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Non Linear, Time Variant Speed Control of a Single Phase Hybrid Switched Reluctance Motor

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Abstract — A high torque ripple in a given motor always presents a challenge for the speed control, since this ripple may lead to excessive actuation and ultimately may even lead to instability. The conventional solution is to low pass filter the measured speed, but this lowers dynamic control performance. This paper presents a new method to accurately calculate the average speed and a time variant approximation of the integrator to achieve good dynamics. Both simulation and measurement shows a good steady state performance, with a steady state error less than 0.4%.

I. Introduction

Domestic pumps used for central heating is increasingly changing from single phase induction motors to permanent magnet motor types. A candidate motor type for a low cost replacement motor is the single phase hybrid switched reluctance motor (SHSRM).

A switched reluctance motor may have a high torque ripple [1][6], in particular for some single phase motors. If the moment of inertia is small then this may also lead to a large speed ripple. If the classical approach is applied to such a drive, aliasing may occur. Aliasing will introduce measured speed components at low frequencies due to frequency convolution. The aliasing can be limited by a low pass filter, but filtering would lead to additional propagation delay between a real disturbance and the control response.

Another approach is to minimize the torque ripple altogether as it is done in [6-9], but if the torque ripple can not be reduced, no known work deals with high torque ripple speed control. The method proposed in [2] uses a phase locked loop (PLL) to synchronize the control with speed of the rotor. For the PLL to be stable, a loop filter is used. Unless the loop filter is dynamically changed, the loop filter limits the controllable speed range.

This paper presents a method for a high torque ripple drive, where a time variant non linear approach simplifies the control of a high torque ripple drive significantly without sacrificing dynamic abilities.

II. The Single Phase Hybrid Switched Reluctance Motor

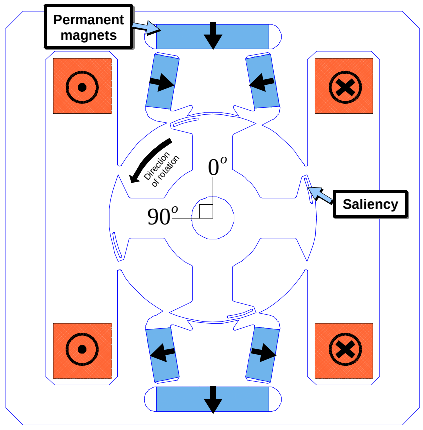

Fig. 1. The hybrid switched reluctance motor is shown here with rotor at an angle of 0 degree. Thick black arrows indicate magnetization of magnets. Due to the saliency in the rotor the normal parking position of the rotor is normally at 1-5 degree.

The motor used is a single phase hybrid switched reluctance motor (SHSRM) with a torque output through one cycle that may vary from 0.06Nm to 1.4Nm. The target application is a pump drive as discussed in [5].

The motor is run in strokes, by alternating between torque produced by the coils and torque produced by the permanent magnets. The motor drive has a single phase asymmetrical inverter and the motor structure is shown in Fig. 2. The machine has six poles on the stator sides, two permanent magnet poles (PM-poles) and four reluctance poles. The permanent magnets in the stator attract a rotor pole when there is no phase current to a parked position. When the two series connected coils are energized, a flux is generated that opposes the flux from the permanent magnets and thus attracts the rotor poles to the reluctance poles. When the coil energy is returned to the DC-link capacitor, the process repeats as the permanent magnet poles attract the rotor once again.

The motor can be described by the following non-linear equations:

$$\begin{bmatrix} \psi \\ i \\ \omega \\ \theta \end{bmatrix} = \begin{bmatrix} \int (v - R \cdot i) dt + \psi_{pm}(\theta) \\ f_i(\psi_{coil}, \theta) \\ \int \frac{\tau_m(\theta, i) - B \cdot \omega - \tau_l}{J} dt \\ \int \omega \, dt \end{bmatrix} \tag{1}$$

where ψ is the phase flux linkage, v is applied phase voltage, R is the phase resistance, i is phase current, ψ_{pm} is the angle (θ) dependent phase flux linkage from the permanent magnets, f_i is the a lookup function that links the coil flux ($\psi_{coil} = \int (v - R \cdot i) dt$) and θ with i, $\tau_m(\theta, i)$ describes the generated motor torque, B is the Coulomb friction, ω is the mechanical speed, τ_l is the load torque, and J is the moment of inertia. Leakage inductance, power converter dynamics and static friction is not considered in (1), but is included in the implemented model used for this paper. See [4] for more details on modelling a SHSRM.

The equations are solved as four simultaneous equations. As can be seen from (1), the speed and angle could be be reformulated in to a state space form. However the flux linkage and current can not be converted in to a state space form. The current (i) is found using the look up table shown in Fig. 3, the output torque is found using the look up table shown in Fig. 2, and the PM flux linkage is found using the look up table shown in Fig. 4.

The output torque for this SHSRM is not constant as can be seen of a plot of the torque look up table shown on Fig 2.

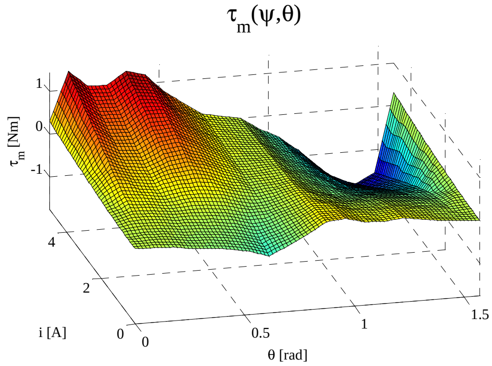

Fig. 2. A plot of the look up table used for finding the torque as a function of the phase current and rotor angle.

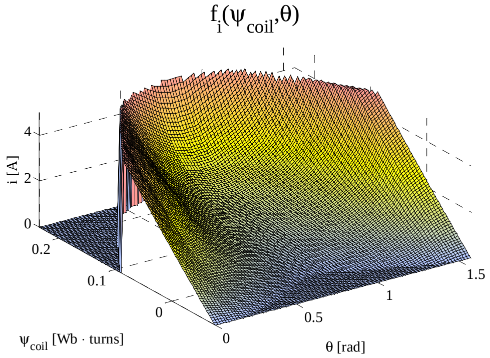

Fig. 3. A plot of the look up table used for finding the current as a function of the coil flux and angle.

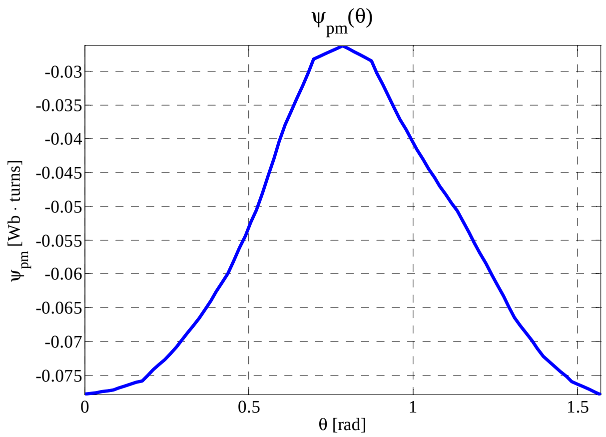

Fig. 4. The permanent magnet flux look up table as a function of the rotor angle.

III. THE CONTROL METHOD

The basic idea in the control proposed in this article is to synchronize the update of the speed loop with the rotor and do proper compensation. The need for handling the control on a per stroke basis was realized in [10], however it only dealt with the current control. The faster the speed of the rotor is the shorter the time between updates. So the update frequency is directly dependent on the state it controls. The proposed control structure is shown on Fig 5.

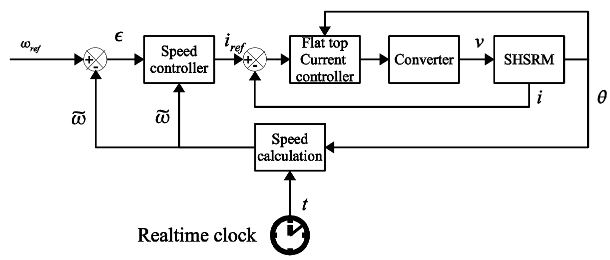

Fig. 5. The proposed control structure. The speed controller and the speed calculation is updated once per stroke. A current reference is generated and used to control the torque production. The turn on and turn of angle for the motor is here kept fixed for test, but changing the angle as a function of speed is not a problem as long as the torque output increases as a function of the reference current.

The inner current loop is a modified proportional current controller. The outer speed loop is updated once per stroke, so the update frequency is not fixed as would be the case for a classical control loop. This is unlike [2], where a PLL is used in connection with angle control to directly control the speed.

The speed controller updates are sequentially numbered (k), and thus the current update is done at the time $t[k]$, and the previous update was done at the time $t[k-1]$. It is important to note that the time between $t[k-1]$ and $t[k]$ varies as a function of the speed.

The measured average speed $(\tilde{\omega})$, defined as the speed at time $t[k]$ $(\omega[k])$, can be calculated as:

$$\tilde{\omega} = \omega[k] = \underbrace{\frac{1}{t[k]-t[k-1]}}_{Averaging\ term} \cdot \underbrace{\int_{t[k-1]}^{t[k]} \omega(t)dt}_{rotor\ distance\ traveled} \quad (2)$$

Since the motor has four poles, one stroke means the rotor has travelled a distance of $\frac{\pi}{2}$ radians. Since the control is kept synchronized with the rotor, the speed is thus calculated by substituting the integral in (2) with a fixed value of $\frac{\pi}{2}$ radians:

$$\omega[k] = \frac{1}{t[k]-t[k-1]} \cdot \frac{\pi}{2} \quad (3)$$

Using (3) the accurate average speed for the period $t[k-1]$ to $t[k]$, can be calculated. The accuracy of the the position of the rotor can be determined.

The error ($\varepsilon[k]$), at the time instance k, between the speed reference (ω_{ref}) and the measured speed is given by:

$$\varepsilon[k] = \omega_{ref}[k] - \omega[k] \quad (4)$$

A simple controller can be made with a proportional controller, but it will have a steady state error. Normally adding an integral term would remove the steady state error, but normal approximations of the continuous time integrator assume a fixed update frequency. The problem here is also that the update frequency is directly linked to the variable it is supposed to control.

The classical Tustin approximation (also known as the bilinear transformation) is actually derived based on a substitution of the Laplace variable s with a Z-domain approximation [3]:

$$s \cong \frac{2}{T_s} \cdot \frac{Z-1}{Z+1} \Leftrightarrow \frac{1}{s} \cong \frac{T_s}{2} \cdot \frac{1+Z^{-1}}{1-Z^{-1}} \quad (5)$$

where T_s is the time between updates.

(5) can be converted from a Z-domain form to a difference equation describing output y due to the input x:

$$\begin{aligned} &\frac{Y(Z)}{X(Z)} = H(Z) = \frac{T_s}{2} \cdot \frac{1+Z^{-1}}{1-Z^{-1}} \Updownarrow \\ &Y(Z) = Y(Z) \cdot Z^{-1} + \frac{T_s}{2} \cdot (X(z) + X(Z) \cdot Z^{-1}) \Downarrow \\ &y[k] = y[k-1] + \frac{T_s}{2} \cdot (x[k] + x[k-1]) \end{aligned} \quad (6)$$

where $Y(Z)$ is the output of the transfer function ($H(Z)$) based on the input $X(Z)$, and k is the time index variable for the difference equation.

In the control method proposed here, the time between samples is not constant. So the term $\frac{T_s}{2}$ will change as a function of the speed of the rotor. The integration is still valid; however the closed loop response would change as if the integral is increasing with speed. However decreasing gain linearly with the speed would be sensitive to variations in the speed that may occur due to sudden load changes. Instead a bilinear term is proposed to compensate for the varying gain by modifying the difference equation in (6) to:

$$y[k] = y[k-1] + \frac{\omega_0 \cdot 2}{\omega[k] + \omega[k-1]} \cdot \frac{k_i}{2} (\varepsilon[k] + \varepsilon[k-1]) \quad (7)$$

where k_i is the integral gain, ω_o is the speed for which the controller is designed. (7) is not valid for zero speed, so a start up method has to be supplied, that ensures that the measured speed is higher than zero. In the simulation and in the tests a fixed initial value of $y[k] = 1.5\,A$ is used. This value is chosen since the motor in the tests will have a quadratic load with a low starting torque requirement.

The output of the speed controller is given as a reference current for the inner current control loop. It is assumed that there is a monotonous relationship between current. The SHSRM is not normally capable of producing negative torque. To ensure that the total drive remains controllable, the load has to be a dissipative load. If the motor needs to lower its speed this can only be achieved by the load dissipating the excess rotational energy and returning the drive to a state where control can be resumed. The machine is here tested with a quadratic dissipative load, and the Coulomb losses are also dissipative. The reference current is given by:

$$i_{ref}[k] = f(x) = \begin{cases} 0, & k_p \cdot \varepsilon[k] + y[k] < 0 \\ k_p \cdot \varepsilon[k] + y[k], & k_p \cdot \varepsilon[k] + y[k] \geq 0 \end{cases} \quad (8)$$

Besides handling saturation of the controllers, the total set of equations necessary to implement the speed control is given by solving (3), (4), (7) and (8) in that order.

IV. Simulation Results

To validate the approach two simulations are performed: One simulation with a fixed update frequency controller with a low pass filter on the measured speed, as can be seen on Fig 6, and another simulation using the proposed method.

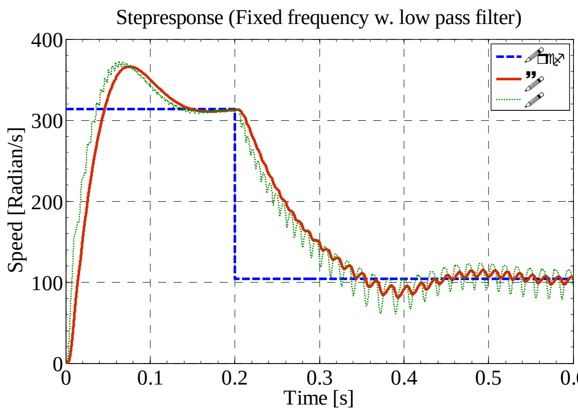

Fig. 6. Simulation of the classical speed control with a fixed update frequency. There is a larger overshoot at high speed compared to low speed. Notice that the estimated speed as seen by the controller has a ripple at low speed. This was done to achieve reasonable dynamics at high speed.

The classical controller shows a sluggish response particularly at low speed. The speed ripple in the estimated speed could be reduced by setting an even lower cut off frequency for the low pass filter, but this would lower the performance at high speed.

The time variant controller step response is shown at Fig 7. The time variant controllers response, shown on Fig. 7, has a slow start up from zero speed since the controller is not valid at zero speed.

The control shows good response at low speed, though at speed the integrator reacts slowly in approaching the reference value past the overshoot. This is due to the fact that motor should generate any negative torque to lower the speed. Lowering the speed is then only achieved by the loss of energy due to the load. This means that response will be asymmetrical, i.e. negative overshoot will look different compared to positive overshoot.

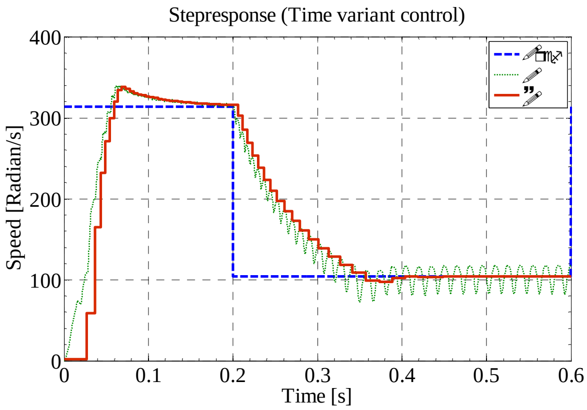

Fig. 7. Simulation of the proposed speed control with a fixed update frequency.

As can be seen on Fig. 7, the time between updates varies as a function of the speed. At 1000 RPM there three times as long per update as compared to at 3000 RPM.

V. Measurement Results

The implementation of the proposed control is done on a Texas DSP F2812, fixed point DSP. The inner current control monitors the angle and signals a parallel main loop, when a current reference update is needed. In the implementation for the proposed control, updates are done four times per revolution. The The machine is controlled by a single phase asymmetric converter connected to the DSP. On the shaft of the prototype motor shown on Fig. 8 is a position encoder mounted. The phase current is recorded on a Textronix TDS3014B sampling oscilloscope.

For testing purposes the motor load is a fan that dissipates approximately 31 W. For the test the motor is kept at 3000 RPM, and the steady state current waveform is captured and compared with the simulated waveform to verify good steady state performance.

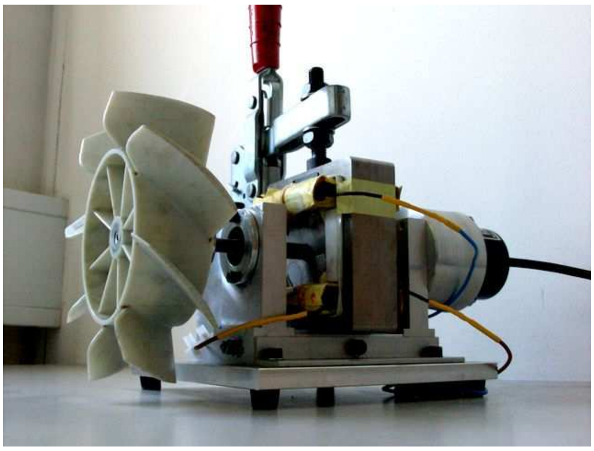
Fig. 8. The prototype motor used for the tests is here shown with the fan that is used as a load for the motor. The motor is rated at an output power of approximately 70W. The motor is mounted in a fixture that enables replacement of the stator stack, and due to this some rotor misalignment may be present. Unlike the final motor, the prototype uses normal ball bearings. In the final motor ceramic wet runner bearings will be used.

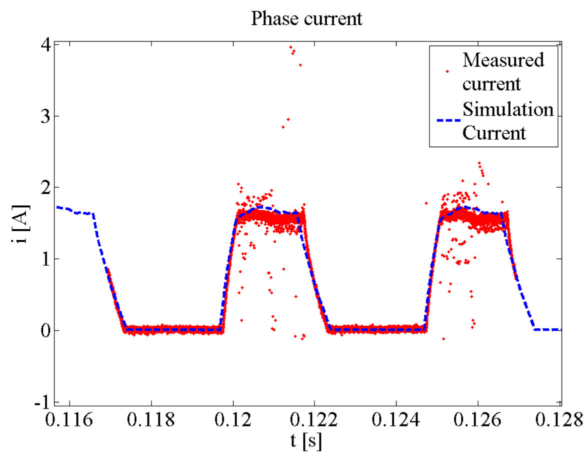

Fig. 9. Measured phase current at 3000 RPM, with a flat top current controller running at 50 kHz

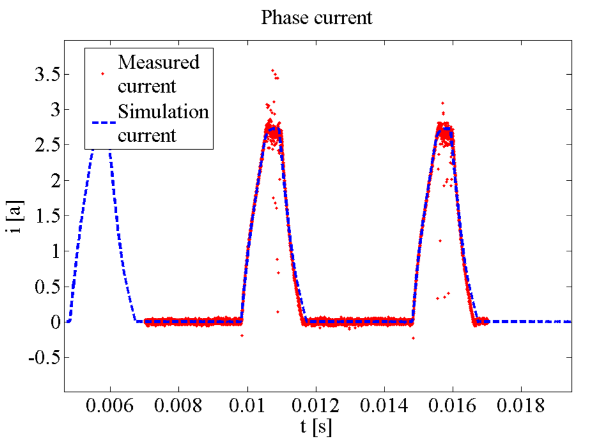

Fig. 10. Measured phase current at 3000 RPM, with a flat top current controller running at 50 kHz. The turn off angle is changed compared to Fig. 8, which leads to a higher torque ripple. Despite the higher torque ripple, the average speed is accurate within 14 RPM.

The simulation and measurement steady state error shows good agreement. The turn off angle variation shown on Fig. 9 and Fig. 10 gives a different speed ripple magnitude, but still the speed error is less than 0.4%.

VI. Conclusion

This paper presented a method where a time variant non linear approach simplifies the control of a drive with a high torque ripple significantly without sacrificing dynamic abilities. The method has two new components: a simple, but accurate, average speed measurement despite a high torque ripple and a new non linear, time variant integrator. The method may also be applied to other motors where torque ripple makes speed control challenging.

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