PERIODICA POLYTECHNICA SER. EL. ENG. VOL. 43, NO. 1, PP. 65–79 (1999)

PARK VECTOR BASED SLIDING MODE CONTROL OF UPS WITH UNBALANCED AND NONLINEAR LOAD

Péter KORONDI* and Hideki HASHIMOTO**

* Department of Automation
Technical University of Budapest
H–1111 Budapest, Budafoki út 8, Hungary
Tel: +36-1-463 2184 Fax: +36-1-463 3163 E-mail:korondi@elektro.get.bme.hu
** Institute of Industrial Science
University of Tokyo
7-22-1, Roppongi, Minato-ku Tokyo 106, Japan
Fax:+81-3-3423 1484, E-mail: hashimoto@iis.u-tokyo.ac.jp

Received: Oct. 14, 1999

Abstract

In this paper an inverter is taken as a member of Variable Structure System (VSS). A new Park vector based variable structure control (VSC) method is proposed.

A modified definition of the Park vector is introduced to handle the effect of zero phase-sequence component caused by an asymmetrical load. The design of a sliding mode controller consists of two main steps. Firstly, the design of the sliding surface, secondly, the design of the control which holds the system trajectory on the sliding surface. A complex sliding surface is proposed. The inverter is switched in such a way that the system trajectory gets as close to the sliding surface as possible. This paper focuses on the switching rule. Experimental results of a 100 KVA inverter are presented.

Keywords: sliding mode, UPS, unbalanced load, Park vector.

1. Introduction

The principle of pulsewidth modulation (PWM) plays a very important role in power electronics [1]. In the field of inverter technology, which produces sinusoidal voltage, a great number of 'optimized PWM' techniques have been proposed in the literature. These types of PWM inverters have very good steady-state characteristics, but the voltage regulator response to a sudden change in the load takes a few cycles, and nonlinear loads may cause high 'load harmonics'. This is not acceptable in Uninterruptable Power Supply (UPS) applications for which instantaneous feedback is preferred [2],[3]. In most papers symmetrical load is assumed. There is an increasing need to supply not only one unit, but also the whole installation by UPS. The energy supply for the group of units of great importance (nuclear power plants, airports, computer and telecommunication centres, etc.) is usually realized by three independent single phase UPS inverters, each having an independent control. This is because of the asymmetrical load of the phases due to the large number of possible single phase units connected to the network. A less expensive solution is to apply a single three phase inverter. In this case asymmetrical load causes problems because voltage in the three phases cannot be controlled independently. This paper proposes a three phase controller which can be used under asymmetrical and nonlinear loading.

In the course of designing the controller, the inverter is taken as a member of Variable Structure Systems (VSS) [4]. The output filter is considered as the controlled plant with its control realized by the inverter branches. It can be asymptotically stable even if it consists of unstable structures. A VSS system usually possesses an operational mode that is insensitive to parameter variations and load disturbances. This state is referred to as the sliding mode. The sliding mode is well known as a powerful tool to realize robustness in motion control systems and power electronics networks. The basic idea of the sliding mode control of a single input, single output (SISO) system can be summarized as follows [4]: It is supposed that the control is discontinuous and the reference signal is smooth enough. If the control undergoes discontinuities in some plane $\sigma(x) = 0$ in the state space then sliding mode may occur in this plane. Usually, $\sigma(x)$ is a linear function of the state variables. To ensure that the system trajectory remains on the sliding surface the condition

$$\dot{\sigma}(x)\sigma(x) < 0 \tag{1}$$

must hold. The robustness of ideal sliding mode control is obtained by infinite frequency switching of the control inputs resulting infinite gain. The price paid for insensitivity is – theoretically speaking – the infinitely high switching frequency. Due to the switching delays and frequency limit of controlled switches, ideal sliding mode does not exist. However, there is an acceptable approximation to ideal sliding mode control.

2. Configuration and Basic Equations

The main circuit and control system of the three phase inverter are illustrated in *Fig. 1*. The figure does not show the battery and charge unit. The filter circuit consists of a special Δ/Y transformer and a capacitor unit connected in parallel to the secondary. For the sake of filtering, the leakage inductance on the primary is increased and the mutual main field inductance is decreased by an air gap in contrast to an ordinary transformer. The load is connected between the star and the R-S-T terminals of the transformer.

The positive directions are shown in *Fig. 1*. This paper is focused on control, so a simplified transformer model is applied for theoretical consideration. Resistances of transformer coils and semiconductor devices are ignored. For the calculations, the transformer ratio of the transformer is chosen in such a way that leakage inductances on the secondary are equal to zero [13]. The primary leakage inductance matrix ($L_l \in R^{3x3}$) and the filter capacitor matrix ($C \in R^{3x3}$) are diagonal. The mutual inductance matrix ($L_m \in R^{3x3}$) is symmetrical. The single phase equivalent leakage and mutual main field inductance are L_{lp} and L_{mp}, respectively. The control system senses three electrical variables (namely the filter circuit output

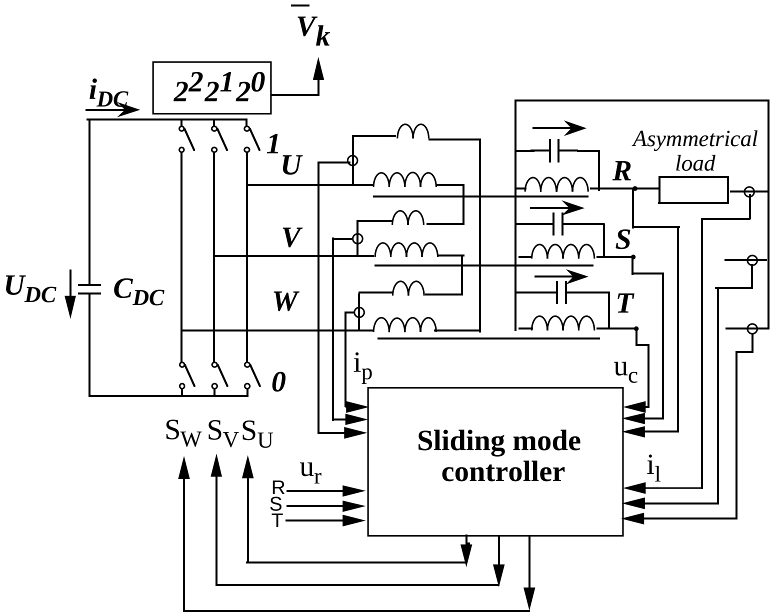


Fig. 1. Simplified diagram of the inverter

voltage, $u_c \in R^3$, the primary phase current, $i_p \in R^3$ and the load current $i_l \in R^3$), in each phase instantaneously. The sinusoidal reference signal, $u_r \in R^3$, is stored in ROM memory. The matrix differential equation for the filter circuit is as follows:

$$Gv = \Omega^{-2}\frac{d^2u_c}{dt} + u_c, \tag{2}$$

where $v \in R^3$ is the line-to-line voltage in the primary side,

$$\Omega^{-2} = L_m\left(L_m + L_l\right)^{-1} L_l C, \tag{3}$$

$$G = L_m\left(L_m + L_l\right)^{-1}. \tag{4}$$

The semiconductors of the inverter are controlled by a circuit which uses the reference signal and the three feed-back variables in each phase taking into account the switching frequency and current limitation requirements. The switches S_W, S_V and S_U are set to 1 if the upper transistor is switched on and 0 if the lower transistor is switched on in the corresponding inverter branch.

3. Sliding Mode Control of a Multi-Input Multi-Output System

The following non-linear state-equations are considered.

$$\dot{x} = f(x) + Bu, \quad x \in R^n, \quad u \in R^m, \quad f \in R^n, \tag{5}$$

where x is the state-variable vector; $f(x)$ is a vector-vector function; B is a positive or negative definite $n * m$ matrix; u is the control vector. The goal of the control is to drive the system into the origin in state space. The switching surfaces of the sliding mode, where control vector components have discontinuities, can be written in the following form

$$\sigma = \Lambda x, \quad \sigma \in R^m, \tag{6}$$

where Λ is an $m * n$ matrix. The condition for existence of sliding mode is

$$\sigma_i \dot{\sigma}_i < 0, \tag{7}$$

where σ_i is the i-th element of vector σ. The first derivative of vector σ with respect to the time is

$$\dot{\sigma} = \Lambda(f(x) + Bu). \tag{8}$$

If the matrix ΛB is not diagonal but its rank m, a T matrix can be introduced to decouple the system into m control loops

$$\sigma_t = T\sigma. \tag{9}$$

In other words $T\Lambda B$ should be diagonal in (10)

$$\dot{\sigma}_t = T\Lambda f(x) + T\Lambda Bu. \tag{10}$$

(It is remarked that matrix T can be time varying [6], in which case there is an additional term in (10)). The condition (7) must be true for each row of vector σ_t. That means that the sign of each row on the right hand side of (10) must be opposite to that of σ_t. If T exists then it can be chosen in a way so that the matrix $T\Lambda B$ is the identity matrix. Let us use a relay type control law

$$u_i = sign(\sigma_i) \tag{11}$$

The condition (7) holds if the absolute value of each element of vector u is bigger than those of vector $T\Lambda f(x)$.

Applying the above method to a three phase inverter, the error of the output voltage

$$u_e = u_r - u_c \tag{12}$$

is controlled. Using the above notation:

$$x = (u_{eR} \ \ u_{eS} \ \ u_{eT} \ \ \dot{u}_{eR} \ \ \dot{u}_{eS} \ \ \dot{u}_{eT})^T, \tag{13}$$

$$u = (S_u \ \ S_v \ \ S_w)^T U_{DC}, \tag{14}$$

$$f(x) = \begin{pmatrix} Z & I \\ \Omega^2 & Z \end{pmatrix} x + \begin{pmatrix} Z_c \\ (\Omega^2 - \omega^2 I_c) u_r \end{pmatrix}, \tag{15}$$

$$\Lambda = (I \quad \lambda I), \tag{16}$$

$$B = -\begin{pmatrix} Z \\ C^{-1} L_l^{-1} G \end{pmatrix} \begin{pmatrix} 1 & -1 & 0 \\ 0 & 1 & -1 \\ -1 & 0 & 1 \end{pmatrix}, \tag{17}$$

where ω is the output angular frequency $Z \in R^{3x3}$ and $Z_c \in R^3$ are the zero matrix, $I \in R^{3x3}$ is the identity matrix and

$$I_c = (1 \;\; 1 \;\; 1)^T.$$

Since det $B = 0$ the method of diagonalization cannot be applied directly. From the point of view of the circuit, it is because of the floating potential of the star point of the load.

Hence, although the inverter has three legs, it is impossible to introduce three independent switching surfaces to control the three phase errors independently. Several papers have solved this problem by introducing two independent surfaces with a third one, on which the system trajectory is automatically held.

4. A Park Vector Based Sliding Mode Control

4.1. Sliding Surface Design

To take the reduced degree of freedom into account a complex Park vector is introduced as a 3-phase to 2-phase transformation

$$\bar{x} = \frac{2}{3}(x_R + \bar{a} x_S + \bar{a}^2 x_T), \tag{18}$$

$$x_0 = \frac{1}{3}(x_R + x_S + x_T), \tag{19}$$

where x_R, x_S and x_T are the time functions of phase value of any three-phase signals and

$$\bar{a} = -\frac{1}{2} + \mathrm{j}\frac{\sqrt{3}}{2}, \quad \bar{a}^2 = -\frac{1}{2} - \mathrm{j}\frac{\sqrt{3}}{2} \tag{20}$$

The main advantage of the Park transformation for a system with symmetrical structure is that the two (real and imaginary) components are decoupled. The three phase values (x_R, x_S and x_T) can be asymmetrical, the system matrixes can be time varying. The only necessary assumption is that they should be symmetrical

[5] at any time instant. Using Park transformation, all the matrixes (including the non diagonal but symmetrical L_m) can be replaced by a scalar real value in (2) – (4). Expressing (2) in Park vector representation

$$\frac{d^2\bar{u}_c}{dt} = \Omega_p^2 G_p \bar{v} - \Omega_p^2 G_p \bar{u}_c, \tag{21}$$

where subscript p denotes the Park equivalent of the corresponding matrixes. Ω_p^2 and G_p are scalar real values

$$\Omega_p^{-2} = L_{mp}\left(L_{mp} + L_{lp}\right)^{-1} L_{lp} C_p, \tag{22}$$

$$G_p = L_{mp}\left(L_{mp} + L_{lp}\right)^{-1}. \tag{23}$$

A stationary reference frame is used. A complex sliding surface

$$\bar{\sigma} = \bar{u}_e + \lambda\dot{\bar{u}}_e = 0 + 0j \tag{24}$$

is introduced, where the control undergoes discontinuities. λ is a time constant type parameter, it defines the transient behaviour.

4.2. Switching Strategies

The inverter is switched so that the system trajectory gets as close to the sliding surface as possible. First, the location of the trajectory is measured. Second, the directions of the change of the trajectory according to the different switching states are derived. Finally, an appropriate switch condition is chosen to force the system trajectory to the sliding surface. If the system is not on the sliding surface, then vector σ indicates the distance of actual system trajectory from the sliding surface.

The direction of change of vector $\bar{\sigma}$

$$\dot{\bar{\sigma}} = \dot{\bar{u}}_e - \lambda\Omega_p^{-2}\bar{u}_e + \lambda\left[\left(\Omega_p^2 - \omega^2\right)\bar{u}_r - \Omega_{pd}^2 G_{pd}\bar{v}_k\right]. \tag{25}$$

Since error in steady state is only a small percentage of the rotating reference vector, $\bar{u}_r$, a good approximation of $\dot{\bar{\sigma}}$ is the difference between the rotating reference and the corresponding switching state vector as shown in *Fig. 2*

$$\dot{\bar{\sigma}}\; c_1\bar{u}_r - c_2\bar{v}_k, \tag{26}$$

where $c_1 = \lambda(\Omega_{pd}^2 - \omega^2)$, $c_2 = -\lambda\Omega_{pd}^2 G_{pd}$.

There are two main approaches of design of a control law for the sliding mode on the surface $\bar{\sigma} = 0 + 0j$. In the first approach, sliding mode exists only in the intersection of the switching surfaces. In this case, the condition for the existence of a sliding mode is that the vectors $\bar{\sigma}$ and $\dot{\bar{\sigma}}$ have opposite components.

$$Re(\bar{\sigma}\dot{\bar{\sigma}}^*) < 0, \tag{27}$$

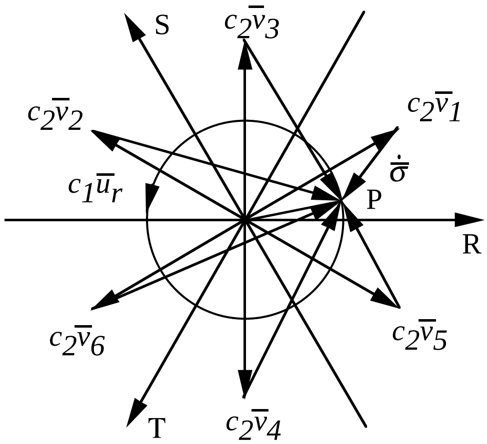


Fig. 2. Switching states and direction of vector change

where $\dot{\bar{\sigma}}^*$ stands for the complex conjugation of $\dot{\bar{\sigma}}$. In the second approach a stable sliding mode may exist on any of the two switching surfaces independently. In this case the condition for the existence of a sliding mode should hold for the two switching surfaces separately.

4.2.1. Triangular Switching Strategy

To adapt the most popular current vector control method [9] for sliding mode control of UPS, one period of the rotating vector, $\bar{u}_r$, is divided into six control intervals denoted by $T1$, $T2$, $T3$, $T4$, $T5$ and $T6$. Four switching states (i.e. only three directions of vector $\dot{\bar{\sigma}}$) are assigned to each control interval. In *Fig. 3* the switching states of control interval $T2$ are shown. Since the switching states are in the vertexes of a triangle that includes $\bar{u}_r$ this switching strategy is referred to as 'triangular' switching strategy. The complex $\bar{\sigma}$ plane is divided into six areas (a, b, c, d, e and f), as shown in *Fig. 3*. The six areas are joined in pairs in each control interval. In each pair of areas there is such an orientation of one of the three possible vectors $\dot{\bar{\sigma}}$ that the condition (27) holds, as shown in *Fig. 3*. Assuming, that the vector $\bar{u}_r$ is pointed at P (control interval $T2$) and the distance vector $\bar{\sigma}$ is in the area a or b, the switching state $\bar{v}_1$ must be switched. If the distance vector $\bar{\sigma}$ is in c or d then state $\bar{v}_3$ must be switched. If the distance vector is in e or f then either state $\bar{v}_0$ or $\bar{v}_7$ must be switched. Regarding the above mentioned facts, the triangular switching strategy is summarized in *Table 1*.

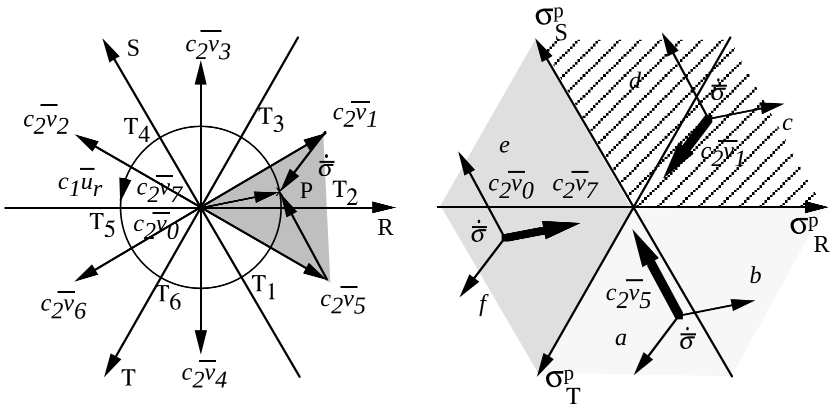


Fig. 3. Triangular switching strategy

Applying the triangular switching strategy some convergence problems emerge in the vicinity of the border of two control intervals discussed in [8] and [10]. In the case of current vector control, the change of vector $\bar{\sigma}$ is the difference between the rotating reference and the corresponding switching state vector. Because of the approximation (26) the convergence problem becomes more serious in the case of UPS control. To avoid these problems a rhomboid switching strategy is introduced [12].

4.2.2. Rhomboid Switching Strategy

In the second type of switching rule, the condition for the existence of a sliding mode should hold for two switching surfaces separately. Usually the two perpendicular components of the Park vector are controlled. Adapting this the switching surfaces are chosen as follows [6]

$$Im(\bar{\sigma}) = 0, \qquad Re(\bar{\sigma}) = 0. \tag{28}$$

These switching surfaces are not suitable for the inverter structure because the Park vector cannot be measured directly.

A simple switching strategy can be applied if only two of the three phase components of vector, $\bar{\sigma}$, are sensed and controlled simultaneously. (Remark: $x_R^p = x_R - x_0$, $x_S^p = x_S - x_0$ and $x_T^p = x_T - x_0$ are referred to as the phase components of the Park vector)

One period of the rotating vector, $\bar{u}_r$, is divided into six control intervals again. In each control interval the state of one branch is locked and only two branches are

Table 1. Triangular switching strategy

Control interval	Areas of the error vector	Switched state
$T1$	f, a	$\bar{v}_5$
	b, c	$\bar{v}_1$
	d, e	$\bar{v}_0, \bar{v}_7$
$T2$	a, b	$\bar{v}_1$
	c, d	$\bar{v}_3$
	e, f	$\bar{v}_0, \bar{v}_7$
$T3$	b, c	$\bar{v}_3$
	d, e	$\bar{v}_2$
	f, a	$\bar{v}_0, \bar{v}_7$
$T4$	c, d	$\bar{v}_2$
	e, f	$\bar{v}_6$
	a, b	$\bar{v}_0, \bar{v}_7$
$T5$	d, e	$\bar{v}_6$
	f, a	$\bar{v}_4$
	b, c	$\bar{v}_0, \bar{v}_7$
$T6$	e, f	$\bar{v}_4$
	a, b	$\bar{v}_5$
	c, d	$\bar{v}_0, \bar{v}_7$

switched. The control intervals and the six areas in the vector, $\bar{\sigma}$, plane are rotated to the previous case (a$'$ is used for the rhomboid strategy). *Fig. 4* shows the case of the control interval $T2'$. The two sensed phase components are σ_R^p and σ_T^p, the locked switch is $S_U = 1$ and the two switched branches are S_V and S_W as shown in *Fig. 4*. For example, if vector $\bar{\sigma}$ is in the area b', c' or d' ($\sigma_T^p < 0$) and the switch S_W is switched to zero then the vector $\dot{\bar{\sigma}}$ has a component which points to the switching line. After similar consideration the rhomboid switching strategy is summarized in the *Table 2*.

4.3. Comparison of Rhomboid and Triangular Switching Strategies

The rhomboid and the triangular switching strategies are compared by simulation. A switching-time delay (T_{del}) and a hysteresis were taken into the account. Nominal symmetrical load was considered.

The total harmonic distortion (THD) and the number of switches (NS) during one period are shown in Table 3.

Applying the triangular switching strategy a satisfactory reduction of distance could not always be achieved in the vicinity of the border of two control intervals.

Fig. 5 shows the operation of triangular and rhomboid switching strategies in the critical areas. In the case of triangular switching strategy, the trajectory of the vector, $\bar{\sigma}$, is chattering between the areas e and d. The amplitude of the vector, $\bar{\sigma}$, may not decrease since (26) gives only an approximation, the real direction of the vector $\dot{\bar{\sigma}}$ belonging to state $\bar{v}_{0,7}$ or $\bar{v}_1$ may have no component that is opposite to the vector, $\bar{\sigma}$. In the case of the rhomboid switching strategy the critical border is between the areas e' and d'. The trajectory of the vector, $\bar{\sigma}$, gets close to the origin after a few switches.

4.3.1. *Asymmetrical Load*

The Park vector does not contain information about the zero phase-sequence component caused by the asymmetrical load. According to the Park vector definition, the regular transformation

$$\begin{bmatrix} x_R^p \\ x_S^p \\ x_T^p \end{bmatrix} = \begin{pmatrix} \frac{2}{3} & -\frac{1}{3} & -\frac{1}{3} \\ -\frac{1}{3} & \frac{2}{3} & -\frac{1}{3} \\ -\frac{1}{3} & -\frac{1}{3} & \frac{2}{3} \end{pmatrix} \begin{bmatrix} x_R \\ x_S \\ x_T \end{bmatrix} \tag{29}$$

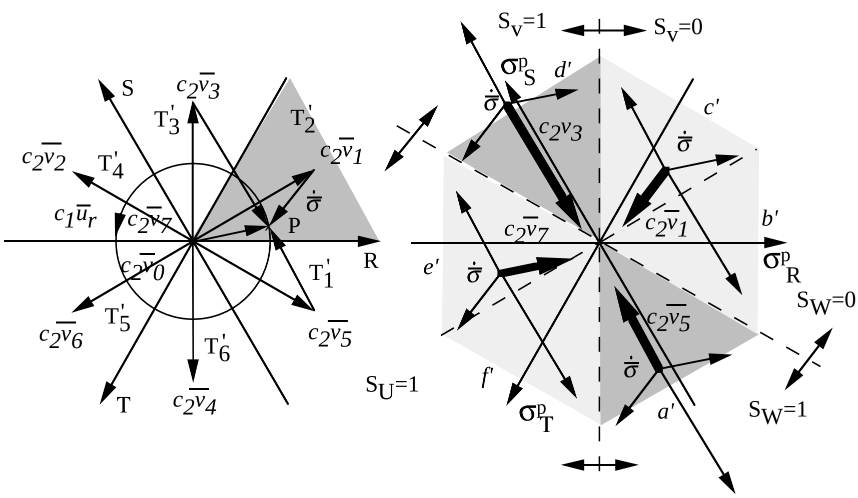


Fig. 4. Rhomboid switching strategy

Table 2. Rhomboid switching strategy

Control interval	Switching law
$T1'$	$S_U = 0.5 + 0.5 sign \bar{\sigma}_R^p$ $S_V = 0$ $S_W = 0.5 - 0.5 sign \bar{\sigma}_S^p$
$T2'$	$S_U = 1$ $S_V = 0.5 - 0.5 sign \bar{\sigma}_R^p$ $S_W = 0.5 + 0.5 sign \bar{\sigma}_T^p$
$T3'$	$S_U = 0.5 - 0.5 sign \bar{\sigma}_T^p$ $S_V = 0.5 + 0.5 sign \bar{\sigma}_S^p$ $S_W = 0$
$T4'$	$S_U = 0.5 + 0.5 sign \bar{\sigma}_R^p$ $S_V = 1$ $S_T = 0.5 - 0.5 sign \bar{\sigma}_S^p$
$T5'$	$S_U = 0$ $S_V = 0.5 - 0.5 sign \bar{\sigma}_R^p$ $S_W = 0.5 + 0.5 sign \bar{\sigma}_T^p$
$T6'$	$S_U = 0.5 - 0.5 sign \bar{\sigma}_T^p$ $S_V = 0.5 + 0.5 sign \bar{\sigma}_S^p$ $S_W = 1$

which transforms the phase values into the Park vector phase component is modified. To control the zero phase-sequence component, a K constant is introduced in the following way:

$$\begin{bmatrix} x_R^{p'} \\ x_S^{p'} \\ x_T^{p'} \end{bmatrix} = \begin{pmatrix} \frac{3-K}{3} & -\frac{K}{3} & -\frac{K}{3} \\ -\frac{K}{3} & \frac{3-K}{3} & -\frac{K}{3} \\ -\frac{K}{3} & -\frac{K}{3} & \frac{3-K}{3} \end{pmatrix} \begin{bmatrix} x_R \\ x_S \\ x_T \end{bmatrix}. \tag{30}$$

If $0 < K < 1$, it can be considered as negative feedback from the zero phase-sequence component. If the load is balanced then constant, K, has no effect on the control since

$$x_R + x_S + x_T = 0. \tag{31}$$

Table 3. Comparison of rhomboid and triangular switching strategies

	$T_{del} = 4\mu$		$T_{del} = 7\mu$	
	Triangular	Rhomboid	Triangular	Rhomboid
THD_R	0.515%	0.203%	0.757%	0.229%
THD_S	0.462%	0.207%	0.706%	0.237%
THD_T	0.588%	0.205%	0.713%	0.210%
NS_U	64	34	36	32
NS_V	82	34	32	32
NS_W	58	34	34	32

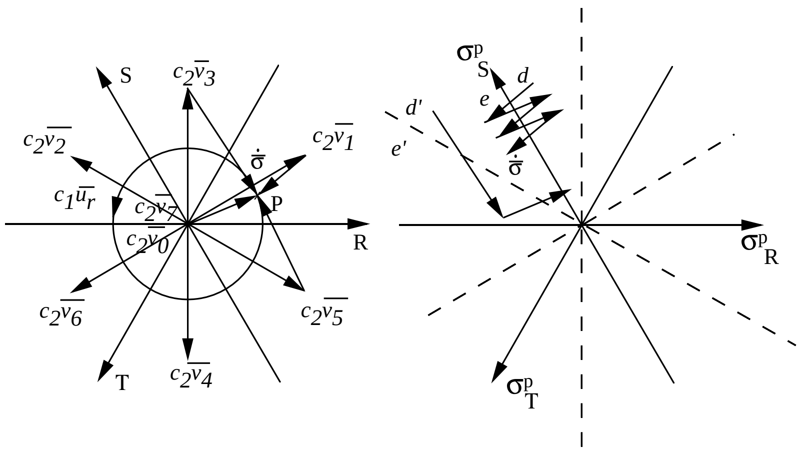


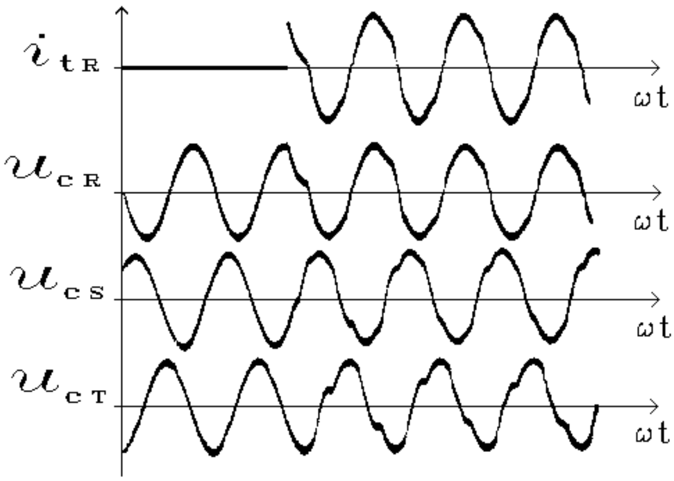
Fig. 5. Switching phenomenon in the vicinity of the border of two control intervals

5. Experimental Result

The parameters of the experimental system are given in *Table 4*.

Table 4. Nominal parameters

Nominal parameters		
Output power	P_n	100 [kVA]
Output voltage	U_c	220 [V]
Output frequency	f	50 [Hz]
Switching frequency	f_{sw}	2 [kHz]



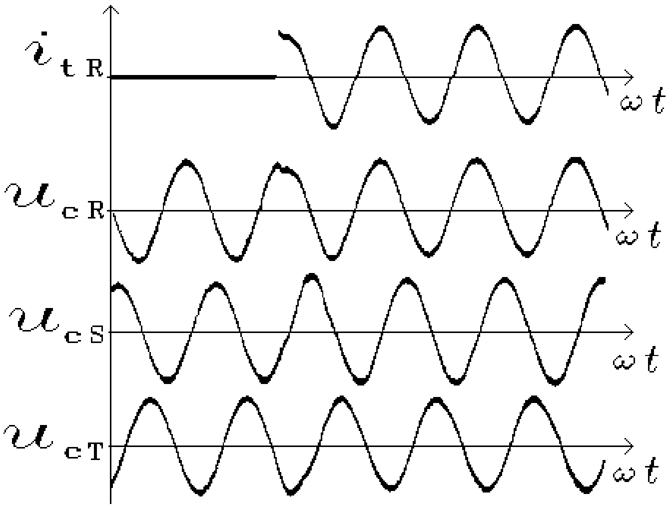
K	$=$	0	I_{lR}	$=$	120A			
U_{cR}	$=$	219 V	U_{cS}	$=$	221 V	U_{cR}	$=$	214 V
THD_R	$=$	4.3%	THD_S	$=$	6.3%	THD_T	$=$	9.8%

Fig. 6. System response to 80% step change in load of phase *R*

Two measurements are compared in *Fig. 6* and *Fig. 7*. In both cases, the rhomboid switching strategy is applied and the load of phase R is changed from 0 to 80% of the nominal load. When $K = 0$, the two controlled phase voltages can perfectly follow the reference signals and the uncontrolled phase voltage can be sinusoidal as well but the uncontrolled phase voltage has phase angle and amplitude errors. At the beginning of each control interval, a transient phenomenon appears which increases the total harmonic distortion (THD) (see in *Fig. 6*). Increasing the value of K, the total harmonic distortions are decreased in all phases but the asymmetry in the RMS values is increased. The optimum was found at $K = 0.7$ when the asymmetry in phase amplitudes was less than 2% and the total harmonic distortions were about 3% (see in *Fig. 7*).

6. Conclusion

This paper has demonstrated that variable structure theory is a useful and practical tool by which inverters can be controlled. The proposed controller structure performs well but it does not need a microcomputer, DSP or transputer. The effect



K	$=$	0.7	I_{lR}	$=$	120 A			
U_{cR}	$=$	219V	U_{cS}	$=$	223V	U_{cR}	$=$	216V
$nTHD_R$	$=$	2.1%	THD_S	$=$	2.8%	THD_T	$=$	3.1%

Fig. 7. System response for 80% step change in load of phase *R*

of asymmetrical load cannot be perfectly eliminated. However, there is a near-optimal solution where the asymmetry in phase amplitudes and the total harmonic distortions is acceptable.

Acknowledgements

The authors wish to thank the Office of Higher Education Support Programs (MKM FKFP 1092/1997, the Control Research Group of the Hungarian Academy of Science and Japanese Society for Promotion of Science for their financial support.

References

[1] HOLTZ, J. (1992): Pulsewidth Modulation – A Survey, *IEEE Trans On Industrial Electronics* Vol. IE-39 no. 5 pp. 410–420.

[2] KAWAMURA, A.– HOFT, R.: Instantaneous Feedback Controlled PWM Inverter with Adaptive Hysteresis, *IEEE Trans.* Vol. IA-20, no, pp. 769–775.

[3] KAWAMURA, A. – YOKOYAMA, T. (1991): Comparison Five Control Methods for Digitally Feedback Controlled PWM Inverters, *4th European Conference on Power Electronics, Firenze Proc.* Vol. 2. pp. 35–40.

[4] UTKIN, V. I. (1993): Variable Structure Control Optimalization, Springer-Verlag.

[5] JARDAN, K. R. – DEWAN, S. B. – SLEMON, G. R. (1969): General Analysis of Three-Phase Inverters, *IEEE Trans. on Industry and General Applications*, IGA-5, No.6 , pp. 672–679.

[6] SABANOVIC, A. – BILALOVIC, F. (1989): Sliding Mode Control of AC Drives, *IEEE Trans.*, Vol. IA-25, no. 1, pp. 70–75.

[7] KORONDI, P. – NAGY, L. – NEMETH, G. (1991): Control of a Three Phase UPS Inverter with Unbalanced and Nonlinear Load, *4th European Conference on Power Electronics Firenze*, Vol. 3. pp. 3–180–3–184.

[8] KORONDI, P. (1993): Comparison of Two Types of Tolerance Band Controlled Converters, *IEEE ISIE'93 Budapest, Proc.* pp. 128–133.

[9] NABAE, A. – OGASWARA, S. – AKAGI, H. (1985): A Novel Control Scheme of Current Controlled PWM Inverter, *IEEE Conf. Record of Industry Applications Society* 1985. pp. 473–477.

[10] NAGY, I. (1993): Control of Bi-directional Power Conversion, *IEEE ISIE'93 Budapest*, Vol. pp. 176–182.

[11] NAGY, I. (1994): Novel Adaptive Tolerance Band Based PWM for Field-Orientated Control of Induction Machines, *IEEE Trans. Industrial Electronics*, Vol. IE-41.

[12] KORONDI, P. – YANG, S-H. – HASHIMOTO, H. – HARASHIMA, F. (1994): Pulse Modulated Variable Structure System Controller For Parallel Resonant Dual Converter, *IEE Japan Ann. Meeting*.

[13] RETTER, GY.: Matrix and Space Phasor Theory of Electrical Machines, Akadémiai Kiadó, Budapest, 1987.