and velocity observer for nonlinear mechanical systems," Automatica, vol. 40, no. 4, pp. 695–700, 2004.

[29] B. Xian, M. S. de Queiroz, D. M. Dawson and M. L. McIntyre, "Output feedback variable structure-like control of nonlinear mechanical systems," *Decision and Control, 2003. Proceedings. 42nd IEEE Conference on*, 2003, pp. 368-373 Vol.1.

[30] R. D. Middlebrook, “A general unified approach to modeling switching-converter power stages,” Int. J. Electronics, vol. 42, pp. 521-550, 1977.

[31] P. Cortes, G. Ortiz, J. I. Yuz, J. Rodriguez, S. Vazquez and L. G. Franquelo, "Model Predictive Control of an Inverter With Output *LC* Filter for UPS Applications," in *IEEE Transactions on Industrial Electronics*, vol. 56, no. 6, pp. 1875-1883, June 2009.

[32] J. Slotine and W. Li, *Applied Nonlinear Control*, Englewood Cliffs, NJ, Prentice Hall, 1991.

[33] M. Krstic, I. Kanellakopoulos, and P. Kokotovic, *Nonlinear and Adaptive Control Design*, John Wiley & Sons, 1995.