A 2DOF PID for Dynamic Control of Drive System Switched Reluctance Motor

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***Abstract*—Traditional Proportional, Integral and Derivative controllers (PID) have been largely employed for the control of industrial variable speed drives due to easy design and good performance they provide. Switched Reluctance Machines (SRM) due to its inherent characteristics are naturally tolerant to phase faults despite performance loss. However, in some cases, such as electric vehicles and power generation, a non-interrupt operation is mandatory or at least required, even under faulty scenarios. Most of the techniques used to improve the performance of SRMs in fault situations are related to the switching feed converter. This paper proposes a control approach based an a 2DOF PID controller which significantly reduces the phase faults effects on the speed of the motor. Furthermore, the high-frequency noise is attenuated when compared to the classical PID controller commonly applied to control such sort of motors. The proposed controller is able to recover the speed of operation faster than a classical approach when a feedforward action is taken into account.**

***Index Terms*—PID 2DOF controller, Conventional PID, SRM, Fault tolerance.**

I. INTRODUCTION

The inherent characteristics of the Switched Reluctance Machines (SRM), such as its simple structure, low-cost operation and wide speed range make it an invaluable choice among industrial machinery. Applications as pumps, vacuum cleaners, servomechanisms and automotive have emerged as a promising machine due to its high power density, operating in four quadrants, and extensive range of operating speed at constant power [1], [2]. So that, speed control plays an important role in many of these applications, especially to electric vehicles (EV), with works already reported in [3], [4], [5].

Many works related to speed control using SRM are presented in [6], [4], [3]. Model Predictive Controllers (MPC) have been considered, as in [4] where a robust control scheme based on Generalized Predictive Controller (GPC) is shown to exhibit low computational cost. Therein, dynamic performance is validated through experimental results, where the controller is compared with a classical root locus PID. Continuous operation of the SRM is an important requirement because it increases both the safety and reliability of the motor. However, SRM is highlighted due to its inherent tolerance to this kind of issues because of its intrinsic electrically isolated phases [7], [8]. Nevertheless, fault tolerant control schemes are always welcome and eventually enhances overall system performance. The fault tolerance does not necessarily imply free of faults, so that, strategies of analysis play an important role, as well as fault diagnosis strategies. Some works also consider new topologies of the machine itself and the switching drive converter design as cited in [9]. The fault tolerance can be seen as the system's ability to maintain its operation when subjected to different types of faults that occur within the system.

The proposed approach handle with the fault situation at the control level, i.e., this paper proposes a control structure that compensates the fault effects that may occur in either the phases of the machine itself or any of the power switch of a specific phase in the converter. The speed control loop is realized using a 2DOF PID controller. The tuning of the loop is based on a standard response of the speed control system. The behavior of the SRM operating in its motor mode is investigated under open circuit phase fault situation. The control strategy which ensures the machine's operation, if one or two phase faults happen, with low output disturbance and noise attenuation of the control signal in high frequencies. Therefore, the goal is to maintain speed operation under a phase fault condition in order to obtain a fast recovery of the desired motor's speed.

The 2DOF PID control strategy is widely applied in classical and modern control. It has been applied successfully to the regulation of disturbance rejections; it also remains at a given setpoint and the follow-up of the controls whose controlled variable is good in monitoring the desired value [10], [11]. In [12], such a 2DOF PID approach is applied for speed control of DC Motor. A 2DOF PID controller is capable of rejecting

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disturbances without increasing the overshoot to the setpoint level. These controllers are also useful for attenuating the influence of variations of the reference signal on the control signal. Therefore, the structure of the 2DOF regulator is expected to try to meet the objectives regulation and tracking properties. This 2DOF adding is intended to provide greater flexibility in the design of the control system [13], [14]. So, in this paper, such a 2DOF PID control strategy is used for controlling the velocity of the electric phase of a SRM. The proposed controller has two adjustable parameters, one only related to tracking performance, and another only concerned with disturbance rejection property, which simplifies the parameter setting process.

This paper is organized as follows. In Section II the model of switched reluctance motor drive is discussed and the SRM speed loop identification is presented. In Section III, the control design strategies are developed with the Classical PID and 2DOF PID. Section IV depicts the simulation and experimental results of the designed SRM control. Finally, Section V gives the conclusion of this paper.

II. SRM SPEED CONTROL SYSTEM

A three-phase, 6 stator poles and 4 rotor poles SRM, 120 V_{dc}, 10 A, whose series resistance is $r = 0.45\ \Omega$ is the base plant. A block diagram of the drive system is shown in Figure 1.

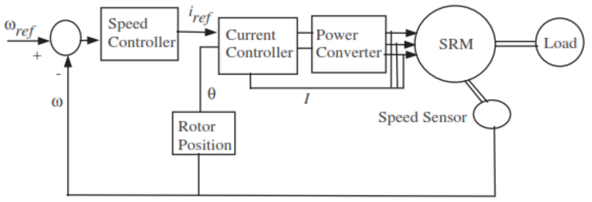

Fig. 1. Block diagram of drive system.

The mathematical model of the SRM is a set of differential equations which are obtained by using standard electric machine theory [15]. These differential equations are:

$$\frac{d\psi_j}{dt} = v_j - ri_j, \quad j = 1, 2, 3 \tag{1}$$

$$\frac{d\theta_r}{dt} = \omega \tag{2}$$

$$\frac{d\omega}{dt} = \frac{1}{J}\left(T - T_l\right) \tag{3}$$

$$T_j = \frac{1}{2} i_j^2 \frac{dL_j}{d\theta} \tag{4}$$

where v_j, i_j and ψ are voltage, current and the flux linkage in the j_{th} phase of the stator, respectively, r is the stator winding resistance, ω is the rotor angular speed, T, is motor torque, T_l, is the load torque, J is the total rotor and load inertia, L_j is the phase inductance and θ is the rotor angular position.

A. SRM Speed Loop Identification

A PID speed controller has been chosen. The controller has an inner current control loop, C_i, and an outer speed control loop, C_v, represented by the block diagram in Figure 2. The speed controller generates a reference current based on the error between the reference speed and actual speed. The current in the designated phase is regulated at the reference level by the current controller.

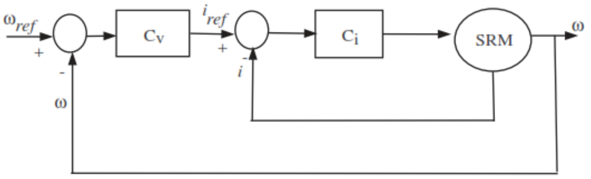

Fig. 2. Overall block diagram of SRM controller.

To acquire a model for the speed loop is necessary to have an inner current controller which ensures reference tracking. There is a lot of research about current controllers. A practical identification based on the setpoint relay method has been presented in [16] where the model transfer function can be represented in a general form as:

$$G(s) = \frac{K_g}{\tau s + 1} \tag{5}$$

where K_g is the gain of the plant and τ is the time constant.

Speed model of SRMs closely relates electromagnetic and mechanical variables. In [6] has presented an identification procedure in which it is suitable to consider the first order model. As a result of this identification procedure, the transfer function of SRM speed loop for discrete-time model considering the sampling period $T_s = 0.01\ s$ and a zero-order holder is given by:

$$G(z) = \frac{0.03259z^{-1}}{1 - 0.996z^{-1}} \approx \frac{0.03259z^{-1}}{1 - z^{-1}} \tag{6}$$

III. CONTROL DESIGN STRATEGIES

A. Classical PID control system

For the SRM speed controller design, a classical PID control strategy known for its simplicity of design and implementation was selected. The control law of a PID controller is given by [17]:

$$u\left(t\right) = K_c * e\left(t\right) + K_i * \int_0^t e\left(t\right)dt + K_d \frac{de\left(t\right)}{dt}, \tag{7}$$

where K_c is the proportional, K_i integral and K_d derivative gain, respectively. In discrete-time, the control law is given by:

$$u(k) = K_c * e(k) + K_i * Ts * \sum_{i=0}^{k} e(i) + \frac{K_d}{Ts} * (e(k) - e(k-1)) \tag{8}$$

$$u(k) = K_c * e(k) + K_i' * \sum_{i=0}^{k} e(i) + K_d' * (e(k) - e(k-1)), \tag{9}$$

where K_c is the proportional gain, T_s is the sampling time, $K_i^{'} = K_i * T_s$ is the integral gain and $K_d^{'}$ derivative gain in discrete-time. Following the same relations, there is still $K_d^{'} = \frac{K_d}{T_s}$. Using the z^{-1} operator as a delay to rewrite the control law as its Z-transform is obtained for the PID controller:

$$u\left(z^{-1}\right) = K_c^{\prime} + \frac{K_i^{\prime}}{\left(1-z^{-1}\right)} + \frac{K_d^{\prime} * \left(1-z^{-1}\right) * (1-p)}{\left(1-p * z^{-1}\right)} \tag{10}$$

The term $\frac{(1-p)}{(1-p*z^{-1})}$ is added as a filter to prevent derivative gain from amplifying high frequency noise.

B. 2DOF PID Control

The ideal PID controller can cause unwanted control actions due to proportional and derivative terms. When there is a sudden change in the reference and the error has large variations in a short time, the actions of these terms can lead to a peak in the control signal, which can lead to saturation. To avoid this type of control signal behavior other strategies involving PI and PID, such as 2DOF are used. A general form of the 2DOF PID controller is shown in Figure 3, where the controller consists of two compensators $C_1(s)$ and $C_2(s)$, and the transfer function $G_d(s)$ from the disturbance d to the controlled variable ω is assumed to be different from the transfer function $G(s)$ from the manipulated variable i_{ref} to ω. This type of strategy uses a feedforward to reduce the effects of proportional and/or derivative action on the reference change without impairing system regulation.

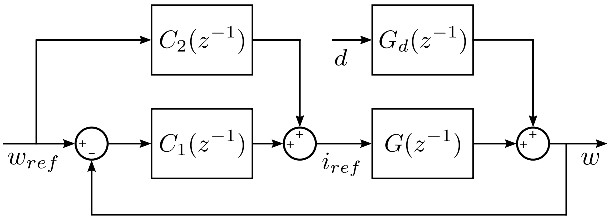

Fig. 3. Feedforward type expression of the 2DOF PID control systems.

The controller $C_1(z^{-1})$ can be an ideal PI or PID while $C_2(z^{-1})$ has the following characteristic, in the case where $C_1(z^{-1})$ is a PID,

$$C_2\left(z^{-1}\right) = -\left(\alpha * K_c^{\prime} + \beta * K_d^{\prime} * \frac{\left(1-z^{-1}\right)(1-p)}{\left(1-p * z^{-1}\right)}\right) \tag{11}$$

The closed loop transfer function of this control system from the setpoint variable ω_{ref} to the controlled variable ω and that from the disturbance d to ω are, respectively, given by

$$\frac{\omega(z^{-1})}{\omega_{ref}(z^{-1})} = \frac{(C_1(z^{-1}) + C_2(z^{-1})) * G(z^{-1})}{1 + C_1(z^{-1}) * G(z^{-1})} \tag{12}$$

$$\frac{\omega(z^{-1})}{d(z^{-1})} = \frac{Gd(z^{-1})}{1 + C_1(z^{-1}) * G(z^{-1})} \tag{13}$$

For the SRM the perturbation considered can be either a load torque or one or two phase faults. According to the closed-loop equations the regulation and the reference change can be treated separately through the coefficients of the controller and the parameters α and β.

C. Controller Design

For the controller design, root locus analysis was used in the discrete domain and the gains $K_c^{'}$, $K_i^{'}$ and $K_d^{'}$ selected so that the poles of the closed-loop are positioned next to $z = 0.93$. Thus there is a rapid response to disturbances (good regulation) for the PID controller parameters were selected to meet this criterion in the closed loop.

The plant used to perform the calculations was obtained using the least squares method and its transfer function is given by Eq. (6). Then the PID controller gains are listed below where the filter pole has been set to $z = 4.5 * 10^{-5}$, while the other closed-loop poles close to $z = 0.93$, as well as in the PI controller, that is, $K_c^{'} = 4.497$; $K_i^{'} = 0.165$; $K_d^{'} = 1.553$; $p = 4.5 * 10^{-5}$

IV. SIMULATED AND EXPERIMENTAL RESULTS

In this section, we present the simulation and experimental results for the output speed of the SRM and reference value of the phase current considering each case of α and β.

The conventional PID controller can be obtained from the 2DOF PID controllers as special cases by choosing 2DOF parameters appropriately. Namely, the conventional PID controller is obtained by setting $\alpha = \beta = 0$, the preceded-derivative PID by setting $\alpha = 0$ and $\beta = 1$, and the I-PD by setting $\alpha = \beta = 1$.

The simulated responses to different values of α and β are shown in Figures 5 to 8. Initially, the reference is 1000 rpm, after 3 seconds of the simulation the reference is changed to 1100 rpm and at 5 seconds a two-phase fault (disturbance) occurs.

Moreover, the maximum load in a fault situation has a direct relation with the machine speed. So, the experimental results shown were carried out with a relation between speed reference and load torque that couldn't cause current saturation. Then, for the speed reference of 1100 rpm, the load torque was 0.1 Nm.

To explore the effectiveness of proposed controller experimental tests have been carried out that are shown in Figures 5 to 8. The proposed 2DOF PID controller was compared with a conventional PID controller. The controllers have been implemented in a prototype using a digital signal processor $TMS320F28335$ (DSP) and a classical asymmetric bridge converter with switching frequency of 25kHz. The experimental setup is presented in Figure 4.

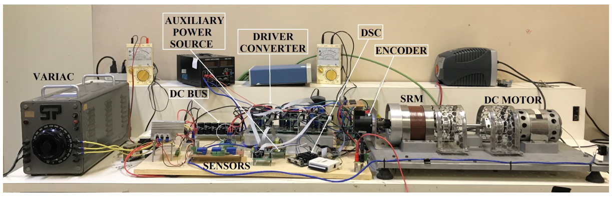

Fig. 4. The experimental setup of the SRM drive.

For values of α and β equal to zero, Figures 5, the equivalent controller is a conventional PID since there is no

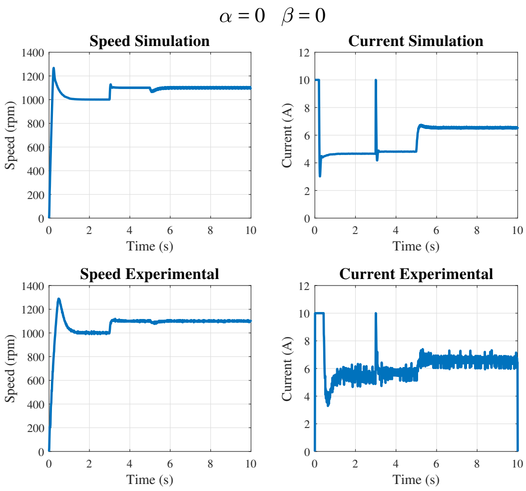

Fig. 5. The simulation and experimental setup of the SRM drive - PID classical.

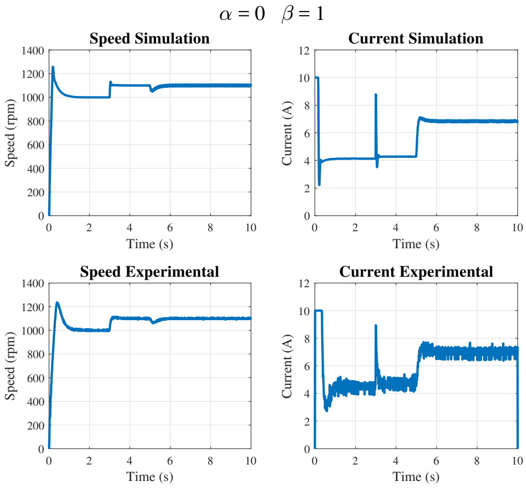

Fig. 6. The simulation and experimental setup of the SRM drive - preceded-derivative PID.

feedforward block action. Thus, even a small change in the reference leads to a peak in the control signal, caused by the aggressiveness with which the controller was designed. By putting the compensating block into action, we notice a reduction in the peak caused by the reference change, as well as a smoother transition between the set points. By making $\alpha = 0$ and $\beta = 1$, Figures 6, the derivative action of the controller is suppressed from the reference change, and abrupt changes in the control signal are due to only to the proportional part. In the same way that doing $\alpha = 1$ and $\beta = 1$, Figures 7, removes the action of the proportional and derivative parts of the reference change, making the controller behavior equal to that of an I + PD controller [11]. In Figure 8 correspond to the action of PI controller considering $\beta = 0$ and $\alpha = 1$

Three performance indexes are proposed for the comparison of experimental results. The largest overshoot (OS) due to a

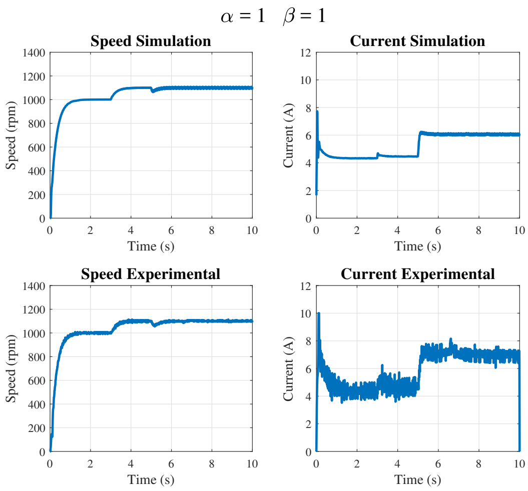

Fig. 7. The simulation and experimental setup of the SRM drive - I-PD controller.

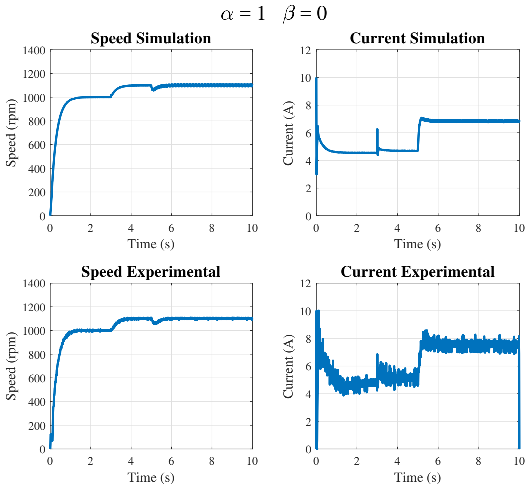

Fig. 8. The simulation and experimental setup of the SRM drive - PI controller.

reference change, the rise time (t_r) for 1000 rpm, and the undershoot (US) caused by the disturbance.

According to Tables I and II, the addition of the feedforward compensator block did not degrade perturbation rejection since changes in α and β values do not cause a considerable influence on the undershoot caused by the disturbance and brought flexibility in the choice of control signal aggressiveness at the reference change. This enables controllers with good control performance to be designed and allow smoother response in variations in the reference, helping to also prevent overlaps in a machine with a non-linear characteristic. Another benefit in adding the compensator block is to prevent very sudden variations in the control signal, thus avoiding situations of saturation of the same.

Therefore, as shown in [14], for the SRM used in this experiment good coefficient were obtained by setting $\alpha = 1$

and $\beta = 1$. In this way, the 2DOF PID controller significantly reduces the phase fault effects on the speed of the motor and maintains acceptable speed response characteristics, such as overshoot in relation the reference, settling time for 1000 rpm and 1100 rpm, maximum peak and undershoot, as presented by the performance indexes shown in Tables I and II.

TABLE I
SIMULATED PERFORMANCE INDICES.

	$\alpha = 0$ $\beta = 0$	$\alpha = 0$ $\beta = 1$	$\alpha = 1$ $\beta = 0$	$\alpha = 1$ $\beta = 1$
(OS)	26.69%	25.87%	0.0447%	0.0625%
(t_r)	$0.13s$	$0.11s$	$0.57s$	$0.58s$
(US)	3.24%	4.75%	3.99%	3.34%

TABLE II
EXPERIMENTAL PERFORMANCE INDICES.

	$\alpha = 0$ $\beta = 0$	$\alpha = 0$ $\beta = 1$	$\alpha = 1$ $\beta = 0$	$\alpha = 1$ $\beta = 1$
(OS)	28.85%	23.62%	0.908%	0.805%
(t_r)	$0.25s$	$0.21s$	$0.55s$	$0.56s$
(US)	1.982%	3.576%	4.113%	4.145%

V. CONCLUSION

Control strategies of the SRM have been widely analyzed in several studies, but the SRM behavior analysis in phase loss conditions are still incipient. This work has proposed a control approach based on a 2DOF PID controller specifically driven to the speed control of SRM in order to ensure the functionality of the bench robustly and efficiently. The performance of the SRM changes as a fault situation occurs reducing the gain in the direct loop. The use of a feedforward action along with a trade-off between dynamical response and robustness achieved by the proposed 2DOF PID has shown its feasibility by means of simulation and experimental results presented. In addition, because of its feedforward action, the proposed 2DOF PID controller is able to recover the speed of operation faster than a classical approach, when a feedforward action is not taken into account.

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