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A Fast and Cost-Effective Control of a Three-phase Stand-Alone Inverter

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***Abstract*—This paper proposes a straightforward control method for voltage control of a three-phase transformer-based inverter in uninterruptible power supplies or distributed generation systems. The approach offers a dual-loop design consisting inner current control loop and outer voltage loop. Sliding mode current controller provides desired bandwidth for voltage controller which consists of a state feedback term for stabilization and resonant term for harmonic damping. The proposed scheme provides fast dynamic response and low total harmonic distortion even for high power inverters with the limitations of switching frequency and LC filter components. Experimental studies for 2KVA linear and nonlinear loads using digital signal processors validate excellent performance of the proposed method.**

Keywords-voltage-source inverter;harmonic damping; resonant controller

I. INTRODUCTION

In recent years, growing demands for renewable energies and distributed generation systems increase attentions to stand-alone inverters and output voltage quality of these converters specially under nonlinear or computer loads. Fast Dynamic response and low THD of inverter voltage are challenging research area for UPS systems as well.

The cascade control method which consists of inner current control loop and outer voltage control loop is widely used in inverter voltage control systems which results in decreasing output impedance of these systems and decoupled output filter dynamics. However, it limits the bandwidth of voltage control loop which is very important for high power inverters with limited switching frequencies. For achieving fast dynamics, Proportional, dead-beat and sliding mode controllers are exploited as current controllers in [1, 2, 3].

Many control methods have been investigated to achieve appropriate dynamics for inverter output voltage while a few papers consider LC filter and switching frequency limitations of transformer-based systems. Special care must be taken to choose filter capacitors for transformer-based inverters considering costs, reactive power and output impedance [4].

Repetitive controller originating from the internal model principle is known as an effective solution for rejection of periodic errors in a dynamic system [5]. This method has been successfully applied to UPS systems in [5] and [6]. However, the control scheme is not simple and straightforward.

Adaptive voltage control scheme is performed in [7] and [8] which includes an adaptive compensating term for system uncertainties and a stabilizing term. In [4], input-Output feedback linearization is applied to nonlinear model of a three-phase UPS to make it linear and the tracking control is implemented based on pole placement technique. A model predictive control scheme is presented in [9] with high speed computations. The above schemes are presented for three-phase three-wire inverters but they should be extended for transformer-based and four-wire systems since the International Electrotechnical Commission Standard 62040-3 introduces single-phase rectifier loads as reference nonlinear loads in test requirements of UPS systems. Unlike the reviewed literature, a four-wire system is considered in this paper and its performance under single-phase reference loads is investigated which is applicable to UPS systems. Moreover, the method is simple and easy to implement.

Resonant controller methods have been shown excellent performance for selective harmonic damping of inverter voltage under nonlinear loads. In this approach, conventional control schemes such as state feedback [2] and proportional-integral PI controller [10] are designed for stabilizing and fast dynamic response of the system while resonant terms are

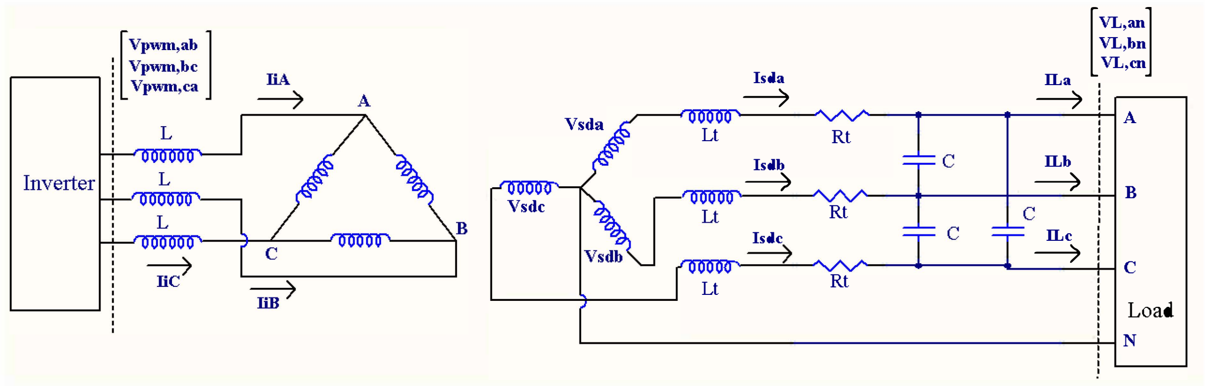

Figure 1. Three-phase inverter, filter chokes, delta-wye transformer, filter capacitors and load.

exploited to eliminate certain harmonics during several fundamental periods [11]. Therefore, this scheme which is also exploited in this paper can be successfully applied to high power inverters with switching frequency and bandwidth limitations.

In [12], a robust control method is proposed for a three-phase inverter with delta-star transformer. A fast digital sliding mode method is used as current controller while voltage control system consists of state feedback term and resonant terms. However, six filter capacitors are used before and after transformer which increase degree and cost of the system. In addition, the primary voltages of the transformer must be measured or observed.

This paper proposes a new modeling and design method based on the above approach for three-phase inverter by eliminating three filter capacitors and voltage sensors, thus providing reduced system order and less computational costs. In section II, state space equations of this system are obtained and section III describes the controller design procedure. The effectiveness of the proposed method is validated through experimental results.

II. System Description

Power circuit of a three-phase inverter with delta-why transformer is shown in Fig. 1. Using isolation transformer, output voltage leveling and neutral connection can be provided. The inverter phase currents are denoted as I_{iA}, I_{iB} and I_{iC}, the secondary transformer currents are denoted as I_{sda}, I_{sdb} and I_{sdc} and the load phase currents are denoted as I_{La}, I_{Lb} and I_{Lc}. It should be mentioned that the transformer is considered of D1 vector group.

Applying Kirchhoff's laws in transformer primary and secondary circuits and considering transformer turns ratio tr, the following equations are obtained:

$$C\begin{bmatrix}-2 & 1 & 1\\ 1 & -2 & 1\\ 1 & 1 & -2\end{bmatrix}\begin{bmatrix}\dot{V}_{LA}\\ \dot{V}_{LB}\\ \dot{V}_{LC}\end{bmatrix}+\begin{bmatrix}I_{sda}\\ I_{sdb}\\ I_{sdc}\end{bmatrix}=\begin{bmatrix}I_{La}\\ I_{Lb}\\ I_{Lc}\end{bmatrix} \quad (1)$$

$$\begin{bmatrix}\dot{I}_{iAB}\\ \dot{I}_{iBC}\\ \dot{I}_{iCA}\end{bmatrix}=\frac{-3tr}{3Ltr^2+L_t}\begin{bmatrix}V_{LA}\\ V_{LB}\\ V_{LC}\end{bmatrix}-\frac{R_t}{3Ltr^2+L_t}\begin{bmatrix}I_{iAB}\\ I_{iBC}\\ I_{iCA}\end{bmatrix}+\frac{3tr^2}{3Ltr^2+L_t}\begin{bmatrix}V_{pwm,AB}\\ V_{pwm,BC}\\ V_{pwm,CA}\end{bmatrix} \quad (2)$$

In these equations, V_{LA}, V_{LB} and V_{LC} are the load phase-to-neutral voltages; I_{iAB}, I_{iBC} and I_{iCA} are the line-to-line currents of transformer primary side (i.e. $I_{iAB}= I_{iA}-I_{iB}$); and $V_{pwm,AB}$, $V_{pwm,BC}$ and $V_{pwm,CA}$ are the inverter line-to-line voltages which are the control inputs of the system. Leakage impedances of the transformer are lumped at the secondary side and are denoted as L_t and R_t.

Applying transformer voltage and current equations and transforming to αβ0 stationary reference frame, the state space equations are obtained as:

$$\begin{cases}\dot{X}=AX+Bu+Ed\\ y=CX\end{cases} \quad (3)$$

$$X=\begin{bmatrix}V_L^{\alpha,\beta}\\ I_i^{\alpha,\beta}\end{bmatrix} \quad (4)$$

$$A=\begin{bmatrix}0 & \dfrac{1}{9Ctr}\\ \dfrac{-3tr}{3Ltr^2+L_t} & \dfrac{-R_t}{3Ltr^2+L_t}\end{bmatrix} \quad (5)$$

$$B=\begin{bmatrix}0\\ \dfrac{3tr^2}{3Ltr^2+L_t}\end{bmatrix} \quad (6)$$

$$E = \begin{bmatrix} \frac{-1}{3C} \\ 0 \end{bmatrix} \quad (7)$$

$$C = \begin{bmatrix} 1 & 0 \end{bmatrix} \quad (8)$$

where u is $V_{pwm}^{\alpha,\beta}$ and d is the load current considered as a disturbance.

III. Controller Design

This section describes control system approach which consists of voltage and current control loops. The voltage controller output provides reference inputs for digital sliding mode controller of the transformer primary currents.

A. Current control loop

The discretized state space model of the system for inverter current control based on (3) and zero order hold method is obtained as:

$$\begin{cases} X(K+1) = A_d X(k) + B_d u(k) + E_d d(k) \\ y(k) = C_i X(k) \end{cases} \quad (9)$$

$$C_i = \begin{bmatrix} 0 & 1 \end{bmatrix} \quad (10)$$

A sliding mode manifold for tracking of current reference $i_{cmd}(k)$ can be chosen as [12]:

$$s(k) = C_i X(k) - i_{cmd}(k). \quad (11)$$

The control input $u(k)$ is designed to be the solution of:

$$s(k+1) = C_i A_d X(k) + CB_d u(k) + CE_d d(k) - i_{cmd}(k) = 0 \quad (12)$$

Then the control law can be given by:

$$u(k) = (C_i B_d)^{-1} [i_{cmd}(k) - C_i A_d X(k) - C_i E_d d(k)] \quad (13)$$

Therefore, the reference voltages of SVM modulator are produced by sliding mode current controller. The feed-forward behavior of the controller provides a fast and accurate dynamic response for the current loop.

B. Voltage control loop

In order to design the voltage control system, the dynamics of inner current loop must be included in the system state space equations. Considering (9) and (13), the overall system equations can be written as:

$$\begin{cases} X(k+1) = A_p X(k) + B_p u_p(k) + E_d d(k) \\ y(k) = CX(k) \end{cases} \quad (14)$$

where,

$$A_p = A_d - B_d (C_i B_d)^{-1} C_i A_d, \quad (15)$$

TABLE I. System Parameters

Parameter	Value
Rated transformer power	5 KVA
DC bus voltage	300 V
Switching Frequency	7.143 KHz
Sampling Frequency (T_s)	70 µs
Output Filter Inductor (L)	1.2 mH
Output Filter Capacitor (C)	15 µF
Transformer Turns Ration (tr)	1.7765
Transformer Equivalent Leakage Inductance (L_t)	400 µH
Transformer Equivalent Leakage Resistance (R_t)	0.48 Ω

$$B_p = B_d (C_i B_d)^{-1}, \quad (16)$$

$$u_p(k) = i_{cmd}(k). \quad (17)$$

Considering system parameters in Table (1), controllability and observability of the system can be proved which satisfies the conditions described in [12] for designing the resonant and stabilizing controllers in voltage control loop. Since the primary side of the transformer has delta connection, control of zero sequence components of the output voltage is not possible. Moreover, voltage resonant controllers are designed for 1st, 5th and 7th harmonics.

Taking η as output vector of the resonant compensator and considering voltage regulation error $e = V_{ref} - V_{load}$, the dynamic equations of this compensator can be written as:

$$\dot{\eta} = A_s \eta + B_s e \quad (18)$$

where,

$$\eta = \begin{bmatrix} \eta_1 \\ \eta_5 \\ \eta_7 \end{bmatrix}, \quad (19)$$

$$A_s = \begin{bmatrix} A_{s1} & & \\ & A_{s5} & \\ & & A_{s7} \end{bmatrix}, \quad (20)$$

$$B_s = \begin{bmatrix} B_{s1} \\ B_{s5} \\ B_{s7} \end{bmatrix}, \quad (21)$$

$$\eta_i = \begin{bmatrix} \eta_{i1} \\ \eta_{i2} \end{bmatrix}, \quad (22)$$

$$A_{si} = \begin{bmatrix} 0 & 1 \\ -\omega_i^2 & 0 \end{bmatrix}, \quad (23)$$

$$B_{si} = \begin{bmatrix} 0 \\ 1 \end{bmatrix}, \quad (24)$$

and ω_i shows harmonic frequencies for i= 1, 5 and 7 [13].

By discretizing resonant compensator equations and combining them with plant model in (14), the augmented system equations is written as:

$$\hat{X}(k+1) = \hat{A}\hat{X}(k) + \hat{B}u_p(k) + \hat{E}_1 d(k) + \hat{E}_2 y_{ref}(k) \quad (25)$$

where the reference input is $y_{ref}(k)=V_{ref}(k)$, the augmented state vector is $\hat{X} = [X^T \quad \eta^T]$ and the coefficient matrices are:

$$\hat{A} = \begin{bmatrix} A_p & 0 \\ -B_{sd}C & A_{sd} \end{bmatrix}, \quad (26)$$

$$\hat{B} = \begin{bmatrix} B_p \\ 0 \end{bmatrix}, \quad (27)$$

$$\hat{E}_1 = \begin{bmatrix} E_d \\ 0 \end{bmatrix}, \quad (28)$$

$$\hat{E}_2 = \begin{bmatrix} 0 \\ B_{sd} \end{bmatrix}, \quad (29)$$

and A_{sd} and B_{sd} are discretized coefficients of (18) [12].

As demonstrated in [12], the stabilizing compensator to stabilize the augmented system is designed based on discrete-time linear quadratic state feedback (LQR) method which yields u_p as:

$$u_p = K^T \hat{X} = [K_0 \quad K_1] \begin{bmatrix} X_p \\ \eta \end{bmatrix}, \quad (30)$$

by minimizing the following cost function:

$$J = \sum_{k=0}^{\infty} \hat{X}(k)^T Q \hat{X}(k) + \varepsilon u_p(k)^T u_p(k). \quad (31)$$

Special care must be taken for selection of Q and ε weights. The weights correlated with plant states primarily determine transient response of the system while weights of resonant compensator states have considerable effects on tracking error and harmonic damping. Moreover, appropriate value for ε can prevent saturation or over-modulation problems in space vector modulation.

Fig. 2 demonstrates the block diagram of the designed control system. Based on system parameters described in Table (1), the closed-loop bode diagram of the system is shown in

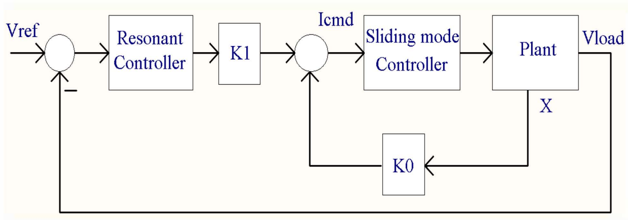

Figure 2. Block Diagram of voltage and current control systems

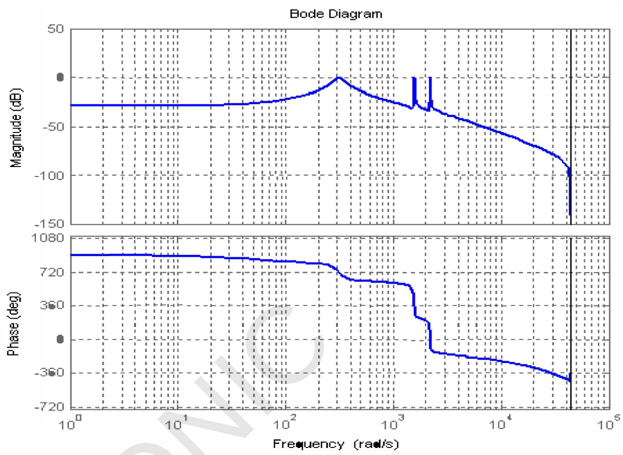

Figure 3. Bode plot of the closed loop voltage controlled system

Fig. 3 which demonstrates unity gain of the system at the desired frequencies.

IV. Experimental Results

The system of Fig. 1 has been practically implemented using the 32-bit fixed-point DSP-based controller TMS320F28335 for a 2KVA inverter with components and sample times as described in Table (1). Three 100Ω resistive loads and three single-phase rectifier loads are used as linear and nonlinear loads, respectively. The load components in different load step changes are implemented based on International Electrotechnical Commission Standard 62040-3.

A proportional-integral (PI) controller is also designed in synchronous reference frame such as [14] and is implemented with the same sample times in order to compare system performance and harmonic levels. Fig. 4 shows output phase voltages of the proposed system under resistive load. Total harmonic distortion of the waveforms is about 1.5%. It is worth mentioning that operation of the system under unbalanced resistive load (disconnection of one or two phases) is rather the same as Fig. 4 with the same THD which demonstrates excellent performance of the proposed method under unbalanced linear loads.

The performance of the proposed system under nonlinear loads is shown in Fig. 5 illustrating phase voltages and one phase current of load. Fig. 6 shows operation of PI control method under the same conditions. The results reveal excellent performance of the proposed method especially voltage peaks because of compensation of 5th and 7th harmonics. The voltage THD in the proposed approach is measured about 2.8% while in the PI control system is 5.1%. It is worth mentioning that 3rd voltage harmonic is not compensated in this system because of delta connection of transformer primary windings.

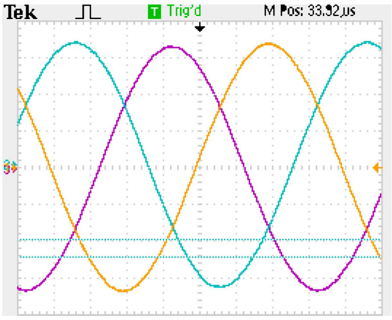

Figure 4. Experimental Results for load voltages. Operation of proposed control method under three single-phase resistive loads. Voltage Scale: 100V/div.

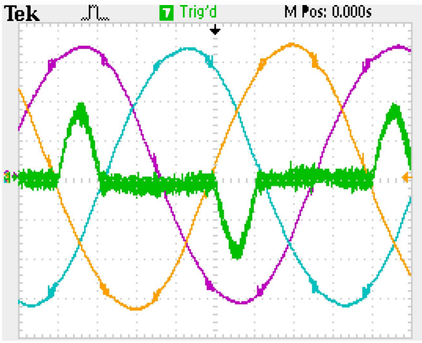

Figure 5. Experimental Results for load voltages and current. Operation of proposed method under three single-phase nonlinear loads. Voltage Scale: 100V/div, Current Scale: 5A/div.

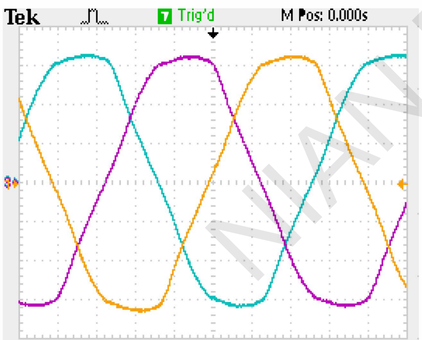

Figure 6. Experimental Results for load voltages. Operation of PI control method under three single-phase nonlinear loads. Voltage Scale: 100V/div.

TABLE II. PERCENTAGE OF VOLTAGE HARMONICS UNDER NONLINEAR LOAD IN PI AND PROPOSED SYSTEMS

Harmonic Order	Proposed Method (%)	PI Control (%)
2	0.2	0.26
3	2.04	2.32
4	0.09	0.1
5	0.08	3.57
7	0.06	0.95
9	0.78	0.69
11	0.34	0.58
13	0.29	0.34
THD	2.8%	5.1%

V. CONCLUSION

This paper has proposed a fast and straightforward control approach for control of three phase inverters in stationary reference frame with fast sliding mode current control and resonant compensation terms on 5th and 7th voltage harmonics. Modeling and control of the inverter with three-phase isolating transformer was demonstrated which is appropriate for UPS or distributed generation systems.

In order to improve previous works on this system, the proposed approach decreases number of filter capacitors and voltage sensors which results in reduction of system order and computational costs. Experimental studies on 2KVA linear and nonlinear loads exhibited superior performance of the proposed method in comparison with the conventional PI control approach. Output voltage harmonic levels in percentage of fundamental harmonic are shown in Table (2) for proposed and PI methods. This table demonstrates prominent reduction of the desired harmonics in the proposed method.

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