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MODEL REFERENCE ADAPTIVE CONTROL OF RIPPLE REDUCED SRM DRIVES

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Abstract

The paper proposes a model reference adaptive control method for SRM drives. The main goal of the drive control is to improve dynamical performance by compensating for the motor nonlinearities. The proposed ripple reduced method changes only the turn-on and the turn-off angle in function of the speed and current reference. One of the advantages of using this method is that it does not need the real-time calculation or measuring the motor torque. Simulation and experimental results are presented.

Keywords: switched reluctance drives, adaptive control, simulation.

1. Introduction

In motion control systems are robustness against parameter changes and disturbance rejection of main interest. The model reference adaptive control has the following features:

- It makes the compliance of the system with varying operational conditions possible and ensures the behavior of the controlled system according to the prescribed reference model.
- It means such a special type of adaptive systems which results in nonlinear control systems. This is the reason why the analytical analysis is completed by Lyapunov stability criterion or by

  hyper-stability principle.
- Its planning and application are closely related to the using of computer methods.
- Simple realisation of the control algorithm.

In this paper the application of a model reference adaptive control to switched reluctance motors is presented.

2. Drive System

The block scheme of the examined drive system with switched reluctance motor (SRM) is shown in *Fig. 1.* An SRM of 6/8 pole and 4 kW rated power was used.

The supply unit consists of three main blocks, namely the RECTIFIER, the FILTER and the INVERTER. The inverter is a pulsed width modulated (PWM) one, marked with QP in the figure and it contains one-one switching transistor per phase and a brake chopper, not shown in the figure. The common point of phase windings is supplied by the PWM inverter. It is of autonomous operation and has an inner current control loop. The other ends of phase windings are connected to the phase switching transistors.

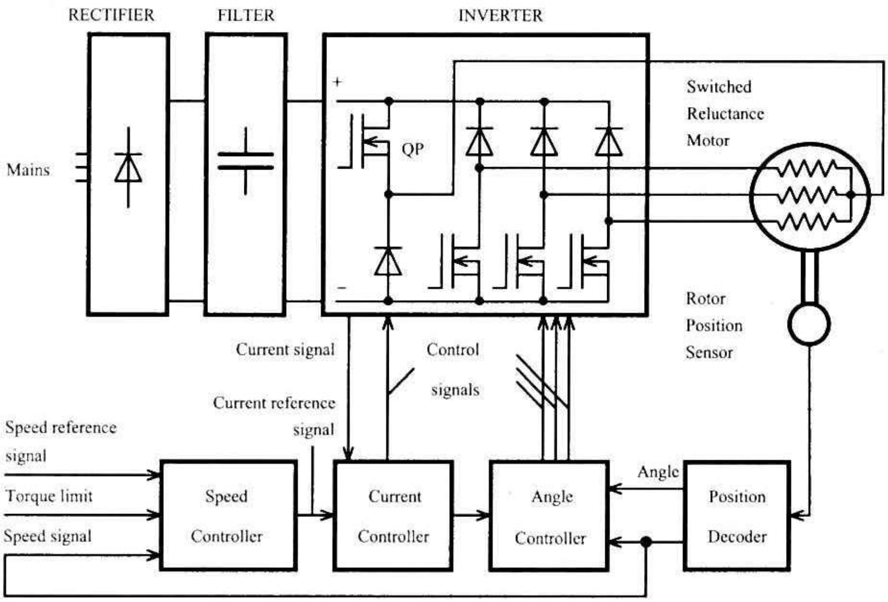

Fig. 1. Block scheme of drive system

It follows from the operational principle of SRM [1] [3] that its phase windings are to be excited at a well determined angle of the rotor position in an appropriate order. This is why a Rotor Position Sensor is to be mounted on the shaft of the motor. In our case the position sensor is a resolver. It can be calculated from the pole numbers that the phase switchings have to follow each other by 15 degree. The resolver is supplied by an oscillator circuit, their signals are evaluated by a Position Decoder.

The Position Decoder has two outputs: the Angle and Speed signals. Based on the two signals, the Angle Controller composes the Control signals for the phase switching transistors.

Fundamentally, SRM drives have two control loops, the outer one is the speed loop, Speed Controller and the inner one is the current loop, Current Controller.

The output signal of the Speed Controller serves for a Current reference signal of the Current Controller. The hardware and software tools together fulfil the two-loop control. The Current Controller produces the control signal for the PWM inverter, and receives the Current signal from the PWM inverter at the same time.

At testing and fault detection there is a possibility to connect an intelligent terminal or a computer via RS 232 series data line to the drives. The fundamental part of the control unit is a single-chip microcontroller. It contains an 256x8 on-chip RAM, four 16 bit timer/counters, a fast 32 bit division unit and a 16 bit multiplication unit, 12 multiplexed input 8-bit A/D converter with programmable reference voltage, two full duplex serial interfaces, and a compare/capture unit. The instruction execution rate of the microcontroller is 1 MIPS.

3. Control of SRM Drives

3.1. Current Control

The Current Controller is totally based on its hardware solution. Based on the current reference signal, it controls the PWM inverter of fix frequency by installing an analog controller. The current feedback also comes from the PWM inverter.

For the control of the sum of phase currents (*Fig. 1*) a simpler four-transistor inverter is suitable and a six-transistor one is not necessary as in the case of control of phase currents independently from each other. But the detriment of the previous solution is that the torque pulsation can be decreased in a smaller degree by changing the turn-on and turn-off angles.

Namely, in the case of the constant current reference signal the current increase is limited by the switched-off, but conducting phase current as the regulator controls the sum of two phase currents. The increase of the phase current can be forced by the modification of the current reference signal [6] at starting of the conducting state:

$$i_r = u\sum_{j=1}^{3} C_j + \sum_{j=1}^{3}(1 - C_j)i_j, \tag{1}$$

where:

i_r is the current reference signal,
i_j is the current signal of phase,
u is the output of the speed controller,
C_j is the control signal of phase (0 or 1).

The supplement of the first member of *Eq.* (1) makes the overlap of the phase conduction possible, while the effect of the second member is to increase the reference signal with the current of the switched-off, but not current-free phase.

3.2. Speed Control

A model reference adaptive control is used for the speed control [4], [7], [8]. Such an adaptive control has been successfully elaborated by using a suitably chosen Lyapunov function to compensate the gain of the speed control loop.

The Lyapunov function has been chosen in such a way that the model error should be decreased asymptotically and the gain of the speed control loop and the load should be compensated. In this case we have had to assume that the change of the loop gain and the load is smaller than the speed of the adaptation.

The arrangement of the speed control loop is shown in *Fig. 2* where an adaptational signal, signed by g is added to the error signal. Applying the signal adaptation control, a P type controller of K_p gain can ensure zero speed error as the adaptation signal g can produce a current reference signal to compensate the loading current at zero speed error. The controlled loop has been approximated by an integral element. The time constant of the closed current control loop has been neglected only in the interest of simpler adaptation algorithm.

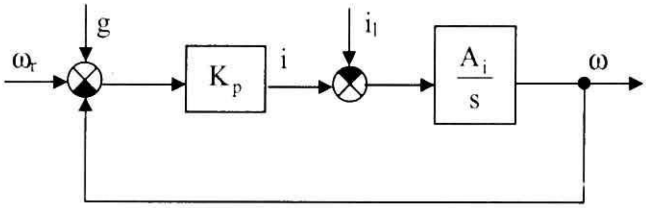

Fig. 2. Block scheme of the speed control loop

The symbols of *Fig. 2* are as follows:

$A_i = k_m / J_m$
k_m is the torque coefficient of the motor,
J_m is the total inertia related to the motor shaft,
i is the current of the motor,
i_l is the current equivalent to the load-torque,
ω is the speed of the motor,
ω_r is the speed reference signal.

The relation between the acceleration current of dynamic torque and the speed can be written by the following transfer function:

$$Y_{\omega(i-i_l)} = \frac{A_i}{s}. \tag{2}$$

Regarding the block diagram of the motor the following differential equation is valid for the closed loop:

$$\dot{\omega} + A_i K_p \omega = A_i K_p(\omega_r + g) - A_i i_l. \tag{3}$$

The feature of the closed speed control loop has been taken into consideration by a parallel control model to be expressed by a one-storage proportional element. The differential equation of this first order system is:

$$\dot{\omega}_m + q_m \omega_m = q_m \omega_r, \tag{4}$$

where the index m refers to the model and q_m is the reciprocal of the model time constant.

Using *Eqs.* (3), (4) and introducing the expression $\varepsilon = \omega_m - \omega$ for the model error, the dynamic equation for the error is as follows:

$$\dot{\varepsilon} + q_m \varepsilon = (q_m - A_i K_p)(\omega_r - \omega) + A_i(i_l - K_p g). \tag{5}$$

The adaptation signal $g(t)$ can be written in the following form:

$$g(t) = g_1(t)(\omega_r - \omega) + g_2(t). \tag{6}$$

Substituting *Eq.* (6) for (5):

$$\dot{\varepsilon} = -q_m \varepsilon + b_1(\omega_r - \omega) + b_2, \tag{7}$$

where:

$$b_1 = q_m - A_i K_p(1 + g_1(t)),$$
$$b_2 = A_i(i_l - K_p g_2(t)).$$

Let us compose the following Lyapunov function to produce the signal $g_1(t)$ and $g_2(t)$:

$$V = \frac{1}{2}\varepsilon^2 + \frac{1}{2}(\beta_1 b_1^2 + \beta_2 b_2^2), \tag{8}$$

(where β_1 and β_2 are positive constants).

The time-derivation of the Lyapunov function is:

$$\dot{V} = \varepsilon\dot{\varepsilon} + \beta_1 b_1 \dot{b}_1 + \beta_2 b_2 \dot{b}_2. \tag{9}$$

Substituting *Eq.* (7) for (8):

$$\dot{V} = -q_m \varepsilon^2 + (\omega_r - \omega) b_1 \varepsilon + b_2 \varepsilon + \beta_1 b_1 \dot{b}_1 + \beta_2 b_2 \dot{b}_2. \tag{10}$$

If

$$\dot{b}_1 = -(\omega_r - \omega)\varepsilon/\beta_1 \tag{11}$$

and

$$\dot{b}_2 = -\varepsilon/\beta_2,$$

then

$$\dot{V} = -q_m \varepsilon^2 \tag{12}$$

and it ensures the asymptotical stability of the model error. On the basis of *Eqs.* (7), (11) and by assuming that the variations of A_i can be neglected compared to the speed of adaptation, the following adaptation algorithm is valid:

$$\begin{aligned} \dot{g}_1(t) &= \gamma_1 \varepsilon(\omega_r - \omega), \\ \dot{g}_2(t) &= \gamma_2 \varepsilon, \end{aligned} \tag{13}$$

where γ_1 and γ_2 are positive constants, the free parameters of the adaptation. Taking the relations (6), (13) into consideration the following equation comes true:

$$g(t) = \gamma_1(\omega_r - \omega) \int \varepsilon(\omega_r - \omega)\, \mathrm{d}t + \gamma_2 \int \varepsilon\, \mathrm{d}t. \tag{14}$$

The block diagram of the control circuit introducing the adaptational signal $g(t)$ furthermore $g_1(t) = \text{const.}$, $A_i = \text{const.}$ can be seen in *Fig. 3* for the nextcoming analysis. The adaptational gain factor γ_2 gives the reciprocal of the integrating time constant of controller type PI, assuming $g_1(t) = 0$.

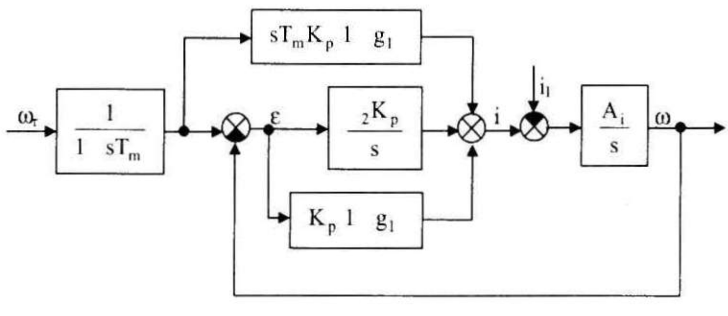

Fig. 3. Block scheme of the adaptive speed control loop

It follows from the determination of the Lyapunov function, that $b_1(t)$ will be equal to zero at the end of the adaptation. Using *Eq.* (7) the following relation is valid:

$$1 + g_1(t) = q_m/(A_i K_p). \tag{15}$$

It can be seen from relation (15) and the block diagram of the control circuit that the task of adaptation signal $g_1(t)$ is to ensure the constant gain in the loop.

For fulfilling the constant integrating time constant, it is preferable to substitute γ_2 by $\gamma_2(1 + g_1(t))$. In such a way the neglection of the time constant of current control loop can be compensated.

In the case of the sampled procedure the adaptation algorithm will be as follows:

$$\begin{aligned} g[k] &= g_1[k](\omega_r[k] - \omega[k]) + g_2[k], \\ g_1[k] &= g_1[k-1] + \Gamma_1 \varepsilon[k](\omega_r[k] - \omega[k]), \\ g_2[k] &= g_2[k-1] + \Gamma_2 \varepsilon[k](1 + g_1[k]). \end{aligned} \tag{16}$$

It follows from the deduction that the asymptotical stability is valid if the model error does not change the sign. If the sign will change then the adaptation has to be stopped for one sampling step in general.

The control in recursive form supplemented with the adaptation algorithm will be as follows:

$$\begin{aligned} u[k] &= u[k-1] + K_i \varepsilon[k](1 + g_1[k]) + K_p((1 + g_1[k])e[k] \\ &\quad - (1 + g_1[k-1])e[k-1]), \\ g_1[k] &= g_1[k-1] + \Gamma_1 \varepsilon[k](\omega_r[k] - \omega[k]), \end{aligned} \tag{17}$$

where

$u[k]$	is the output of speed controller at the k-th sampling period,
$e[k]$	is the error signal $(\omega_r[k] - \omega[k])$,
$\varepsilon[k]$	is the model error $(\omega_m[k] - \omega[k])$,
$\omega[k]$	is the speed feedback signal,
$\omega_r[k]$	is the speed reference signal,
$g_1[k]$	is an adaptation signal,
$[k-1]$	is the previous sampling period,
Γ_1	is a positive constant,
K_p, K_i	are the proportional and the integral constants.

This control has been tested by the simulation program developed in the Department. First the adaptation has been examined without load and current limitation as the motor and the adaptation do contain non-linearities. In the interest of the adaptation stability the speed of change of the adaptation signal g_1 has to be limited. The signal g_1 can result in significant oscillations without limitations as the change of the signal is possible in discrete times.

The current limitation can result in further problems. This limitation hinders the tracking of the model, to the effect of the above signal g_1 will be too large or it can change in the reverse direction. For the elimination of the above problem the signal g_1 is not to be changed in the period of the limitation.

The digital output signal determined by the software is converted by a D/A converter to an analog reference signal. The sampling time of speed loop is about 3 ms. The speed feedback signal is determined by calculating the difference between the actual position value and the previous one of the resolver to digital converter. The measuring period (3 ms) ensures an accuracy of ± 5 rpm.

3.3. *Angle Control*

The Position Decoder in *Fig. 1* contains the resolver to digital converter, the oscillator circuit and two latches. The resolver to digital converter is set to a 12-bit resolution, which corresponds to a disc with 1024 marks (using the usual quadratic encoder interface). The microcontroller reads the code at every sampling period of 244 ms and calculates the speed at every 12^{th} sampling period. In such a way the Angle signal of the rotor and the Speed signal are calculated by the software.

The smallest digit bit of the digital output of the resolver to digital converter gets to the input of one port of the microcontroller. The Timer 2 circuit of the microcontroller counts the change of this signal. By the counting the angle resolution is divided into halves, which corresponds to a disc with 512 marks. At matching the contents of Timer 2 and compare register, the output signal of compare register changes to a level, determined earlier. All three motor phases have one-one compare register and one-one output belonging to them. Three outputs control one-one phase switching IGBT via the Buffer circuit.

The user program calculates the firing angle at every 15 degree, it sets the contents of the compare register and prepares the output to start the firing at matching the compare events. The output signal change, induced by the hardware of microcontroller starts an interrupt request at the same time, based on which the interrupt routine loads the compare register with a value corresponding to the turn-off angle and prepares the output for switching-off the transistor.

As Timer 2 can count in one direction only, it is unable to follow the reverse of speed through the hardware. For this reason and because of starting problems, the user program controls the outputs directly taking the position angle of Resolver to digital converter into account below a predetermined speed. At small speeds the time-lag caused by sampling and the calculation result in a small angle error. This additional error is smaller than 2% related to the 15 degree range below 150 rpm. The frequent shift between control modes can be avoided by applying a suitable hysteresis band.

At high speeds the application of the firing in-advance is necessary, which completes the software by using the Speed signal. As a result of calculations, the contents of compare register will change.

It is worth remarking that a minimal turn-on time is necessary for the safe operation of PWM inverter of fixed frequency. The minimal turn-on time results in a minimal voltage at the star-point of motor phase windings. Because of the minimal voltage at no and small load, either small or zero speed reference signal will result in a high speed of the motor. To prevent this detrimental effect the Current controller reads the Current signal by the help of an inner analog-digital converter of the microcontroller and will intervene in the process if it seems necessary.

There are two ways for solving the above problem: either we inhibit the PWM inverter for some strokes or we switch off the phase transistors accomplishing a secondary choppering. The latter solution has been chosen, represented by the arrow between the Current Controller and the Angle Controller in *Fig. 1*.

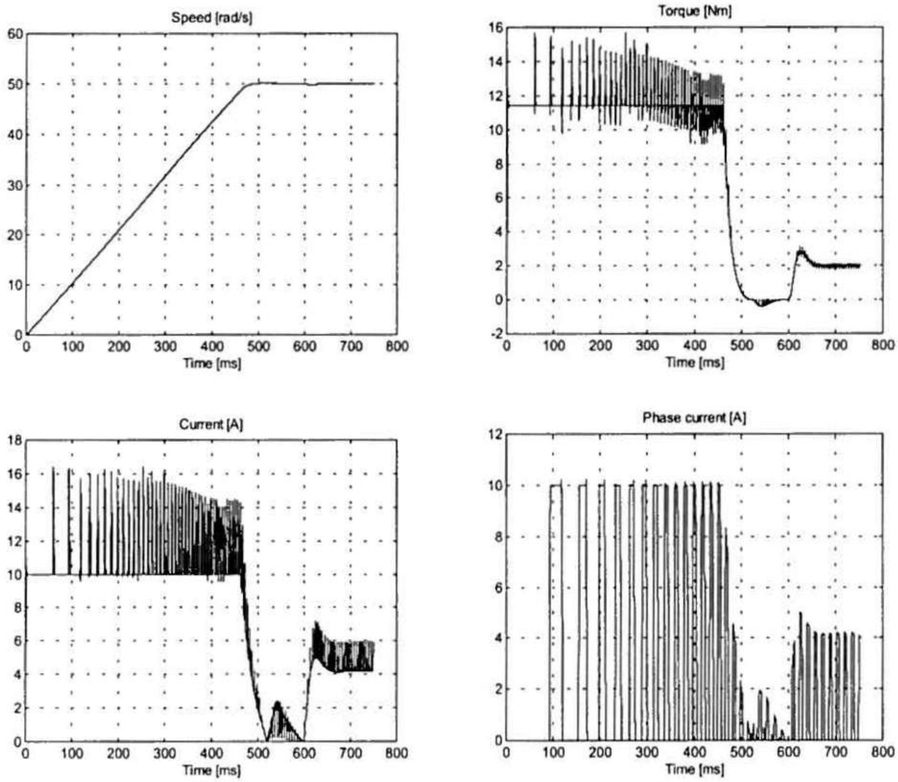

Fig. 4. Simulation results with speed controller of PDF-type

3.4. Ripple Reduced Control Method

The ripple free operation can be realized only with a current waveform depending on the angle, speed and torque [5]. The proposed ripple reduced method changes only the turn-on and the turn-off angle in function of the speed and current reference. The optimum turn-on and turn-off angles of the SRM drive have been determined by computer simulation based on the measured results of the analysed drive. The optimum solution has been fulfilled by four cycles embedded into each other. Two outer cycles give the current and speed reference signals, while two inner ones provide the turn-on and turn-off angles. By this one-one optimum angle pair can be determined to all operating points.

It can be considered an interesting result that the criterion of the minimum torque pulsation does not provide an optimum solution in all cases. The torque pulsation will be minimum in the speed-current plane only in case if the torque of the motor is relatively small. For this reason a good result can be achieved in a way if the relative, i.e. compared to the torque of motor, torque pulsation is minimised.

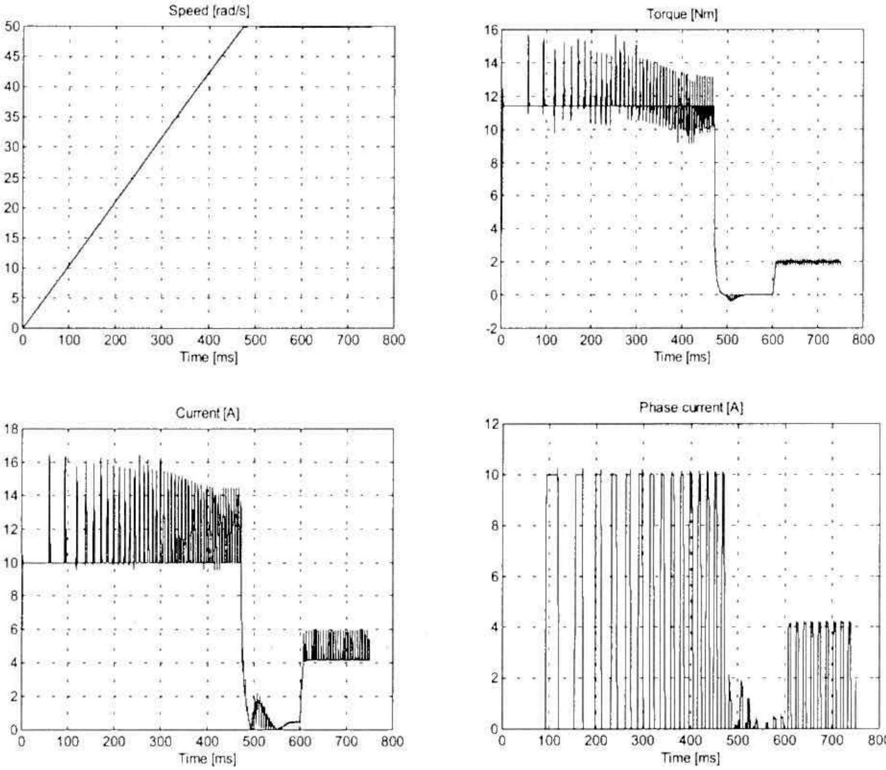

Fig. 5. Simulation results with model reference adaptive speed control

The angle control of the drive determines the actual turn-on and turn-off angles with a two-variable interpolation from the results stored in a look-up table and calculated by the above method.

4. Results

In *Fig. 4* and *Fig. 5* two of many executed simulations are shown. *Fig. 4* shows the run-up with speed controller of PDF-type (an integral element with Proportional and Derivative Feedback [2]), while *Fig. 5* with model reference adaptive control (*Eq.* (17)) and in both cases with turn-on and turn-off angles depending on the speed and current reference and with current reference compensation (*Eq.* (1)). The effect of sharp load-change ($t = 600$ ms) can also be seen.

The tests were completed by the described drive system. The test results have supported our theoretical investigations. The oscillograms in the following figures illustrate some typical starting curves and wave forms. The loading machine was a

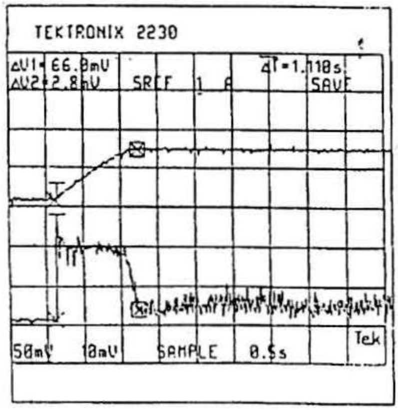

Fig. 6. Oscillograms of the speed and current, PDF-type control

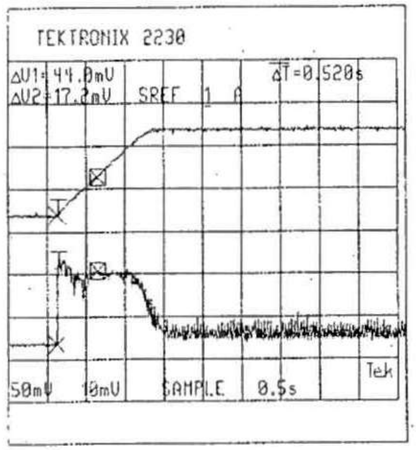

Fig. 7. Oscillograms of the speed and current, MRAC-type control

DC motor. Its inertia is about a triple of that of SRM.

The experiences show that the model reference adaptive control suggested in this paper works without overshooting. Though this method requires a longer calculation period it is less sensible to the variations of parameters.

Figs. 6 and *7* show the speed and current curves in the course of starting without current reference compensation. In all figures the upper curve is the speed (1000 rpm), the lower one is the current flowing in the common point of stator windings (10 A/div). *Figs. 6* and *7* are related to the no-load operation mode. A PDF-type control is applied in the case of *Fig. 6*, while a model reference adaptive control (MRAC) is used in the case of *Fig. 7*.

5. Conclusion

The paper proposes a simple control method for SRM drives. The proposed ripple reduced control method changes only the turn-on and the turn-off angles depending on the speed and current reference. The modification of the current reference is suggested for a simpler four-transistor inverter. The experiences show that the model reference adaptive control suggested in this paper works without overshooting. Though this method requires a longer calculation period it is less sensible to the variations of the parameters.

The simulations and experimental results demonstrate that the proposed method is a promising tool to control the SRM drives.

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