A Permanent Magnet Synchronous Motor Drive by using a Robust Adaptive Control Strategy

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Abstract: This paper presents a robust adaptive control strategy for permanent-magnet synchronous motor ($PMSM$) drive systems. This control scheme is based on an adaptive pole placement control strategy integrated to a sliding mode control scheme. This control strategy is applied to the machine current control loop. With this non-standard control strategy, the $PMSM$ drive systems achieve a fast transient response and robustness to parametric uncertainties and disturbances occurred on the machine parameters owing to saturation, temperature changes and loading conditions. Simulation and experimental results are presented for showing the effectiveness of this proposed closed-loop current control strategy.

I. Introduction

Nowadays, $PMSM$ are gradually replacing standard dc drives in a large number of industrial applications. The $PMSM$ motor drives have full advantages such as compactness, efficiency, robustness, reliability, and shape adaptation to the working environment [1, 2]. Substantial efforts have been developed for achieving a suitable robust control technique which exploit the efficiency and the fast dynamic capabilities of the $PMSM$. There have been several papers describing applications of various strategies for current controllers [3–7]. With these control strategies, $PMSM$ obtains equivalent performance of a separately excited dc motor [6]. However, the control of a $PMSM$ is a difficult task because its dynamics is strongly nonlinear [6]. The problem might also arise on the machine electric parameter variation due the saturation, temperature changes and loading conditions [6]. For that reason, the use of an adaptive approach is an interesting alternative. Two approaches can be used for implementing adaptive controllers: the model reference adaptive control ($MRAC$) and the adaptive pole placement control ($APPC$) [8, 9]. $MRAC$ requires cancellation of the plant zeros in an effort to make the closed-loop plant transfer function equals to that of the reference model. $APPC$ presents the combination of a pole placement control law with a parameter estimator. This scheme can be used to control a wide class of LTI plants with unknown parameters [9]. Usually this control technique has been used in the indirect approach, because it is easy to design and are applicable to a wide class of LTI plants and, do not require to be minimum phase or stable. Recently, a hybrid control strategy, composed by the integration of the $MRAC$ and the sliding mode control (SMC) structure has been proposed in [10] to achieve a suitable transient response and robustness to parametric uncertainties. In this technique, the SMC is employed for reducing the complexity of the parameter identification of $MRAC$. However, it preserves the main drawbacks of the $MRAC$ control schemes.

This paper presents a different hybrid control structure. It is obtained by integrating the $APPC$ control law to the SMC for reducing the complexity of plant parameter estimation. It is named variable structure adaptive pole placement control ($VS-APPC$). This integration allows to aggregate the fast transient response and robustness to parametric uncertainties and non-modelled disturbances. This control strategy is applied to the $PMSM$ current control loop. Simulation and experimental results are presented for showing the effectiveness of this new closed-loop current control strategy.

II. PMSM Modelling

The voltage equations of the $PMSM$ in a rotor reference frame can be described by the following equations:

$$v_{sd}^r = r_s i_{sd}^r + l_{sd}\frac{di_{sd}^r}{dt} - Pl_{sq}\omega_r i_{sq}^r \quad (1)$$

$$v_{sq}^r = Pl_{sd}\omega_r i_{sd}^r + r_s i_{sq}^r + l_{sq}\frac{di_{sq}^r}{dt} + \omega_r\Phi_f \quad (2)$$

where v_{sd}, v_{sq}, i_{sd}^r and i_{sq}^r are dq axis stator voltages and currents, r_s is a rotor resistance; ω_r is the rotor angular speed; Φ_f is the permanent-magnet flux linkage; l_{sd}

and l_{sq} are the dq axis inductances and, P is the $PMSM$ number of pair of poles, respectively. The dynamics of the mechanical drive can be simply described by

$$\frac{J}{P}\frac{d\omega_r}{dt} + \frac{F}{P}\omega_r = T_e - T_l \quad (3)$$

where J is the rotor inertia, F is the viscosity coefficient, T_l is the load torque. The electromagnetic torque developed by the $PMSM$ is given by

$$T_e = k_t i^r_{sq} + \frac{k_t}{\Phi_f}(l_{sd} - l_{sq}) i^r_{sd} i^r_{sq} \quad (4)$$

in which k_t is the torque constant.

Equations (1), (2) and (4), show that the machine dynamic model is nonlinear, as it contains product terms such as machine speed ω_r with i^r_{sd} and i^r_{sq}. Moreover, these equations are strongly coupled. By using a vector control technique and a machine current feedback, the nonlinear model of the $PMSM$, can be linearized and simplified. According to the vector principle, the machine phase current i^r_{sd} is forced to be zero in order to orient all linkage flux in the d axis and to achieve maximum torque per ampere. As a result, the resultant machine model are simplified to:

$$\frac{di^r_{sq}}{dt} = \frac{1}{l_{sq}}\left(v^r_{sq} - r_s i^r_{sq} - \Phi_f \omega_r\right) \quad (5)$$

$$\frac{d\omega_r}{dt} = \frac{P}{J}\left(T_e - \frac{F}{P}\omega_r - T_l\right) \quad (6)$$

where $T_e = k_t i^r_{sq}$ derived from Eq. (4) for $i^r_{sd} = 0$.

From Eq. (6) it can be verified that the machine torque can be controlled by regulating the machine phase current i^r_{sq}.

III. Control System

Figure 1 presents the block diagram of the proposed $PMSM$ vector control strategy. Blocks $VS - APPC$ implements the proposed current control scheme. The reference current of the d-axis machine phase current is regulated to be zero ($i^{r*}_{sd} = 0$). The machine torque is controlled by regulating the phase current i^r_{sq} determined by a feedforward control strategy from the desired reference torque T^*_e. Both current controllers are implemented on the rotor reference frame. Block $e^{-j\delta r}$ transforms the variables on the rotor reference frame into stator reference frame. The rotor position δr is obtained by using of an absolute encoder.

From the equations (1) and (2) and considering that the both dq axis can be decoupled by a suitable control vector strategy, the $PMSM$ current-voltage transfer functions can be given by

$$\frac{I^r_{sdq}(s)}{V^r_{sdq}(s)} = \frac{b_{sd}}{s + a_{sd}} \quad (7)$$

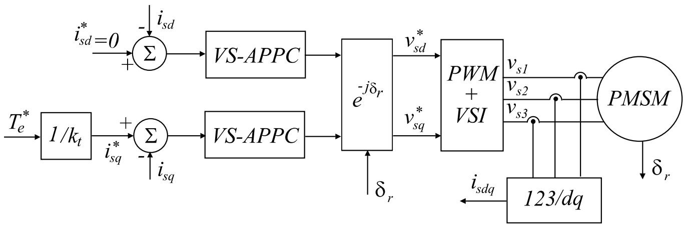

Fig. 1. Block diagram of the proposed $PMSM$ motor drive system.

where the parameters a_{sd} and b_{sd} are r_s/l_{sd} and $1/l_{sd}$ and, parameters a_{sq} and b_{sq} are r_s/l_{sq} and $1/l_{sq}$, respectively. The plant parameters are knowing with uncertainties that can be owing to machine saturation, temperature changes or loading variation.

A. Variable Structure Adaptive Pole Placement

The first approach of $VS - APPC$ was introduced in [11]. However, this approach do not deal with unmodeled disturbances occurred at the system control loop like machine $fems$. To overcome this, a modified $VS - APPC$ is proposed here.

Let us consider the first order $PMSM$ current-voltage transfer function given by Eq. (7). The main objective of the control strategy is to estimate parameters a_{sd}, b_{sd}, a_{sq} and b_{sq} to generate the inputs v^r_{sd} and v^r_{sq} so that the machine phase currents i^r_{sd} and i^r_{sq} following their respective reference currents i^{r*}_{sd} and i^{r*}_{sd} and, the closed loop poles are assigned to those of a given Hurwitz polynomials $A^*_{sdq}(s)$ given by

$$A^*_{sdq}(s) = s^2 + \alpha^*_{1sdq}s + \alpha^*_{0sdq} \quad (8)$$

where coefficients α^*_{1d}, α^*_{0d}, α^*_{1q} and α^*_{0q} determine the closed-loop performance requirements.

To estimate the parameters a_{sd}, b_{sd}, a_{sq} and b_{sq}, consider that the machine voltage-current first order plant for dq machine phases, on the stator reference frame can be described by

$$\dot{i}^s_{sdq} = -a_{sdq} i^s_{sdq} + b_{sdq} v^s_{sdq} \quad (9)$$

An adaptive law can be obtained for generating estimates $\widehat{a}_{sd}(\widehat{a}_{sq})$ and $\widehat{b}_{sd}(\widehat{b}_{sq})$ by using the observed signals v^s_{sdq} and i^s_{sdq}. Considering an arbitrary design constant $a_{md} > 0 (a_{mq} > 0)$, we can add and subtract the term $a_{mdq} i^s_{sdq}$ to Eq. (9) and rewrite the machine plant as

$$\dot{i}^s_{sdq} = -a_{mdq} i^s_{sdq} + (a_{mdq} - a_{sdq}) i^s_{sdq} + b_{sdq} v^s_{sdq} \quad (10)$$

Therefore, the estimative of the machine plant output can be obtained as

$$\dot{\hat{i}}^s_{sdq} = -a_{mdq} \widehat{i}^s_{sdq} + (a_{mdq} - \widehat{a}_{sdq}) i^s_{sdq} + \widehat{b}_{sdq} v^s_{sdq} \quad (11)$$

The estimation error can be determined by

$$e^s_{0sdq} = i^s_{sdq} - \widehat{i}^s_{sdq} \tag{12}$$

which satisfies the differential equation

$$\dot{e}^s_{0sdq} = -a_{mdq}e^s_{0sdq} + \widetilde{a}i^s_{sdq} - \widetilde{b}v^s_{sdq} \tag{13}$$

where $\widetilde{a}_{sdq} = \widehat{a}_{sdq} - a_{sdq}$ and $\widetilde{b}_{sdq} = \widehat{b}_{sdq} - b_{sdq}$ are parameters errors.

In the traditional indirect $APPC$ scheme, adaptive laws driven by the errors e^s_{0sdq} are used for generating estimates $\widehat{a}_{sd}$, $\widehat{b}_{sd}$, $\widehat{a}_{sq}$ and $\widehat{b}_{sq}$. For the proposed $VS-APPC$ strategy, the adaptive laws are replaced by suitable switching laws for reducing the complexity of the estimation of the machine current-voltage parameters. Therefore, by using the Lyapunov-like function

$$V(e^s_{0sdq}) = \frac{1}{2}(e^s_{0sdq})^2 > 0 \tag{14}$$

Admitting that parameters $\widehat{a}_{sd}$, $\widehat{b}_{sd}$, $\widehat{a}_{sq}$ and $\widehat{b}_{sq}$ can be estimated by the following switching laws given by

$$\widehat{a}_{sd} = -\overline{a}_{sd}sgn(e^s_{0sd}i^s_{sd}) \tag{15}$$

$$\widehat{b}_{sd} = \overline{b}_{sd}sgn(e^s_{0sd}v^s_{sd}) \tag{16}$$

$$\widehat{a}_{sq} = -\overline{a}_{sq}sgn(e^s_{0sq}i^s_{sq}) \tag{17}$$

$$\widehat{b}_{sq} = \overline{b}_{sq}sgn(e^s_{0sq}v^s_{sq}) \tag{18}$$

we have that

$$V(e^s_{0sdq}) \le -a_{md}(e^s_{0sdq})^2 \le 0 \tag{19}$$

since the following restrictions are satisfied $\overline{a}_{sd} > |a_{sd}|$ and $\overline{b}_{sd} > |b_{sd}|$ ($\overline{a}_{sq} > |a_{sq}|$ and $\overline{b}_{sq} > |b_{sq}|$) [11]. This guarantees that $e^s_{0sd} = e^s_{0sq} = 0$ and those are the globally asymptotically stable equilibrium points.

The pole placements and the tracking objectives of proposed $VS - APPC$ are achieved by using the following control laws

$$Q_{mdq}(s)L_{dq}(s)V^r_{sdq}(s) = -P_{sdq}(s)(I^r_{sdq} - I^{r*}_{sdq}) \tag{20}$$

which addresses to the implementation of both controllers transfer functions

$$C_{sdq}(s) = \frac{P_{sdq}(s)}{Q_{mdq}(s)L_{dq}(s)} \tag{21}$$

where $Q_{mdq}(s)$ is the internal model of reference currents i^{r*}_{sd} (i^{r*}_{sq}), $P_{sd}(s)$ ($P_{sq}(s)$) and $L_{dq}(s)$ are polynomials (with $L_{dq}(s)$ monic). $Q_{mdq}(s)$ is chosen to satisfy $Q_{md}(s)I^{r*}_{sd}(s) = Q_{mq}(s)I^{r*}_{sq}(s) = 0$. For the first order $PMSM$ current-voltage control plant (see Eq. (7)) and considering that the machine reference currents are constants, the choice of controller polynomials are $Q_{mdq}(s) = s$ (internal model of reference signal i^{r*}_{sd} (i^{r*}_{sq})), $L_{dq}(s) = 1$ and $P_{sd}(s) = \widehat{p}_{1sd}s + \widehat{p}_{0sd}$ ($P_{sq}(s) = \widehat{p}_{1sq}s + \widehat{p}_{0sq}$). By solving the Diophantine equation for desired Hurwitz polynomials $A^*_{sd}(s)$ ($A^*_{sq}(s)$), the coefficients $\widehat{p}_{1sd}$ and $\widehat{p}_{0sd}$ ($\widehat{p}_{1sq}$ and $\widehat{p}_{0sq}$) are determined by

$$\widehat{p}_{1sd} = \frac{\alpha^*_{1d} - \widehat{a}_{sd}}{\widehat{b}_{sd}} \tag{22}$$

$$\widehat{p}_{0sd} = \frac{\alpha^*_{0sd}}{\widehat{b}_{sd}} \tag{23}$$

for the d axis machine phase and

$$\widehat{p}_{1sq} = \frac{\alpha^*_{1q} - \widehat{a}_{sq}}{\widehat{b}_{sq}} \tag{24}$$

$$\widehat{p}_{0sq} = \frac{\alpha^*_{0sq}}{\widehat{b}_{sq}} \tag{25}$$

for the q axis. To avoid zero division on the Eqs. (22)-(25), the switching laws (16) and (18) are modified by

$$\widehat{b}_{sd} = \overline{b}_{sd}sgn(e^s_{0sdq}v^s_{sdq}) + b_{sd(nom)} \tag{26}$$

in which $b_{sd(nom)}$ is the nominal value of b_{sd} and the stability restriction is $\overline{b}_{sd} > \left|b_{sd} - b_{sd(nom)}\right|$. The same procedure is employed for obtaining the modified switching law related to Eq. (18).

The control signals v^r_{sd} and v^r_{sq} generated at the outputs of $VS-APPC$ controllers can be determined by using Eq. (21) as

$$\dot{v}^r_{sdq} = -\widehat{p}_{1dq}\dot{i}^r_{sdq} - \widehat{p}_{0dq}i^r_{sdq} + \widehat{p}_{0dq}i^{r*}_{sdq} \tag{27}$$

The block diagram of the $VS-APPC$ control algorithm for the d phase machine current control loop is presented in Fig. 2. Basically, it is composed by the machine parameters estimation scheme and by the $PMSM$ current control loop. The estimative of the parameters are implemented on the stator reference frame. This reference frame choice is used for increasing the accuracy of the estimation scheme. Therefore, at the output of the system controller and the output of the machine there are two blocks $e^{j\delta s}$ which realize the rotor-stator reference frame transformations. To accomplish the estimation scheme, the estimative of machine phase current $\widehat{i}^s_{sd}$ is compared with machine phase current i^s_{sd}, which generates the estimation error e^s_{0sd}. This error together with inputs $\widehat{v}^s_{sd}$ and $\widehat{i}^s_{sd}$, and set-points $\overline{a}_{sd}$, $\overline{b}_{sd}$ and $b_{sd(nom)}$ are used for calculating estimative $\widehat{a}_{sd}$ and $\widehat{b}_{sd}$ by using Eqs. (15) and (26).

These estimatives update the $PMSM$ model and are also used for calculating the controllers parameters $\widehat{p}_{1sd}$ and $\widehat{p}_{0sd}$, by using Eqs. (22) and (23). The block diagram related to the q machine phase controller can be obtained from this one by substituting subscript q in a place of d.

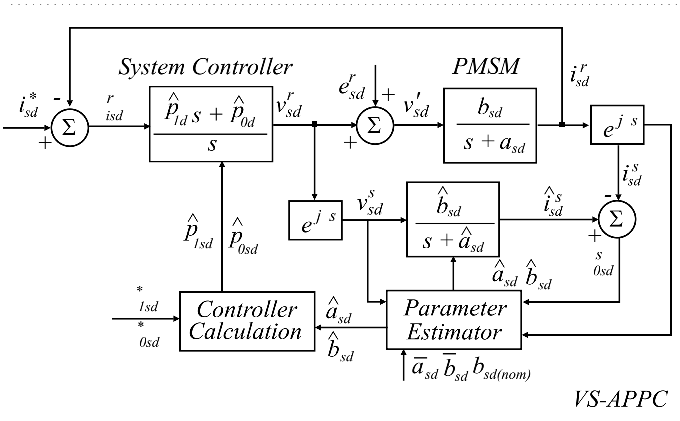

Fig. 2. Block diagram of the proposed $VS - APPC$ current controller.

B. Design of the controllers

To design the proposed $VS-APPC$ controller is necessary to choose suitable polynomials and to determine controllers coefficients $\widehat{p}_{1sd}$, $\widehat{p}_{0sd}$, $\widehat{p}_{1sq}$ and $\widehat{p}_{0sq}$. A good choice criteria for accomplishing the bound system conditions, is to define a polynomial which poles are closed to the control plant time constants. The characteristics of $PMSM$ used in this work are listed in the Table 1. This machine has different current-voltage transfer functions for dq phases because $l_{sd} \neq l_{sq}$ which are given by

$$\frac{I^r_{sd}(s)}{V_{sd}(s)} = \frac{526}{s+347} \tag{28}$$

for the machine current-voltage d phase and

$$\frac{I^r_{sq}(s)}{V^r_{sq}(s)} = \frac{454}{s+300} \tag{29}$$

for the machine q phase. Therefore, it is necessary to choose two different suitable polynomials $A^*_{sd}(s)$ and $A^*_{sq}(s)$ which are

$$A^*_{sd}(s) = (s+347)^2 \tag{30}$$

and

$$A^*_{sd}(s) = (s+300)^2 \tag{31}$$

According to Equations (15), (26) and (22)-(25) the estimative of the $VS - APPC$ current controllers can be obtained as

$$\widehat{p}_{1sd} = \frac{694 - \widehat{a}_{sd}}{\widehat{b}_{sd}} \tag{32}$$

$$\widehat{p}_{0sd} = \frac{120409}{\widehat{b}_{sd}} \tag{33}$$

for the d axis machine phase controller and

$$\widehat{p}_{1sq} = \frac{600 - \widehat{a}_{sq}}{\widehat{b}_{sq}} \tag{34}$$

$$\widehat{p}_{0sd} = \frac{90000}{\widehat{b}_{sq}} \tag{35}$$

for the q axis machine phase controller, respectively. To define the coefficients of the switching laws it is necessary to take into account together the stability restrictions $\overline{a}_{sd} > |a_{sd}|$ ($\overline{a}_{sq} > |a_{sq}|$) and $\overline{b}_{sd} > \left|b_{sd} - b_{sd(nom)}\right|$ ($\overline{b}_{sq} > \left|b_{sq} - b_{sq(nom)}\right|$). Based on the simulation and the theoretical studies, it can be observed that the magnitude of the respective switching laws ($\overline{a}_{sd}$, $\overline{b}_{sd}$, $\overline{a}_{sq}$ and $\overline{b}_{sq}$) determine how fast the $VS-APPC$ controllers converges to their respective references. However, the choice of greater values results in output signals of the system controllers (v_{sd} and v_{sq}) with high amplitudes. High values of output controllers which can addresses to the system to nonlinear behaviors. Thus, a good design criteria is to choose the parameters closed to average values of control plant coefficients a_{sd}, b_{sd}, a_{sq} and b_{sd}. Using this design criteria for the $PMSM$ employed in this work, the following values are obtained $b_{sd(nom)} = 500$, $b_{sq(nom)} = 430$, $\overline{b}_{sd} = 30$, $\overline{b}_{sq} = 30$, $\overline{a}_{sd} = 350$ and $\overline{a}_{sq} = 310$. This solution is not unique and different adjusts can be experimented for different applications. The performance of these controllers are evaluated by simulation and experimental results.

IV. Simulation and Experimental Results

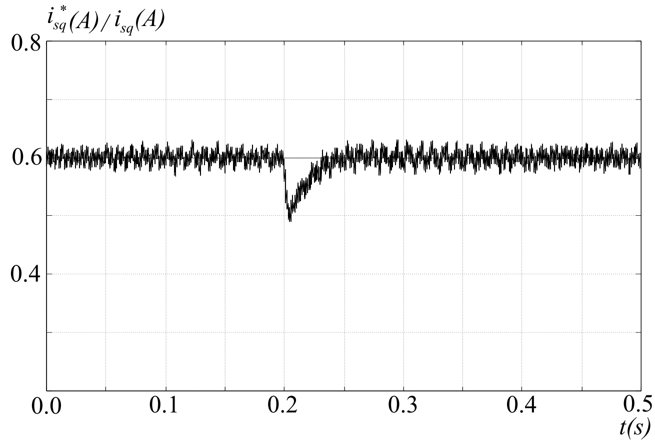

Fig. 3. Simulation results of i_{sq}/i^*_{sq} under an increase of machine stator resistance r_s.

The performance of the proposed $VS - APPC$ control strategy for $PMSM$ motor drive system was evaluated initially by simulation tests. To realize these tests it was implemented a simulation program written in C^{++} language which implements a full $PMSM$ motor drive system. This system is composed by $PMSM$, power inverter, current measurement and $VS - APPC$ control scheme. The $PMSM$ data are listed in Table 1. The full system was simulated with a sampling time of $100\mu s$.

Figures 3-4 present the simulation results for the proposed $PMSM$ motor drive system. In these graphs, the robustness of the proposed control strategy and the identification of the $PMSM$ current-voltage parameters were evaluated. In the first graph it is presented the simulation results of machine phase current i_{sq}^{r} superimposed by its respective reference. In this test, the reference current is settled in $i_{sq}^{r*} = 0.6A$ and at time $t = 0.2s$, the stator resistance r_s is increased from $r_s = 6.187\Omega$ to $r_s = 10.0\Omega$. After this parameter variation, there is some error in the initial transient response which is rapidly reduced as the controlled adapts, and converges to its reference current ($\Delta t < 0.03s$). In the Figs. 4(a) and 4(b) the capability of identification $PMSM$ parameters $\widehat{a}_{sq}$ and $\widehat{b}_{sq}$ are evaluated. These graphs present the average values of those parameters superimposed by their respective nominal values which are $a_{sq(nom)} \cong 187$ and $b_{sq(nom)} \cong 30.7$. In this tests, it can be verified that the average values of parameters $\widehat{a}_{sq}$ and $\widehat{b}_{sq}$ converge approximately to their nominal values. However, it is important to highlight that the proposed control strategy do not guarantee precise estimative of a_{sq} and b_{sq}. Because these estimatives are generated by the use of switching laws. Simulation results demonstrate the effectiveness of the proposed controller. The performance of the proposed controller was also tested by experimental results. To realize these tests, we use an experimental platform composed by a microcomputer equipped with a specific data acquisition card, a control board and a $PMSM$ servodrive. The command signal of servodrive power converter are generated by a microcomputer with a sampling time of $100\mu s$. The data acquisition card employs Hall effect sensors and A/D converters. Figures 5 and 6 show the experimental results for the same test conditions of the simulation experiments. In the Graph 5 it is presented the experimental results the machine phase current i_{sq}^{r} superimposed by its respective reference.

In the beginning, this machine phase current is settled in $i_{sq}^{r*} = 0.6A$. At the time $t = 0.24s$ a current step of $\Delta i_{sq}^{r} = 0.2A$ is applied for evaluating the effectiveness of the control strategy. It can be observed that the machine phase current follows their reference with a good transient response. Graph 6 presents the experimental results fof the machine phase current i_{sd}^{r}. As stated before, this reference current is settled in $i_{sd}^{r*} = 0$. It is important to observe that after time $t = 0.07s$, there is an increase on the current ripple. This is due to the current step imposed on the q machine phase current. It is important to highlight that the time when the current step is imposed do not appear at the same instant at dq machine phase current results, because the physical limitation of our laboratory setup.

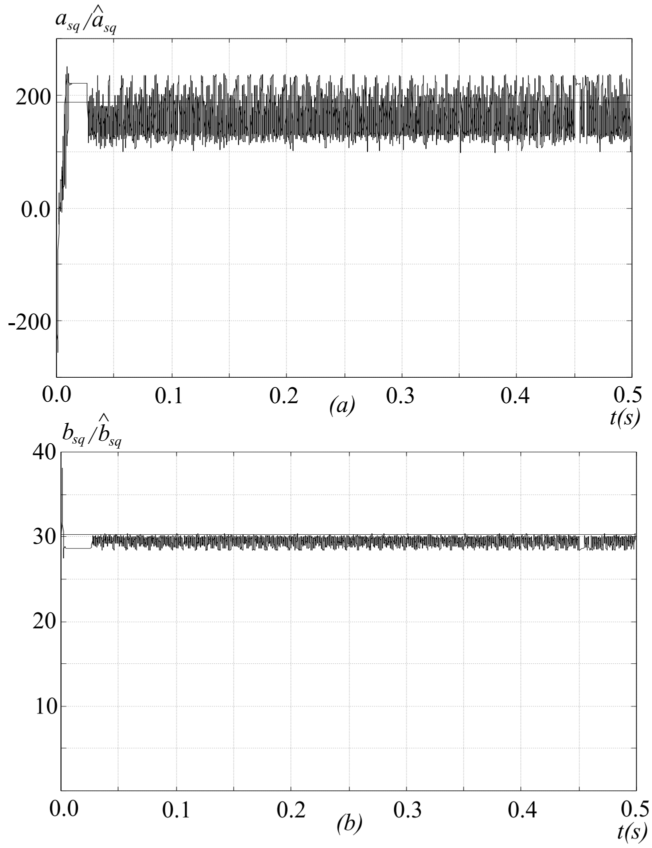

Fig. 4. Simulation results of average values of the estimative of parameters a_{sq} and b_{sq}.

V. Conclusions

This paper has presented a non standard current control strategy for $PMSM$ motor drive system. It is based on a hybrid scheme composed by integration of a variable control structure (SMC) and an adaptive pole placement control ($APPC$). A modification on the controller algorithm was employed for introducing the robustness of unmodeled disturbances on standard $APPC$. With this proposed control scheme, the system is capable to generate the control signal for achieving the desired behavior. A methodology for designing of the parameters of the $PMSM$ proposed current controllers are presented. The capability of the $VS - APPC$ to estimate the machine current-voltage parameters has been also evaluated. The

TABLE 1. $PMSM$ nominal parameters

$Pm = 1.13kW$	$N = 6000rpm$	$p = 4$	$r_s = 6.187\Omega$
$l_{sd} = 24mH$	$l_{sq} = 33mH$	$\Phi_f = 0.1087Wb$	$J = 0.0000084kg.m^2$

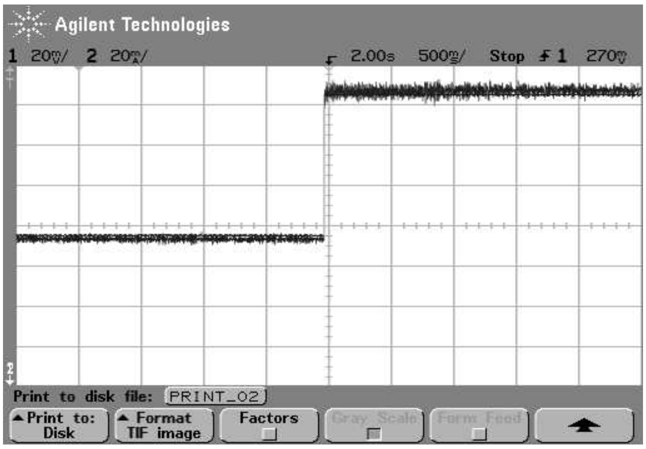

Fig. 5. Experimental results of $PMSM$ phase current i_{sq}^r for a positive step on reference current i_{sq}^{r*}.

simulation and experimental results demonstrate that the proposed $VS-APPC$ has presented a good performance.

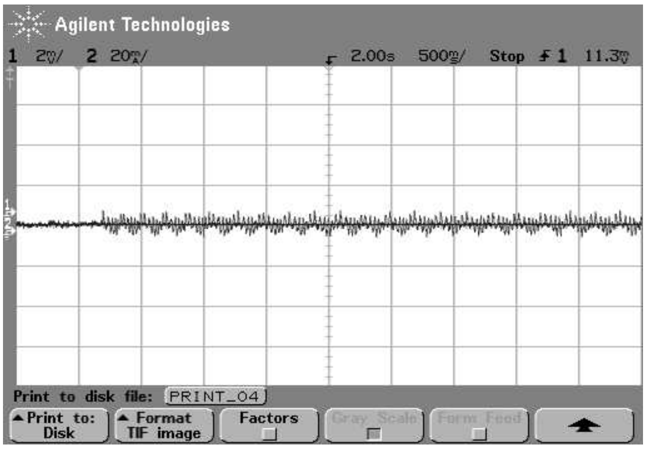

Fig. 6. Experimental results of $PMSM$ phase current i_{sd}^r superimposed by reference current i_{sd}^{r*}.

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