Investigation of Model Reference Parameter Adaptive SRM Drives
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Abstract - Nowadays switched reluctance motor (SRM) drives have been widely used in the field of controlled electric motor drives. 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 ripple free operation can be realize only with an current waveform depending on the angle, speed and torque. 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 the ripple –reduced method that it does not need the real-time calculation or measuring the motor torque. So it can be implemented on a cheap microcontroller. The test of this control method was performed in an experimental drive system. A SRM of 6/8 pole and 4 kW rated power was used. Simulation and experimental results are presented.

I. 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 criterium or by hyper-stability principle.
- Its planning and application is closely related to the using of computer methods.
- Simple realisation of the control algorithm.

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

II. DRIVE SYSTEM

The block scheme of the examined drive system is shown in Fig.1. A 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.

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.

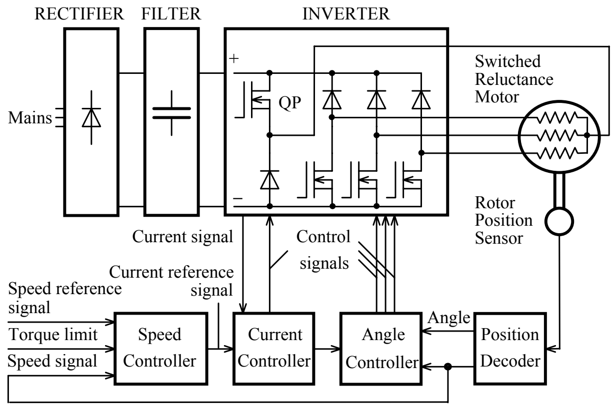

Fig.1. Block scheme of drive system

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.

The fundamental part of the control unit is a single-chip microcontroller. It contains a clock generator with 12 MHz, 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, a compare/capture unit.

The microcontroller executes the instructions of program stored in EPROM memory. The actual values of program parameters and variables are stored in the inner, 256x8-bit

RAM of the microcontroller, which also serves for stack. There is no outer RAM.

The EEPROM memory serves for storing the necessary data, parameters and fault codes after turning off the equipment. The EEPROM memory preserves the data without supply voltage.

Substantial element of the control board is the Resolver to digital converter of type AD2S280. As its operation mode is selected so that it should be in continuous operation, a permanent overwrite in Latch1 and Latch2 takes place. Before reading in the data of the two latches, i.e. to input the digital value of rotor position to the microcontroller, the microcontroller gives an INHIBIT signal for interrupting the overwrite hindering this way the reading of false data to the microcontroller.

We can accomplish a serial and full duplex data transmission according to the standard RS 232 C towards the higher control level. The Serial interface produces the signal levels of data transmission. In this way it is possible to define the Speed reference signal, the Torque limit, other parameters in a digital way and to modify them in operation.

III. OPERATION

The simplified operation principle of the SRM is as follows: When current is passed through the phase (stator) windings the rotor tends to align with the stator poles, that is it produces a torque that tends to move the rotor to a minimum-reluctance position. When a rotor pole is approaching the aligned position of the excited stator pole, positive (motoring) torque is produced, regardless of the direction of the current.

If the phase windings of the stator are supplied by a current of constant peak-value and square-waveform (power semiconductors are excellent for this purpose) with a frequency according to the required speed, then the task seems to be solved, as the SRM is a synchronous motor. But in this case, a significant torque pulsation occurs depending on the load at high currents which results in an additional overheating, while at small currents the motor can fall out of the synchronism and stops. This is why the task is dissolved for two parts. Partly a speed controller is applied, partly the switching on and off of motor phase windings is controlled by a position sensor on the motor shaft. For example at the increasement of speed reference signal originating from the error signal, the controller enlarges the current and the motor will accelerate. At the same time, the shaft position sensor controls the commutations among the phases by ensuring the accelerating torque continuously, otherwise the improper commutations will cause decelerating torque within one period. In the steady-state condition the torque produced by the controller is in balance with the loading torque.

IV. CONTROL OF SRM DRIVE

A. 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.) it is suitable a simpler four-transistor inverter and is not necessary a six-transistor one 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 at starting the conducting state can be forced by the modification of the current reference signal [6]:

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

where:

i_r is the current reference signal,

i_j is the current signal of phase j ,

u is the output of the speed controller,

C_j is the control signal of phase j (0 or 1).

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

B. Speed Control

Based on the Speed reference signal, the Speed Controller produces the Current reference signal taking the Speed signal and the Torque signal into account.

The output signal of speed controller can be used preferably for torque command signal to ensure fast dynamics. One solution for the above task is to insert a torque control loop between the speed control loop and the current one. The formation of an actual torque signal makes the torque control complicated and expensive. Knowing the intermediate dc voltage and dc current, the speed and the efficiency, the approximate calculation of the torque is possible [3]. Because of the above difficulties, other, simpler and more effective methods have been analysed. Their common characteristics are that fast dynamics are ensured by applying a constant gain in the speed control loop.

Neglecting the saturation, the motor torque is proportional to the square of current. This means that the current reference signal can be composed from the torque reference signal, produced by the speed controller by the help of a square-root function after composing its absolute value. The motoring and generating operation modes of the drive can be determined from the sign of torque reference signal and the momentary direction of rotation. The change of operation mode is fulfilled by the Angle Controller, illustrated by the arrow between the Current Controller and Angle Controller in Fig.1.

The saturation of motor, depending greatly on the motor construction, complicates the transient analysis of the system further. In consequence of the saturation, the square relation between the current and the torque will be valid approximately only.

Furthermore, the inner voltage of motor at high speed causes an additional problem, since the current control of PWM cannot produce constant current, given by the command signal. The decrease of torque can be partitionally compensated by angle control, namely by advancing the turn-on angle, resp. by changing the turn-off angle. Due to the above, the starting time of drive increases, on the one hand, and the gain of speed control loop decreases, on the other hand. As at high speed the constant value of maximum torque cannot be reached because of physical limitations, the only purpose for the controller is to maintain the constant gain of loop. This goal can be achieved at least in two ways.

One solution for solving the task is the compensation in the function of speed and torque, using calculated values from a look-up table, stored in the memory.

The other solution is the application of an adaptive control. The up-to-date control theory and the computer hardwares make the application of adaptive control possible, taking the actual parameters of control circuit changing during the operation into account. For controlling electric drives, algorithms ensuring fast adaptation and having simultaneously a fairly few calculation requirement can be applied effectively [4].

A model reference adaptive control is used for the speed control [4] [7] [8] [9] [10]. Such an adaptive control has been successfully elaborated by using a suitable chosen Lyapunov function to compensate the gain of the speed control loop.

The adaptive control of servo-drives with cascade arrangement is most effective if it is applied in the inner loop containing the effect of variable parameters directly, i.e. the inertia (J_m) and/or torque factor (k_m). The speed control implemented with PI controller is of cascade arrangement in fact as it contains an inner, proportional feedback loop (PF controller, [2], Fig.2.). A one-storage proportional element can describe this inner loop neglecting the time constant of the closed current control loop. By this our adaptation algorithm will be simplest.

The relation between the accelerating current determining the dynamic torque and angular velocity can be given by the following transfer function:

$$Y_{\omega,(i-i_t)}(s) = \frac{A_i}{s}; \tag{2}$$

where $A_i = \dfrac{k_m}{J_m}$.

The arrangement of control circuit can be seen in th Fig. 2.

The section determined by the transfer function $Y_{\omega,(i-i_t)}(s)$ is fed back by a proportional member of gain K_P. The task is to change the gain K_P in such a way that the product $A_i K_P$ should remain constant despite the change of parameter A_i.

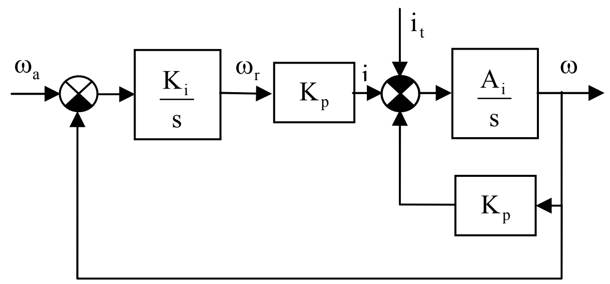

Fig.2. Block scheme of speed control

The transfer factor of the inner closed loop is given by the reciprocal $(1/K_P)$ of the feedback member that is not constant because of the torque factor change. In the consequence of above the loop gain of the outer speed control loop would change as well. In the interest to get a one-storage element with unity transfer factor we have to insert a member with gain K_P between the integrator of PF controller and reference signal of the inner loop. The one-storage reference model with time constant T_m gets the sum of the input signal of above member (ω_r) and the signal ω_{tm} compensating the load effect for the model. So the dynamics of the reference model can be described by the following differential equation:

$$\dot{\omega}_m T_m + \omega_m = \omega_r + \omega_{tm} \tag{3}$$

in which the index m refers to the model.

If we divide Eq. (3) with T_m and apply the designation $q_m = 1/T_m$, we get the following equation:

$$\dot{\omega}_m + q_m \omega_m = q_m(\omega_r + \omega_{tm}) \tag{4}$$

The differential equation of the one-storage controlled plant is as follows:

$$\dot{\omega} + (K_p A_i)\,\omega = (K_p A_i)\,\omega_r - A_i i_t \tag{5}$$

The factor K_P can be described as the sum of K_{P0} determined on the mean A_i and ΔK_p accomplished by the adaptation algorithm. So thus:

$$K_p A_i = (K_{p0} + \Delta K_p)\,A_i = q + \Delta q \tag{6}$$

where K_{P0}, and q are constant.

In this case we assume that the change of A_i is slow from the viewpoint of adaptation, therefore the effect of this change can be neglected.

Substituting Eq. (6) into Eq. (5) we get:

$$\dot{\omega} + (q + \Delta q)\,\omega = (q + \Delta q)\,\omega_r - A_i i_t \tag{7}$$

By using Eq. (4) and (7) and substituting the expression of model error $\varepsilon = \omega_m - \omega$ the dynamic equation will be:

$$\dot{\varepsilon} = -q_m \varepsilon + x\,\omega - x\,\omega_r + q_m \omega_{tm} + A_i i_t \tag{8}$$

where $x = (q + \Delta q) - q_m$.

For the sake of that the system should follow the model, the error of dynamics is stable asymptotically. In the interest of determination Δq the following Lyapunov function should be composed:

$$V = \frac{1}{2}\left(\varepsilon^2 + \beta\, x^2\right), \tag{9}$$

where β is a positive number.
At choosing the Lyapunov function both purposes, i.e. the termination of the model error and loop gain deviation have been taken account.
The time derivative of the Lyapunov function is:

$$\dot{V} = \varepsilon\ \dot{\varepsilon} + \beta\, x\, \dot{x} \tag{10}$$

Substituting Eq. (8) into Eq (10) the following equation is valid:

$$\dot{V} = -q_m \varepsilon^2 + \varepsilon\, x\, \omega - \varepsilon\, x\, \omega_r + \varepsilon\,(q_m \omega_{tm} + A_i i_t) + \beta\, x\, \dot{x} \tag{11}$$

If

$$\varepsilon\, x\, \omega - \varepsilon\, x\, \omega_r + \beta\, x\, \dot{x} = 0\,,$$

that is

$$\dot{x} = \varepsilon\,(\omega_r - \omega) / \beta \tag{12}$$

and

$$\varepsilon\,(q_m \omega_{tm} + A_i i_t) < 0 \tag{13}$$

then

$$\dot{V} < -q_m \varepsilon^2\,. \tag{14}$$

The above equation is a negative definite function that shows the asymptotic stability of the error dynamic Eq. (8).

By using the relations (6), (8) and (12) the following adaptation algorithm is true:

$$\Delta \dot{K}_p = \gamma\, \varepsilon\,(\omega_r - \omega) \tag{15}$$

where γ may be an arbitrary positive number.

The inequality (13) shows how we have to change the signal ω_{tm} representing the load of model.

If

$$\varepsilon > 0, \quad \text{then} \quad \omega_{\mathrm{tm}} < -\frac{|i_t|_{\max} A_i}{q_m}, \tag{16}$$

respectively if

$$\varepsilon < 0, \quad \text{then} \quad \omega_{\mathrm{tm}} > \frac{|i_t|_{\max} A_i}{q_m}.$$

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.

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

The digital output signal of speed controller 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 ensures an accuracy of ± 5 rpm.

B. 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 in and calculates the speed at every 12th sampling period. In such a way the Angle signal of 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 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.

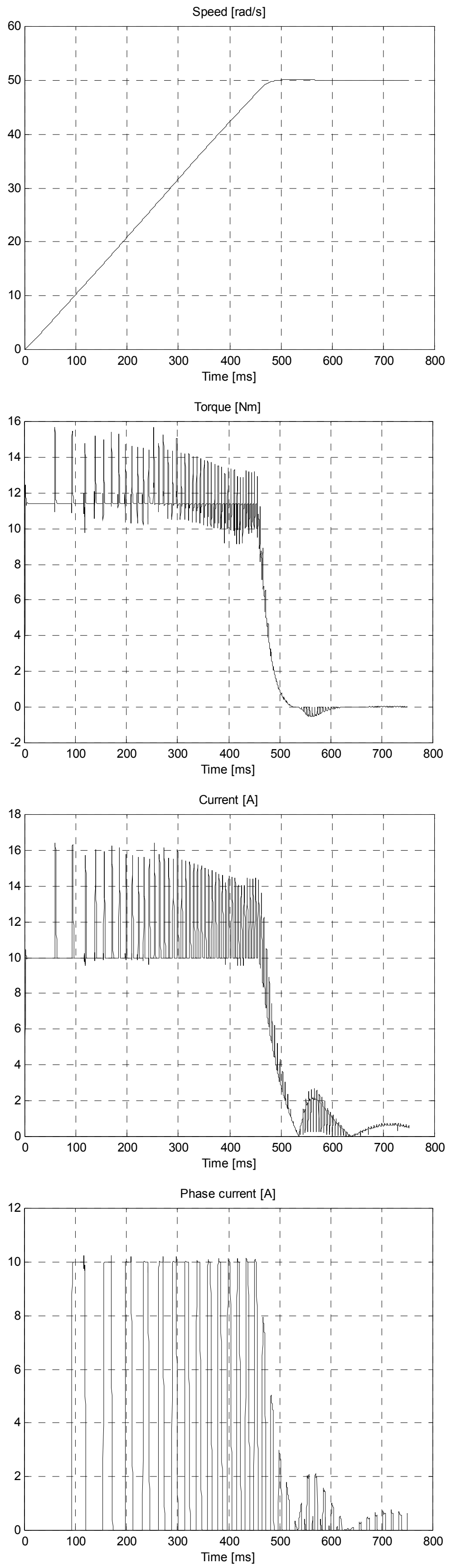

Fig.3. Simulation results with PF-type speed control (ΔK_p =0)

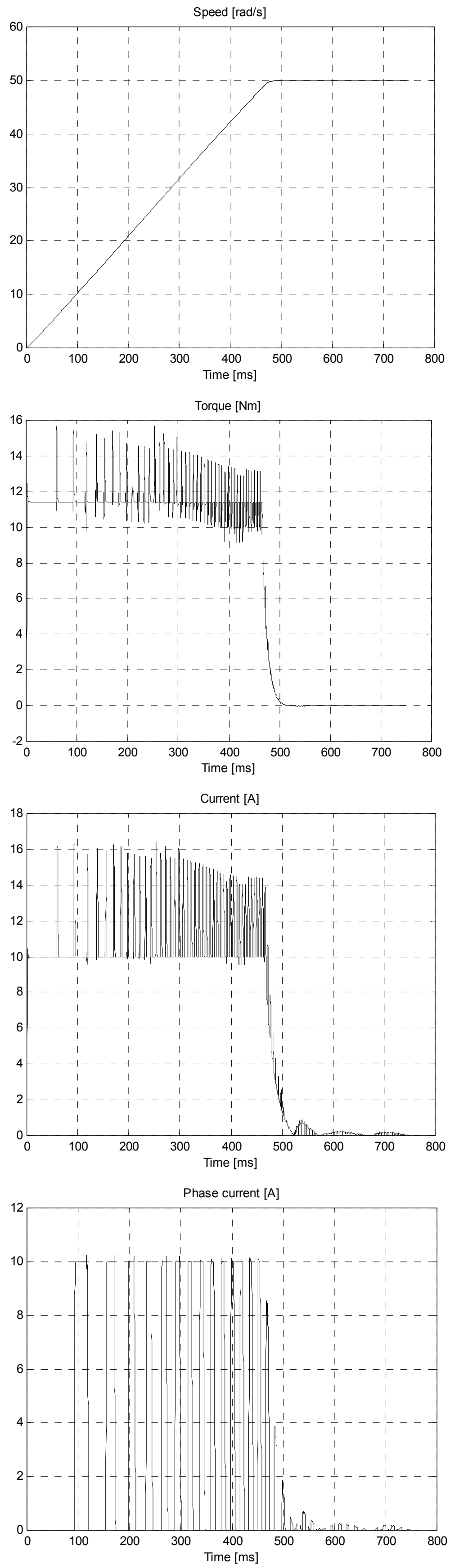

Fig.4. Simulation results with model reference parameter adaptive speed control

At high speeds the application of the firing in-advance is necessary, which the software completes 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 a 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 in 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.

C. Ripple reduced control method

The ripple free operation can be realize 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 has 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 criteria 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 that case if the torque of the motor is relatively small. For this reason a good result can be achieved in such a way if the relative, i.e. compared to the torque of motor, torque pulsation is minimised.
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.

IV. RESULTS

In Fig. 3 and Fig.4 two of many executed simulations are shown. Fig. 3 shows the run-up with speed controller of PF-type (an integral element with Proportional Feedback, $\Delta K_p = 0$), while Fig. 4 with model reference parameter adaptive control (Eq. 15, Eq. 16) 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 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 DC motor. Its inertia is about a triple of that of SRM.

Fig. 5. shows the speed and current curves in the course of starting without current reference compensation.
The upper curve is the speed (1500 rpm), the lower one is the current flowing in the common point of stator windings (10 A/div). It is related to the no-load operation mode.

V. CONCLUSIONS

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 adaptive control suggested in this paper works without experiences show that the model reference parameter 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. The simulations and experimental results demonstrate that the proposed method is a promising tool to control the SRM drives.

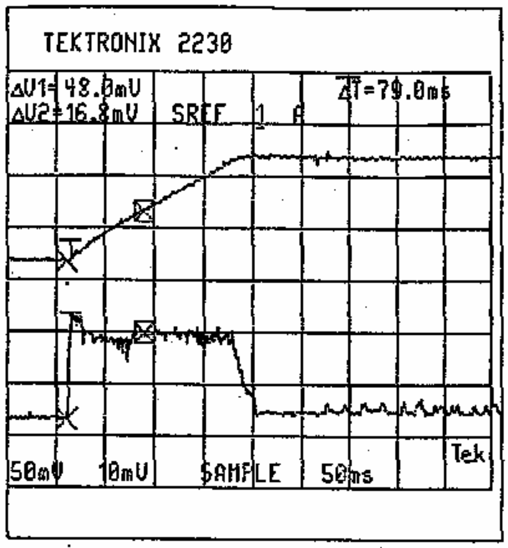

Fig.5. Oscillograms of the speed and current, model reference parameter adaptive-type control

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