A NEW SIMPLIFIED CURRENT CONTROL FOR SWITCHED RELUCTANCE MOTOR DRIVES

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Abstract **- A new simplified current control method for Switched reluctance motor (SRM) based on Proportional Integral (PI) auto-tuning controller has been developed. The method uses a set-point relay framework to extract relevant information about the process dynamic. During the on-line model of SRM winding current, this novel approach keeps the PI controller in a closed loop to improve the process safeness and model accurateness. Furthermore, the modified Ziegler-Nichols method is adopted to calculate the PI parameters. Several laboratory experimental tests has been performed to evaluate the method using a 12/8, 3-phase SRM.**

I. Introduction

Switched reluctance motors (SRMs) are potentially attractive for many industrial applications for their simple structure, high speed, and low cost advantages. However, as SRM utilization in industry is still relatively limited compared to other machine types, the practical drive implementation issues have not been fully explored. In particular, the inner control loop could benefit from additional research. The goal of this paper is to focus on the development of a fully digital implementation of the inner current control loop for high-performance SRM drives operating with high saturation. However, the nonlinear characteristics in the operation of an SRM complicate the analysis and the control of the motor, and limit its applicability to high performance applications, such as servo systems.

One of the primary disadvantages of SRMs is the torque ripple. The torque is often controlled via an inner current loop, with a nonlinear relationship between torque and current. To achieve good torque linearity and low torque ripple, accurate command tracking is required from the current controller. At low speed, current profiling may be utilized to reduce torque ripple. Several researchers have proposed current profiling based methods to minimize torque ripple [1], [2], [3] and [4]. The most common strategies of current controllers can be classified as hysteresis and the linear controllers. Each scheme has its advantages and drawbacks with regard to insensitivity to parameters variations, accuracy, robustness, and dynamic response over the entire speed range. The advantages of hysteresis current controllers lie in their simplicity and their providing fast responses and good accuracy, because they act quickly. However, there are well known disadvantages such as variable-frequency switching, and high ripple current, making it undesirable for many applications. This is mainly due to a lack of sliding mode trajectory in hysterisis controller. A PI controller with switching frequency constant has advantages such as easy digital implementation, low ripple current, and simple structure. However, due to the aforementioned nonlinear plant characteristics, good performance and stable operation are difficult to achieve over the full operating range. The main difficulties in both cases are the electrical time constant (L/R) and dynamic disturbance (back EMF) which vary widely with current and rotor position causes poor performance of the controller [4].

AC small signal modeling technique has been widely used in power converters for many years [5]. This paper presents a new practical and simple approach for modeling and controlling of a SRM current loop and a design methodology for PI current controllers was developed based on setpoint relay method [6]. By means of the relay control, small oscillations on the output current are generated. Based on the measurements of the frequency and amplitude of the oscillations, two parameters of a first-order-plus-deadtime model at the oscillation frequencies are derived. The control design is done without on-line iteration and the PI parameters of the current loop are determined analytically and instantaneously using the modified Ziegler and Nichols method [6], [7] and [8]. The advantage of this method is a low demand processing time. Experimental results are included to verify the performance of the proposed controller.

The new simplified mathematical model for the SRM current loop is a suitable tool for control design. The setpoint relay method provides good information about a operation point to determine the parameters of a first-order-plus-deadtime model.

II. On-line SRM Modelling Scheme - Process Identification

A. Describing Function Method

Describing function analysis is a method that can be used to predict and approximately analyze nonlinear behavior. The basic idea is to approximate a nonlinearity by a linear equivalent and then use different frequency domain techniques to analyze the resulting system [9].

Since a relay is a nonlinearity, this analysis is used. The use of ideal on-off relays limits the identification of the frequency response to only the phase crossover point. This restriction is due to the describing function of the ideal relay being a real valued function. The basic principle of relay feedback method can be explained following Fig. 1. The SRM current transfer function, which is going to be

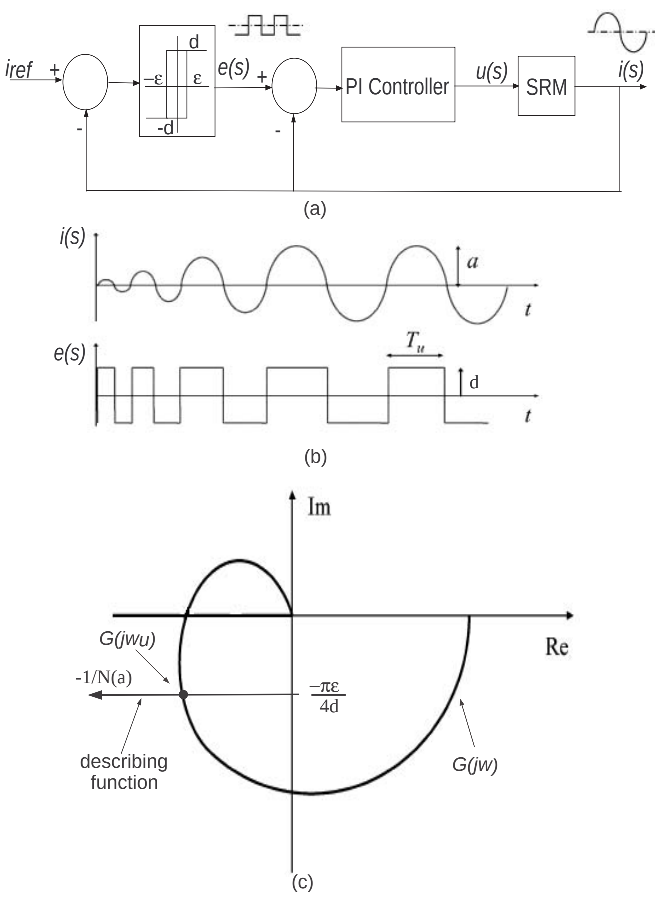

Fig. 1 (a) Block diagram of Process under PI control with setpoint relay. (b) Typical waveforms of relay feedback. (c) Nyquist diagram of $G_p(s)$ and relay block.

controlled, denoted with $G_p(s)$, is regulated firstly using a relay see Fig. 1(a). Thus, an oscillation on signal with period T_u (or angular frequency ω_u) and amplitude a is generated see Fig. 1(b). The oscillation waveform is almost sinusoidal assuming that the low-pass filter action of $G_p(s)$ filters the higher harmonics generated by the relay. The condition for the oscillation to be maintained is the following:

$$N(a).G(j\omega_u) = -1 \tag{1}$$

where $N(a)$ is the relay transfer function modeled using the describing function method. Practical implementation of the scheme of Fig. 1 usually requires the introduction of a hysteresis function in the relay block. This is usually needed in order to avoid multiple zero-crossing due to the noise in the sensed variable. When the hysteresis function is used, the describing function of the relay is:

$$N\left(a\right) = \frac{4d}{\pi a} e^{-j\sin^{-1}(\varepsilon/a)} \tag{2}$$

where d is the amplitude of the relay, ε is the width of the hysteresis and a is the amplitude of the resulting oscillations at the output of the plant under the control of the hysteresis relay. Condition (1) is also represented in the Nyquist diagram of Fig. 1(c). Function $1/N(a)$ can be represented as a straight line parallel to the real axis, in the complex plane.

In the configuration of Fig. 1, for a hysteresis relay of d and ε values, under limit cycling conditions with frequency ω_u and amplitude a, the operating point approximately satisfies

$$\frac{4d}{\pi a} e^{-j\sin^{-1}(\varepsilon/a)} G\left(j\omega_u\right) = -1 \tag{3}$$

where $G(j\omega_u)$ is the frequency response of the linear process. Hence,

$$\left|G\left(j\omega_u\right)\right| = \frac{\pi a}{4d}, \angle G\left(j\omega_u\right) = -\pi + \sin^{-1}\left(\varepsilon/a\right) \tag{4}$$

Eq. 4 shows that for a given set of d and ε, the gain and phase of $G(j\omega)$ can be determined by observing a and ω_u from the resulting limit cycles.

B. Process Model

This work used a two-quadrant-chopping converter. Based in ac small signal modeling technique used in power converters, and assuming the duty-cycle is d, with the average voltage applied to the phasing winding during one switching period, the output current to duty-cycle transfer function can be derived [5]:

$$G_{id}(s) = \frac{T_{uo}}{1 + s/\omega_o} \tag{5}$$

With

$$T_{uo} = \frac{2V_{dc}}{R_a + \omega K}$$

and

$$\omega_o = \frac{R_a + \omega K}{L}$$

where ω is the speed; K is the gradient of the inductance varying with the rotor position; L is the average inductance over one switching period; V_{dc} is the power supply. Because the L, K varies with the rotor position, the output current to duty-cycle transfer function contains a moving single pole.

The phase current can be measured on-line using the setpoint relay experiments. The control scheme comprises of a relay and an extra feedback signal to the setpoint, as shown in Fig. 1. The basic idea of the setpoint relay is similar to that of relay feedback experiment [8] except that it have the advantage of oscillate inside a "safe operation zone" due the performance of the PI controller in forward loop and the existing control system is always in closed-loop during the tuning. The control parameters are determined from the knowledge of the critical gain

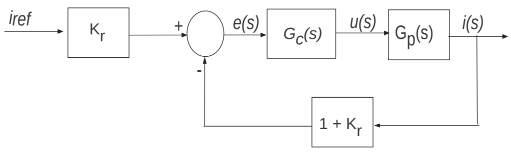

Fig. 2 Diagram of the setpoint relay.

and the critical period. The relay output is connected to the setpoint of the existing PI controller.

Assume that the gain of the relay is K_r, which corresponds to its describing function, the transfer function of the controller is $G_c(s)$ and the transfer function of process is $G_p(s)$, the error $e(s)$ can be calculated as:

$$e(s) = K_r i_{ref} - (1 + K_r)\, v(s). \tag{6}$$

where i_{ref} is the setpoint; $u(s)$ is the controlled variable or process variable. Hence the block diagram can be rearranged as shown in Fig. 2. The loop transfer function is $(1+K_r)G_p(s)G_c(s)$. The equivalent gain or describing function for the relay element is [10]:

$$K_r = \frac{4d}{\pi\sqrt{a^2 - \varepsilon^2}}$$

C. Setpoint relay under PI control

The relay based identification method can give only one critical point on the Nyquist curve of the open loop transfer function, 1(c), which is able to determine two unknown parameters in the process model. Therefore, we adopt the transfer function between the duty-cycle and the output voltage, (5), plus deadtime model, given by [7]:

$$G_p(s) = \frac{K_p e^{-\theta s}}{\tau s + 1}. \tag{7}$$

where K_p is process gain, τ is the time constant and θ is the dead time.

It is noteworthy that model (5) is an exact expression for SRM converter proposed by [5], it was also used to interpret the relay autotuning method by [7]. This transfer function will be applied to the drive of the SRM, shown in Fig. 1. Considering a PI controller with the following transfer function as:

$$G_c(s) = K_c\left(1 + \frac{1}{T_i s}\right). \tag{8}$$

where T_i is the integral time. Then, the system is forced to oscillate at the resonant frequency ω_u. When sustained oscillation is developed:

$$\arg\left[(1 + K_r)G_p(jw_u)G_c(jw_u)\right] = -\pi. \tag{9}$$

and

$$|(1 + K_r)G_p(jw_u)G_c(jw_u)| = 1. \tag{10}$$

where w_u, is the oscillation frequency. From 10 we obtain:

$$\tau = \frac{(1 + K_r)K_p\sqrt{T_i^2\omega_u^2 + 1}}{T_i\omega_u^2}). \tag{11}$$

In the same way, we obtain the estimate of deadtime in model of 7, from 9 we as:

$$\theta = \frac{\left[\frac{\pi}{2} + \tan^{-1}(T_i\omega_u) - \tan^{-1}(T\omega_u)\right]}{\omega_u}. \tag{12}$$

Defining

$$K_u = \frac{2\pi\tau}{T_u} \tag{13}$$

and

$$T_u = \frac{2\pi}{\omega_u} \tag{14}$$

it is easily seen that with the above assumptions K_u is the gain that brings the system to stability boundary. The ultimate gain K_u and the ultimate period T_u are thus easily found from a relay experiment [11].

Experimentally the output variable $i(t)$, oscillation amplitude a and period T_u were measured and the parameters of equation (7) were obtained on the basis of equations (11) and (12). The plant gain can be determined from the steady state or by the other corresponding way.

Eq. 5 show that the parameters of the transfer function of the converter, which varies with the rotor position and rotor speed. During the dwell angle, θ_{dwell}, of each phase, the inductance L varies between minimum value Lu to maximum La, and the inductance gradient K may be either zero or $\frac{L_a - L_u}{\theta_{dwell}}$. Because the original uncompensated system contains a moving pole and the gain of the uncompensated transfer function varies with the rotor position, it is impossible to design a fixed PI compensator to maintain constant gain and phase margin at different rotor position. There are two ways to design a compensator to make system stable. One is to design a fixed compensator at worst case to make system stable under all rotor position. Designing a timing varying compensator with rotor position to maintain stable performance all the time is another way; however, this is hard to realize in the digital signal processing (DSP) program.

The relay method identify two points of transfer function of converter when consider the oscillation of process. Then, the time constant, τ, consider the effect of variation parameters with the rotor position and rotor speed. For the speed change in relay setpoint method, we have the others operations point mainly the value of reference current.

III. Current Controller Design

To verify the proposed method to model of SRM current loop, then based on the estimates of deadtime and time constant, it is possible to use the Ziegler-Nichols reaction curve formula to calculate the PI parameters directly. In this work, it was adopt the modified Ziegler and Nichols method for determining the parameters of the PI controller. The PI controller parameters are given by [8]:

$$K_c = K_u r_b \cos(\phi_b). \tag{15}$$

$$T_i = -\frac{T_u}{2\pi\tan(\phi_b)}. \tag{16}$$

where r_b and ϕ_b are module and phase specified, how design parameters.

IV. Experimental Results

In order to verify the proposed method to model for the SRM current loop, a hardware prototype was implemented with digital signal processor TMS320F2812 when running at $25kHz$ switching frequency. During this test bench setup, a 12/8 SRM with three-phase, $120Vdc$, $2.5A$ is used. It has 12 stator poles and 8 rotor poles. Each phase circuit comprises of single-tooth coils wound on one set of four stator poles, 90 mechanical degrees apart. The inductance profile of each phase is repeated in 45 mechanical degrees. In the whole control process, only one phase winding is conducted at any time and no overlap between phase current. So, the maximum dwell angle of each phase is 15 mechanical degrees. The parameters of PI controller used in this project are designed. The two-quadrant-chopping converter is used for the motor drive. Rotor speed, $220rpm$ and input voltage, $120V_{dc}$ are assumed in relay setpoint identification. The work was undertaken using the experimental setup shown in Fig. 3. The estimated values of current and voltage are updated every $40\mu s$ ($25kHz$).

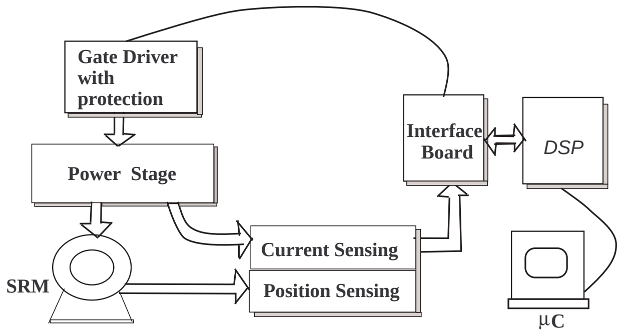

Fig. 3 The setup system.

To verify the proposed method, it was carry out the setpoint relay experiments for SRM drives with the following values: phase current in winding of $2.5A$, amplitude of hysteresis $d = 1.5$, width of hysteresis $\varepsilon = 0.2$. The sustained oscillation for all SRM phases are shown in Fig. 4.

From equations (11) and (12), it was found $\tau = 0.0861s$; $\theta = 0.0195ms$; $T_u = 0.000776rad$; $K_u = 697.22$, and amplitude $a = 0.35$. The parameters of the PI controller, Fig. (1), were established in $K_c = 100.$ and $T_i = 2.22 \times 10^{-4}$. From equation (6):

$$G_p(s) = \frac{0.8e^{-0.000195s}}{0.08611s + 1}. \quad (17)$$

To validate this approach further extensive experimentation was performed and favorable outcomes were obtained. In this way the other experiments with different values for ε with amplitude of hysteresis fix, $d = 1.5$, and same existing PI controller setting were obtained. Several situations are shown in Table 1.

For one of value, $\varepsilon = 0.2$, the estimated parameters are used for auto-tuning of PI controller and was obtained

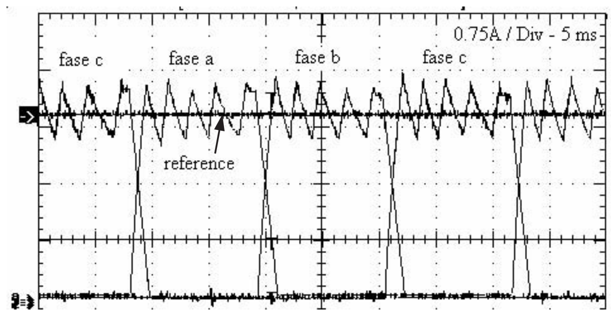

Fig. 4 Output of relay setpoint: phases A,B and C.

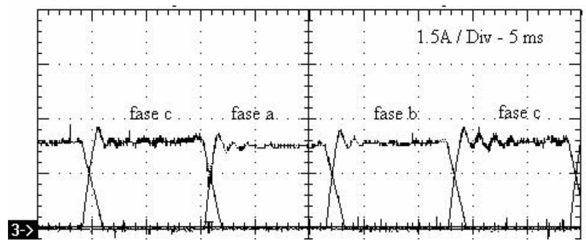

Fig. 5 Currents of SRM phases with PI controller.

by modified Ziegler-Nichols method with $r_b = 0.29$ and $\phi_b = 46^o$, that obtained gain proportional $K_c = 140.45$, integral time $T_i = 0.199ms$, where the operation frequency is given by $\omega_u = 588.235rad/s$. Fig. 5 shows a basic implementation take into account a load in the SRM shaft (implemented through a dc motor).

In order to access the usefulness of the proposed method experimental tests were repeated for different setpoint of current ($2.5Ato3.5A$) and rotor speed ($220rpm$ and $300rpm$), respectively. Figures 6 and 7 show the waveforms of current with PI controller.

By using an iterative procedure, adaptive control can be considered that presumes an identification method. The tuning of PI parameters is obtained directly by including the controller in the relay feedback and by adjusting the controller parameters based on the specified performance.

The application of setpoint relay is done in relative low speed for the SRM mainly due to two reasons: the first is because the current control is an issue only in low speed, where current normally is limited by chopping. The use of single pulse control is typically used at and above nominal speed. The second reasons is because with the increases of speed sustained oscillation can not be developed then

Table 1 Identification Results from Setpoint Relay for SRM

ε	$T_u[rad]$	K_u	$\tau[ms]$	$\theta[ms]$
0.15	0.00072	5064.23	0.0576	0.19
0.1	0.0076	637.29	0.077	0.19
0.2	0.0077	697.22	0.0861	0.195

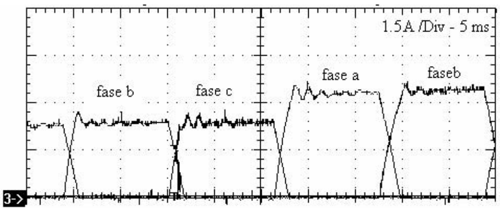

Fig. 6 Currents of SRM phases with PI controller and setpoint change, $\omega = 220rpm$.

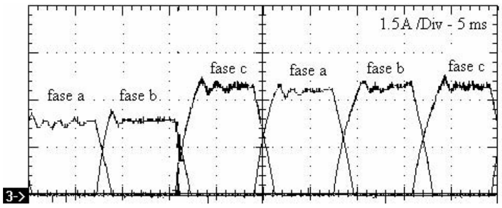

Fig. 7 Current waveforms of SRM phases with PI controller and setpoint change, $\omega = 300rpm$.

the relay setpoint method is not suitable.

V. Conclusions

The main contribution of this paper is the use of relay feedback for identifying the SRM loop current transfer function. The identification is accomplished in frequency domain and the results are used for designing a PI control loop based on gain and phase specifications. Relevant information are extracted from limits cycles established by relay set-point scheme and the describing function method is used for modeling purpose. The proposed SRM modeling scheme does not require a priori knowledge of motor characteristics. To verify the proposed method the modified Ziegler-Nichols PI tuning was used to calculate the controller parameters. Experimental results are presented and demon- strate the feasibility of the SRM modeling and controlling scheme.

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