Sensorless Control of a SRM at Low Speeds and Standstill Based on Signal Power Evaluation

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Abstract - **The Switched Reluctance Motor (SRM) has gained considerable attention in recent years mainly because of its low production costs, which result from its simple structure. To increase its simplicity further, research has been done to replace the regular control, which needs a position sensor, with sensorless control. To date most papers on the sensorless control of a SRM have dealt with the running machine and little attention has been paid to very low speeds or standstill. This paper presents a method which is designed to estimate the rotor position at low speeds and at standstill. The method uses a sinusoidal high frequency (600 Hz) current signal which can be injected into the motor by the converter. The voltage signal which depends on the rotor position is measured and its signal power is evaluated. Thus the rotor position can be estimated. Practical results are presented demonstrating the feasibility and accuracy of this method.**

I. INTRODUCTION

Due to the advantages of sensorless control, such as increased robustness, smaller total volume, no sensor-cable and lower costs, sensorless control methods for all kind of machines have been developed. The main methods for the sensorless control of a switched reluctance motor (SRM) can be categorised as follows:

- **voltage integration / flux methods** (e.g. [1] and [2]): The basic concept of this method is to integrate the voltages of the energised phases. As the phases of a SRM are not continuously excited the integrator can always be reset to zero when a phase is not energised. Therefore, a drift of the integrator does not cause any problems. The integrated voltages yield the flux which is a function of the rotor position. The measured flux is often compared with a threshold. This threshold corresponds to a certain rotor position which indicates when the next phase is to be excited.

- **inductance measurement by current wave-form detection or by current pulses** (e.g. [3] and [4]): As the inductance of a SRM changes with the rotor position, it can also be used to estimate the rotor position. By monitoring the current wave-form (i.e. measuring di/dt) the inductance can be obtained. Current pulses are injected into the non-energised phases of the motor at standstill and low speeds. Frequently the amplitudes of the current pulses of a certain time length are measured and compared with a threshold as well.

- **signal modulation techniques** (e.g. [5] and [6]): Signal modulation techniques inject a sinusoidal current in a non-energised phase and measure the sinusoidal voltage or vice versa. Voltage and current are determined by the complex resistance which mainly consists of the position depending inductance. Both the phase, the amplitude and the frequency of the signal can be modulated and used for the detection of the rotor position. An additional signal generator injects the sinusoidal high frequency (5 kHz in [5] or 25 kHz in [6] respectively) signal into the machine. Semiconductor switches connect or disconnect the signal generator with the machine.

- **Luenberger Observer / Kalman Filter based methods** (e.g. [7] and [8]): The Luenberger Observer and the Kalman Filter are similar methods of implementing the sensorless control. They use a model of the SRM in state space format and they both have a Luenberger Observer type structure. The regular currents and voltages of the machine are measured and used as input values for the observer or the filter respectively. The Kalman Filter also takes measurement and system noise into account in its calculation of the observer gains. The position and the speed of the machine can be estimated by including them into the state vector of the SRM model. Both methods offer a continuous estimate of the rotor position whereas most other methods yield just one switching pulse per electric cycle. But they also need a very high computing capacity.

A more detailed survey of the literature on sensorless controlled SRM's is e.g. given in [9] and [10].

Although the startup from standstill can be achieved with the methods presented above, no special attention has been paid to the problems of sensorless speed control at very low speeds or the sensorless position control.

II. CONTRIBUTION OF THIS PAPER

This paper presents a sensorless control method which is specifically designed for low speeds and standstill. If the motor also runs at higher speeds one, of the above described sensorless methods must be used.

Basically, the method presented in this paper is a signal or amplitude modulation technique. It differs from the amplitude modulation method presented in [5] or [6] in two ways. First, no additional converter / hardware is needed to inject the modulation signal into the machine. Instead, the converter of the motor itself can be used. And second, a different method is used to evaluate the modulation signal. With this method a continuous estimate of the rotor position can be obtained, which could be used for current shaping purposes at low speeds in order to reduce the torque ripple of the SRM or for a sensorless position control of the SRM.

III. SWITCHED RELUCTANCE MOTOR

Because of its geometry, and the fact that in general only one pole of the SRM is excited at a time, it is generally assumed that the flux linkage within one pole depends only on the rotor position and the current in this pole. This implies that the cross-coupling between the phases is neglected. Under these assumptions equation (1) describes the electrical behaviour of a switched reluctance motor:

$$u_j = R_j \cdot i_j + \frac{d\,\Psi_j(i_j,\theta)}{dt} \Leftrightarrow \tag{1}$$

$$u_j = R_j \cdot i_j + \underbrace{\frac{\partial \Psi_j(i_j,\theta)}{\partial i}}_{L_{inc}(i_j,\theta)} \cdot \frac{d i_j}{dt} + \underbrace{\frac{\partial \Psi_j(i_j,\theta)}{\partial \theta} \cdot \frac{d\theta}{dt}}_{back\text{-}emf} \tag{1a}$$

The subscript j ($1 \le j \le m$) represents the stator phase, where m is the number of all stator phases. The flux-linkage, Ψ, is a function of both the current, i, and the rotor position, θ. Ψ is shown in figure 1. The saturation for higher currents and thus the magnetic non linearity can clearly be seen.

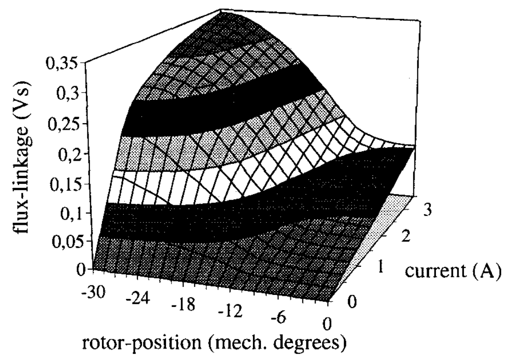

Figure 1: *Flux linkage of one phase over rotor position and current*

If the speed is low, the back-emf can be neglected. Moreover, the ohmic voltage drop can be neglected too if di/dt is high, as it is the case when high frequencies are used. With these simplifications the phase voltage equals the voltage drop over the inductance. Furthermore, the amplitude of i must be kept small, so that saturation does not occur. Then $L_{inc}(i_j,\theta)$ depends only on θ. Finally equation (1) can be simplified as:

$$u_j = L(\theta)\frac{d\,i_j}{dt} \tag{2}$$

Equation (2) is the basic equation for the amplitude modulation. A sinusoidal current with a constant frequency and a constant amplitude will result in a voltage of which the amplitude varies with $u = f(L(\theta))$ - that is with θ.

IV. DESCRIPTION OF THE METHOD PRESENTED IN THIS PAPER

The amplitude of the voltage signal still has to be measured. Sampling the voltage and using each value u_k alone is very noise sensitive. Therefore, some kind of averaging has to be done. The method presented here is based on evaluating the power of the voltage signal u(t). The power of a periodic signal is defined as (e.g.[11]):

$$P(u(t)) = \frac{1}{T}\int_{-T/2}^{T/2} u^2(t)\,dt \tag{3}$$

As u depends on θ, $P(u)$ depends on θ, too. The power of u(t) for a sinusoidal signal equals:

$$P(u(t)) = \tilde{u}^2 \tag{4}$$

where $\tilde{u}$ is the r.m.s. value of u. If the signal u is sampled, its signal power may be calculated as:

$$P(u_k) = \frac{1}{N}\sum_{j=1}^{N} u_k^{\,2} = \tilde{u}^2 \tag{5}$$

where N samples must make up one period of the signal. When the machine turns, the position will change within the period $T_{modulation}$ of the signal:

$$\Delta\theta = \frac{d\theta}{dt}\cdot T_{modulation} = \omega \cdot T_{modulation} \tag{6}$$

For $f_{modulation} = 600$ Hz $\Leftrightarrow$ $T_{modulation} = 1.\overline{66}$ ms and $\omega = 200$ rpm $\Delta\theta$ equals already 2 mechanical degrees for the 8/6 motor used in this paper. That is why this method is also limited to small speeds. To run the motor at higher speeds in sensorless control based on an amplitude modulation higher frequencies would be necessary. Therefore, [5] and [6] must use an additional hardware.

The signal power P of the modulated voltage was evaluated according to equation (5). It was a priori measured once for several equidistant values of θ ($\Delta\theta = 1.5°$) and stored in a table from 0° to 60°, because 60° is the electric cycle of a 8/6 SRM. A linear interpolation is used to obtain the signal power between the stored values. Another way to obtain the signal power for a given rotor position and a given sinusoidal current is to calculate it using equations (2) and (3). In this case an additional calculation would have to be carried out and the inductance $L(\theta)$ would have to be measured and stored in a table. Therefore, it is better to measure and store $P(\theta)$ directly.

In sensorless control mode the signal power P of the modulated voltage is measured and compared to its stored value $\hat{P}$ at the estimated rotor position $\hat{\theta}$. This yields an error ε' which is proportional to $P-\hat{P}$ and which depends on $\theta-\hat{\theta}$. However, the error signal ε' is not proportional to $\theta-\hat{\theta}$, because $L(\theta)$ and consequently the signal power $P(\theta)$ are non-linear functions of θ. As the error signal will be used in order to track the rotor position θ, an error in the rotor position $\theta-\hat{\theta}$ should always results in the same error signal. In other words the error signal should be proportional to $\theta-\hat{\theta}$ and ε' should be linearized. This is done by multiplying ε' with $d\theta/dP$ at the estimated rotor position. This derivative can be obtained from the stored values of $P(\theta)$, as well. Finally, one gets an error signal ε which is proportional to $\theta-\hat{\theta}$. This context is once again shown in figure 2.

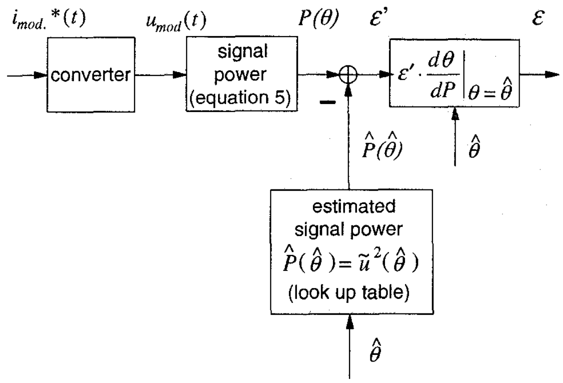

Figure 2: *Block diagram of the used algorithm*

The error signal ε is the input of a tracking observer which is shown in figure 3. As ε is equivalent to $\theta-\hat{\theta}$ the observer is driven by $\theta-\hat{\theta}$. If ε' was chosen as the input of the observer, the dynamic or the poles of the observer would change with the rotor position. The observer tracks both the speed, $\hat{\omega}$, and the position, $\hat{\theta}$, of the machine. It is also described and used for the sensorless control of an AC machine e. g. in [12].

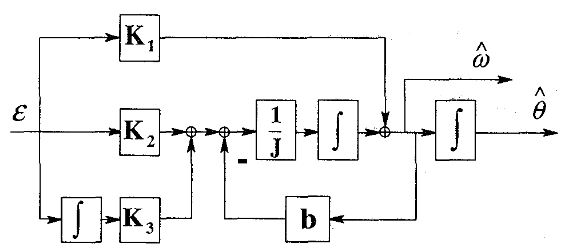

Figure 3: *Block diagram of the tracking observer* [12]

One final problem which still has to be discussed is when the modulation signal is to be injected into a phase. This shall be explained in figure 4. The first task is to energise the phases in order to run the motor. This depends on the sign of the torque. The areas in which the motor is excited are shown both for positive and negative torque. In these areas $L(\theta)$ is the steepest resulting in the biggest positive or negative torque. The modulation should occur when $L(\theta)$ is steep as well, so that a small variation in θ results in a relatively large variation of $L(\theta)$ and therefore a large variation of $u(\theta)$ and $P(u(\theta))$. This means the modulation signal should be injected in the negative torque area when there is a control command for positive torque and vice versa. This is usually done for signal modulation techniques. If the machine turns, the modulation signal has to be switched amongst the phases.

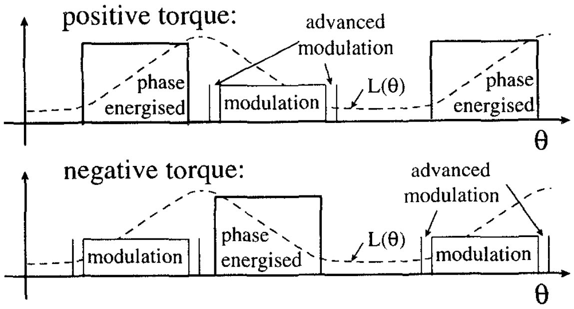

Figure 4: *Areas when a phase is energised and when the modulation signal is injected*

Before the signal power can be calculated at least N samples have to be taken. Therefore, the modulation must be advanced or the area in which the signal is injected must be enlarged. Now, the modulation starts in a new phase while the signal power of another phase is still used for the estimation of θ. The voltage is already sampled and stored so that there are N values when the signal power of this phase must be used for the estimation of θ. Another problem which has to be taken into account is a sudden change in the sign of the commanded torque. The phase which was energised has to be modulated and the phase which was modulated has to be energised when the sign changes. In this case the commanded torque must be delayed for a very short time so that the energised phase can be de-energised and modulation can start.

V. Measurements

All measurements shown in this paper were done in sensorless control. The motor is a fractional horse power 8/6 motor. The algorithm was implemented on a Digital Signal Processor TMS 320 C40 but as the computational effort of the algorithm is very small it could be implemented on a microcontroller as well. In order to take a slower processor into account the sample time of the DSP was set to 330 µs although the algorithm just needs about 30 µs. Another 100 µs are needed for the communication between the DSP and the Host-PC and storing measurement data, which are not needed in an industrial drive. The frequency of the modulation signal was chosen to 600 Hz such that just five samples of the measured voltages were used to determine its signal power. By using more samples the accuracy of the measured signal power can be increased.

The converter is an IGBT converter which has two switches per phase. This algorithm can only be implemented if such a converter is used. Converters which have less than two switches per phase cannot inject both the modulation signal and the regular current into two different phases at the same time. As a converter for a SRM can only inject positive currents, the modulation current must have a small offset:

$$i_{modulation} = i_{offset} + i_{offset} \cdot \sin(2\pi \cdot f_{modulation} \cdot t) \tag{7}$$

The offset chosen was small (i_{offset} = 0.075A) and the frequency high ($f_{modulation}$ = 600Hz) so that they do not affect the control of the drive.

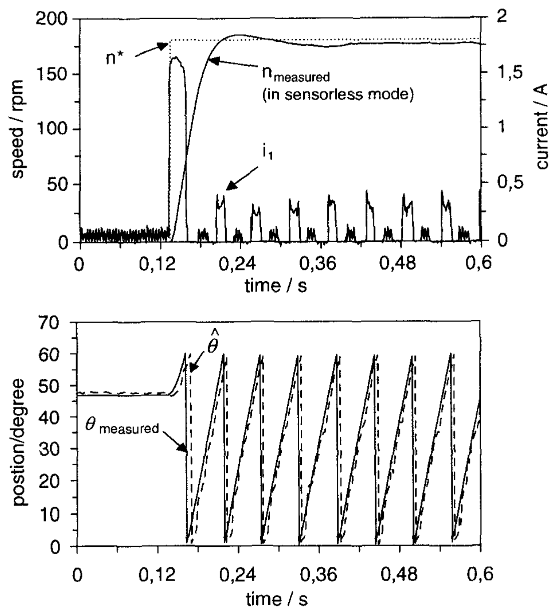

Figure 5: *Start up in sensorless control*

top: *reference speed* (n*), *measured speed* ($n_{measured}$) *and current in phase 1* (i_1)

bottom: *estimated* ($\hat{\theta}$: - - -) *and measured position* ($\theta_{measured}$: ——) *in mechanical degrees*

Figure 5 shows a start up from standstill to 180 rpm. The current in phase 1 demonstrates the ratio between the modulation signal and the regular current used for the control of the machine. As shown the modulation signal is injected in-between the current pulses. Estimated and measured position deviate slightly. The maximum position estimation error lies within about ± 3 mechanical degrees. The position is limited to 60 mechanical degrees as this corresponds to the electric cycle of a 8/6 SRM as mentioned above.

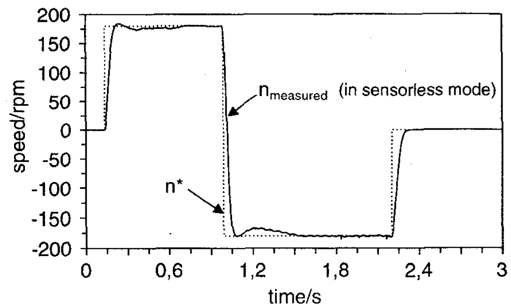

Figure 6: *Start up, speed reversal and stop in sensorless control, reference speed* (n*) *and measured speed* (n_{measured})

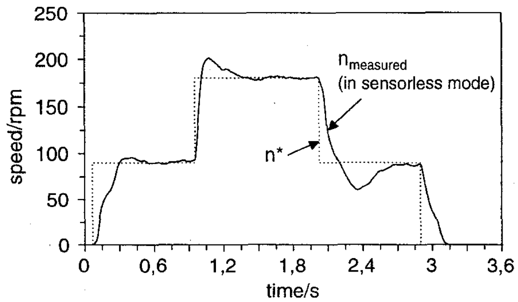

Figure 7: *Change in reference speed in sensorless control; reference speed* (n*) *and measured speed* (n_{measured})

Figures 6 and 7 show different step changes in the reference speed. Speed reversal and stop to standstill can be handled by the sensorless control. The very short delay of the commanded torque, if its sign changes, does not impose a problem. A sudden load variation can be seen in figure 8. The sudden load step still causes a big drop in the actual speed but the actual speed returns to its reference value. Under load the speed ripple increases. The bottom of figure 8 shows the current in phase 1 during the first load step. The ratio between the modulation signal and the load current can be seen again.

Finally, figure 9 shows a step in the reference position of a sensorless position control loop. A simple P-controller was added to the PI-controller of the speed loop, which was used in the measurements shown above. As can be seen a sensorless position control loop can be realised with the proposed algorithm. Since the torque / ampere ratio of the SRM varies largely with the rotor position, the position control is non-linear with rotor position. The P / PI controller structure can not compensate this non-linearity. The dynamic and accuracy still vary with the actual rotor position. This would also happen if a position sensor was used.

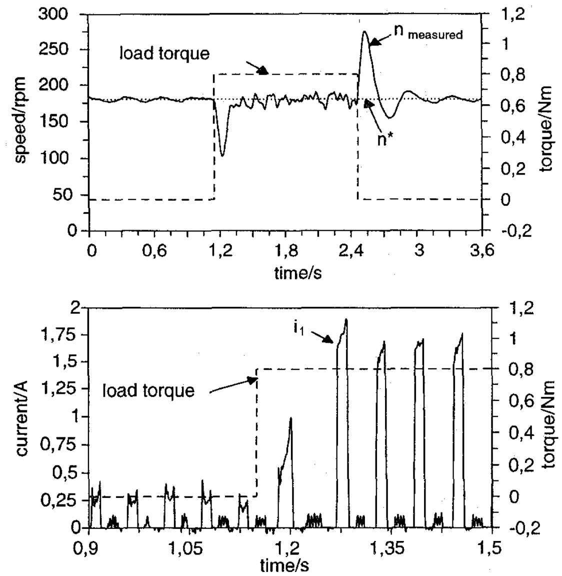

Figure 8: Sudden load variation in sensorless control

top: *reference speed* (n*), *measured speed* (n_{measured}) *and load torque*

bottom: *current in phase 1 and load torque*

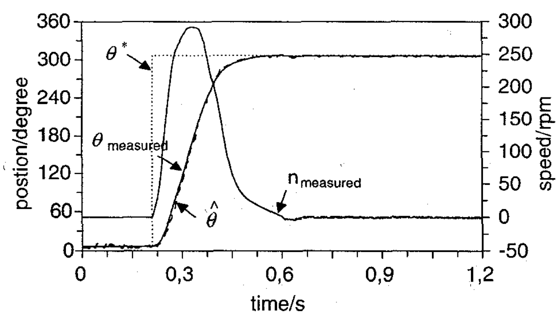

Figure 9: *Change in reference position in sensorless control; reference position* (θ*), *estimated position* ($\hat{\theta}$: – – –), *measured position* (θ_{measured}: ——) *in mechanical degrees and measured speed* ($n_{measured}$)

The next steps in improving this method will focus on increasing the bandwidth of the control loop. One major problem was caused by the varying switching frequency of the converter. As the converter has a hysteresis current control the switching frequency varies with the inductance. This causes problems in filtering the voltage before sampling it. A PWM converter with a constant switching frequency should yield better results.

VI. Conclusion

This paper presented a new method for the sensorless control of a switched reluctance machine. The method is based on an amplitude modulation. A sinusoidal current with a constant amplitude is injected into the machine by the converter. The modulated voltage which depends on the rotor position is measured and its signal power is evaluated. The measured signal power is compared to the signal power that was stored in a table for different rotor positions. The difference between measured and stored signal power drives an observer which tracks both position and speed of the switched reluctance machine. The method is designed for very low speeds and standstill. Measurements show the feasibility and accuracy of this method both for a sensorless speed control and sensorless position control.

VII. References

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VIII. Appendix

Machine data:

motor type	8/6
rated torque	2 Nm
rated speed	1200 rpm
rated current	3.5 A
minimal inductance	28 mH
maximum inductance	280 mH
stator resistance	5.12 Ω