Current Reduction in Stepping Motor Applications using an Adaptive PI controller based on Linearized Dynamics ⋆

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Abstract: Stepping motors are used in numerous applications because of their low manufacturing cost and simple open-loop position control capabilities. The bulk of the widely used full-step open-loop stepping motor drive algorithms are driven at maximum current to avoid step loss. This non-optimal way of control leads to low efficiency. In order to use stepping motors in a more optimal way, closed-loop control is needed. A previously described sensorless load angle estimation algorithm, solely based on voltage and current measurements, is used to provide the necessary feedback without using a mechanical position sensor. In this paper, an adaptive PI controller which optimizes the current level based on the feedback of the estimated load angle is introduced. Although the current - load angle dynamics are highly non-linear, an adaptive PI controller with the settling time of the current reduction as design constraint is worth considering. Especially because few tuning parameters are required. The described method is complimentary to the popular methodology used to drive a stepper motor, which is based on step command pulses. Measurements validate the proposed approach.

Keywords: Adaptive and robust control, Applications of PID control, Stepping motor, Sensorless motor control, Load angle, Fractional horsepower machines

1. INTRODUCTION

The absence of an expensive position sensor makes stepping motors very appealing for low-power positioning. The rotor position of the machine can be controlled by sending step command pulses. Every time a step command pulse is sent by the user, the rotor of the machine makes a discrete rotation. In this way it is easy to control the position without the explicit feedback of a mechanical position sensor. The two-phase hybrid stepping motor principle is illustrated in Fig. 1(a-b). The stator is equipped with concentrated windings while the multi-toothed rotor is magnetized by means of axially oriented permanent magnets. The north-stack and south-stack of the rotor each have a number of rotor teeth and are shifted with a half tooth pitch relative to each other. By magnetising phase A, the rotor teeth are attracted by the excited stator phase (A^+ and A^-). When a new full-step command pulse is given, the excitation of one phase is released while the second phase is excited.

When the motor is overloaded due to too high load torque or acceleration demands, the relation between the setpoint and the actual rotor position is lost. In most cases this step loss will not be noticed by the stepper controller and will result in malfunctioning of the application. Until today, in order to reduce the possibility of step loss in the current applications, the motor is driven at limited velocity, maximum current level or is over-dimensioned (Lu et al. (2016); Orrs et al. (2016)). This means that the bulk of the stepping motors are driven in a non-optimal way with a low efficiency as result as illustrated in Derammelaere et al. (2014b).

The basic open-loop algorithms are unsatisfactory to drive a stepper motor efficiently. For this purpose, vector-control algorithms, as used in permanent-magnet synchronous machines (PMSM) (Seilmeier and Piepenbreier (2014); Park and Sul (2014)) and induction machines (IM) (Yoon

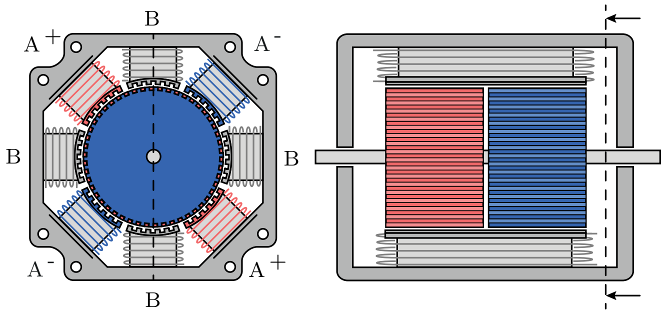

Fig. 1. Two-phase hybrid stepping motor with 50 rotor teeth per stack, front view (a) and cross section (b)

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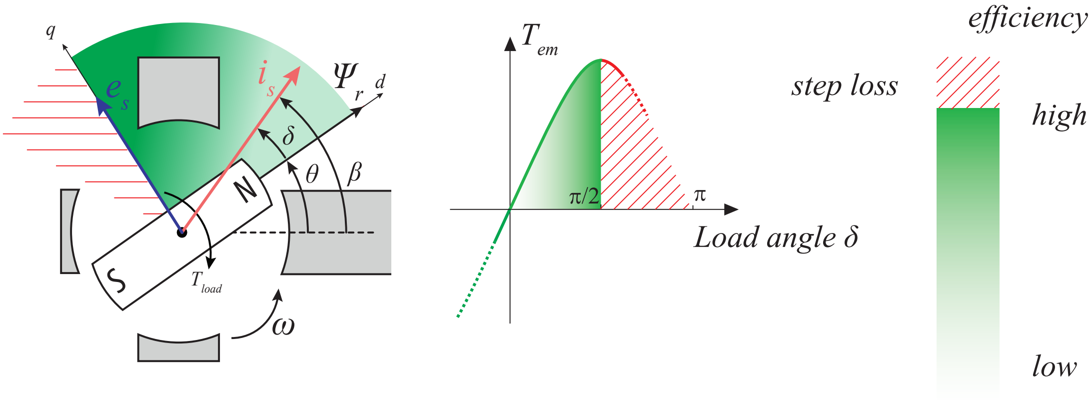

Fig. 2. Relationship between the motor torque T_{em} and the load angle δ at constant current level i_s

(2010); Stojić et al. (2015)), are of interest. The advanced stepping motor drive algorithms described in Kim et al. (2011); Le and Jeon (2010) use control loops typical for PMSM machines. This means that positioning by using step command pulses is impossible when these methods are implemented. This hinders the implementation of these methods in industry.

Therefore in this paper, a closed-loop load angle controller is proposed. The main added value is in the design of a control which adapts the stepping motor current level based on the estimated load angle. A previously described sensorless load angle estimation algorithm, solely based on voltage and current measurements, is used to provide the necessary feedback without using a mechanical position sensor. An important advantage of this approach is the fact that optimal performance can be obtained without changing the control architecture for the stepping motor user. The latter means that the user can still control the position by sending and counting step command pulses while the adaptive PI control introduced in this paper continuously optimises the current level. Although, the current - load angle dynamics are highly non-linear, an adaptive PI controller with the settling time of the current reduction as design constraint is worth considering. Especially because few tuning parameters and no mechanical parameters are required compared to Derammelaere et al. (2017).

2. LOAD ANGLE

The equation describing the electromagnetic motor torque is essential to have the necessary understanding in the stepping motor drive principle. The electromagnetic motor torque can be quantified based on the interaction between the stator flux linkage space vector $\mathbf{\Psi}_s$ and the stator current space vector $\mathbf{i}_s$ (Lin and Zheng (2006)).

$$\mathbf{T}_{\text{em}} = \mathbf{\Psi}_s \times \mathbf{i}_s \tag{1}$$

By neglecting saturation and splitting up the stator flux linkage to a dq-reference frame (Fig. 2) which is fixed to the rotor flux, the electromagnetic torque can be written as with L_d and L_q respectively the direct and quadrature inductance:

$$\mathbf{T}_{\text{em}} = (\mathbf{\Psi}_r + \mathbf{i}_d.L_d + \mathbf{i}_q.L_q) \times \mathbf{i}_s \tag{2}$$

Elaboration of the vector products leads to an equation describing the electromagnetic torque as a function of i_s and the load angle δ, defined as the angle between $\mathbf{i}_s$ and the rotor flux $\mathbf{\Psi}_r$ (Fig. 2):

$$T_{\text{em}} = \psi_r.i_s.\sin(\delta) + \frac{L_d - L_q}{2}.i_s^2.\sin(2\delta) \tag{3}$$

The first term in (3) describes the torque generated by the interaction between the permanent magnet rotor flux $\mathbf{\Psi}_r$ and the stator current $\mathbf{i}_s$. This term depends on the sine of the load angle δ. Because of the multi-toothed rotor and stator construction of a hybrid stepping motor, the reluctance effect will increase the maximal electromagnetic torque. This reluctance effect is represented by the second term in (3) and varies sinusoidally with twice the load angle δ.

Fig. 2 shows the relationship between the motor torque T_{em} (reluctance torque excluded) and the load angle δ at constant current level i_s. If the current level is constant, the generated motor torque becomes bigger when the load angle increases. The higher the load angle the more energy efficient the machine is used. When the load angle exceeds the optimum $\frac{\pi}{2}$, the generated motor torque decreases and step loss occurs.

2.1 Load angle estimation

The load angle δ reflects the capability of the stepping motor system to follow the position setpoint and is therefore interesting to estimate. The load angle δ is the angle between the current vector $\mathbf{i}_s$ and the rotor flux vector $\mathbf{\Psi}_r$ and thus equals to β - θ (Fig. 2). Unless an encoder is used, the location θ of the rotor flux vector $\mathbf{\Psi}_r$ is unknown. Therefore to estimate the load angle, the back-EMF is considered. Based on Lenzs law the resultant back-EMF vector $\mathbf{e}_s$ induced in the stator windings by the rotor flux $\mathbf{\Psi}_r$ can be written as:

$$\mathbf{e}_s = C\frac{d\mathbf{\Psi}_r}{dt} \tag{4}$$

As a result, $\mathbf{e}_s$ leads $\mathbf{\Psi}_r$ by $\frac{\pi}{2}$. Therefore, the load angle can be redefined as:

$$\delta = \frac{\pi}{2} - (\angle\mathbf{e}_s - \angle\mathbf{i}_s) \tag{5}$$

In the previous equation, the location of the current and the back-EMF vectors $\angle\mathbf{i}_s$ and $\angle\mathbf{e}_s$ are unknown. Because the phase current can be measured easily, the problem of estimating the load angle is reduced to a problem of estimating the position of the back-EMF. The back-EMF can be estimated based on the voltage equation of the stator windings. There is no interaction between the two phases because they are perpendicular to each other and constant inductance is assumed (Bendjedia et al. (2012)). Therefore, the mutual inductance is neglected and the back-EMF can be written as:

$$e_s(t) = u_s(t) - R_s.i_s(t) - L_s\frac{di_s}{dt} \tag{6}$$

The derivative in eq. (6) will cause problems if the measured current contains noise. Determining the derivative of noisy signals would result in distorted estimations. Therefore Derammelaere et al. (2014a) suggests to write (6) in the frequency domain, where ω represents the signal pulsation:

$$E_{\mathrm{s}}(\mathrm{j}\omega) = U_{\mathrm{s}}(\mathrm{j}\omega) + R_{\mathrm{s}}I_{\mathrm{s}}(\mathrm{j}\omega) + \mathrm{j}\omega.L_{\mathrm{s}}I_{\mathrm{s}}(\mathrm{j}\omega) \tag{7}$$

According to this method, only electrical parameters such as the stator resistance and inductance and the complex representation of the easily measurable phase current and voltage are needed to estimate the load angle. In stepping motor applications the position and speed setpoints are determined by step command pulses sent by the user as long as no step loss occurs. This means that the position and speed are always known and consequently also the instantaneous signal pulsation ω is known. The complex components $U_{\mathrm{a}}(\mathrm{j}\omega)$ and $I_{\mathrm{a}}(\mathrm{j}\omega)$ of the measured voltage and current signals are determined via transformation of the signals from the time to the frequency domain. De Viaene et al. (2017) described an estimator based on Phase Locked Loop which is able to determine the complex components even during speed transients.

3. LOAD ANGLE CONTROLLER

The large majority of the stepping motors in industry are driven in open-loop using a full, half or micro-stepping algorithm. These algorithms impose a stator current vector $\mathbf{i}_{\mathrm{s}}$. In these typical stepping motor drives, the angular position of the stator current vector is determined by step command pulses. Many commercial stepping motor drives also allow to adjust the current vector amplitude, labelled i_s in Fig. 3. Based on i_s and the step command pulses sent by the user, the transformation to the two-phase system is made and the current controller injects the desired two-phase currents in the motor. By doing so, the position of the rotor can be controlled in open-loop. The advantage of this method is that the position of the rotor can be directly imposed. The disadvantage is that the position of the permanent rotor flux $\mathbf{\Psi}_{\mathrm{r}}$ has not been taken into account to inject the two phase currents in order to achieve optimal torque generation.

In Derammelaere et al. (2017), a closed-loop load angle controller is suggested which is complementary to the typical stepping motor drives. This PI controller only adjusts the current level to obtain the desired load angle to reach the optimal torque/current ratio. In other words, the controller determines the amplitude of the stator current vector while the position of this vector is determined by step commands. The current is reduced from the nominal level to the minimum current necessary to drive the motor at a specific speed and load torque setpoint. Information of the load angle is used to control a stepping motor in an energy-efficient way. This approach is challenging as the current - load angle dynamics are highly non-linear. Therefore, Derammelaere et al. (2017) presented a linearized model. In this way, the linear theory in s-domain can be used to tune a PI load angle controller. As an outcome, both K_{p} and T_{i} depend on the mechanical damping b, the inertia J, the torque constant C_{T}, the setpoint load angle δ^*, the damping factor ζ and the load angle $\delta_{i_{\max}}$ at maximum current level. The last one is easy to determine because Derammelaere et al. (2017) proposed to start the stepper motor operation at maximum current I_{max} for every new speed setpoint. The damping factor ζ is used as design parameter of the closed loop system. The settings of K_{p} and T_{i} are visualized in Table 1.

Table 1. Settings of the PI controller based on linearized dynamics

K_{p}	$-\dfrac{J.\left(\dfrac{-b-\sqrt{b^2-4JC_T.I_{\max}.\dfrac{\sin(\delta_{i_{\max}})}{\tan(\delta^*)}}}{2.C_T.I_{\max}.\dfrac{\sin(\delta_{i_{\max}})}{\tan(\delta^*)}}\right)^2}{4.\zeta^2.C_T.\sin(\delta^*)}$
T_{i}	$\dfrac{2.C_T.I_{\max}.\dfrac{\sin(\delta_{i_{\max}})}{\tan(\delta^*)}}{-b+\sqrt{b^2-4JC_T.I_{\max}.\dfrac{\sin(\delta_{i_{\max}})}{\tan(\delta^*)}}}$

The time responses of the load angle controller strongly depend on the setting of the damping factor ζ but the main drawback of the method described in Derammelaere et al. (2017) is that the system parameters need to be known. Therefore in this paper, a simpler design procedure of the PI settings requiring less settings of the user is proposed. A novel procedure for finding the PI settings based on the current reduction time is described in the next section.

3.1 Adaptive PI controller based on current reduction time

The typical PI controller can be written by the following transfert fuction:

$$K_p\left(\frac{s+1/T_i}{s}\right) \tag{8}$$

The I-action will integrate the difference between the setpoint δ^* and the estimated load angle $\hat{\delta}$. In this way, the integrator can reduce the current from maximum current level $I_{\max}$ to the current level I_{required} at which the setpoint load angle is achieved.

$$\frac{K_p}{T_i}\int(\delta^*-\hat{\delta})\,\mathrm{d}t = I_{\max} - I_{\mathrm{required}} \tag{9}$$

To determine the parameters K_p en T_i in a simple way, the assumption is made that the PI controller will reduce the current level in such a way that the error will decrease linearly as illustrated in Fig. 4. The linearly increase of the load angle δ in Fig. 5, which shows the controlled current

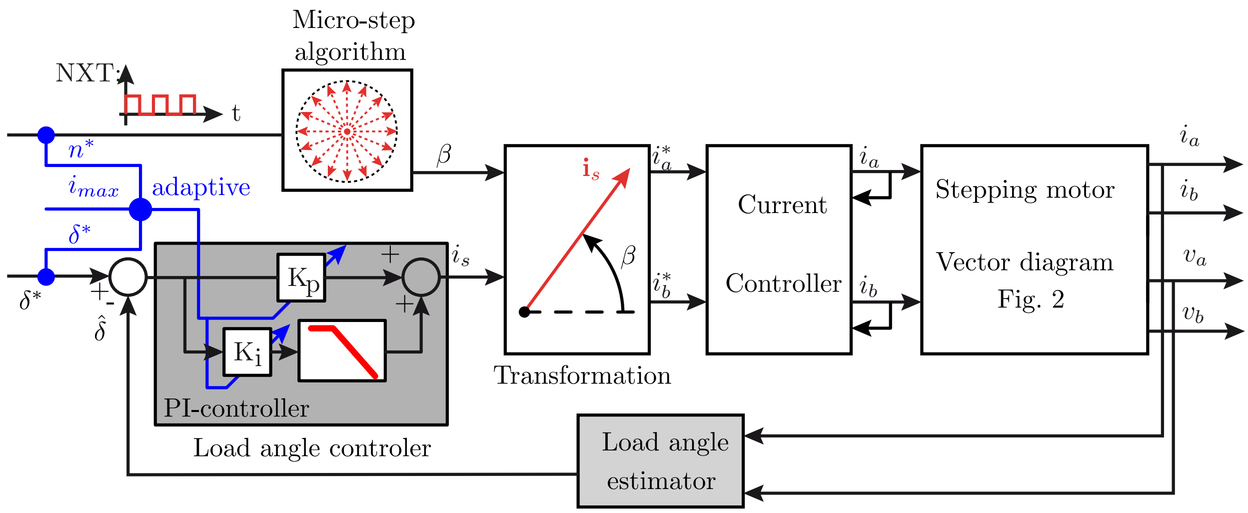

Fig. 3. Typical stepper motor drive principle where current vector $\mathbf{i}_s$ position β is determined by step command NXT pulses sent by the user extended with adaptive PI load angle controller

reduction and load angle optimisation using PI settings based on linearized process dynamics (Table 1), indicates that this assumption is justifiable.

The time required to reduce the current level $I_{\max}$ to I_{required} is defined as the design constraint T_a. Because of the simplified representation of the error reduction $\delta^* - \hat{\delta}$ as illustated in Fig. 4, eq (9) can be written as the area indicated by the triangle.

$$\frac{K_p}{T_i}\frac{1}{2}\left(\delta^* - \delta_{i_{\max}}\right)T_a = I_{\max} - I_{\text{required}} \qquad (10)$$

In this way, the ratio of K_p over T_i is given by:

$$\frac{K_p}{T_i} = \frac{I_{\max} - I_{\text{required}}}{0,5\left(\delta^* - \delta_{i_{\max}}\right)T_a} \qquad (11)$$

The time T_a required to reduce the current to the required current level is the design constraint. The maximum current level $I_{\max}$, the setpoint load angle δ^* are known. The load angle $\delta_{i_{\max}}$ at maximum current level can be measured. The current level I_{required}, at which the setpoint load angle is achieved, has to be calculated. For this, the motor operation is considered at steady-state at which the motor is loaded with a certain torque T_{load} at a certain speed ω.

To calculate I_{required}, the linear torque-load angle relation of (3) is used:

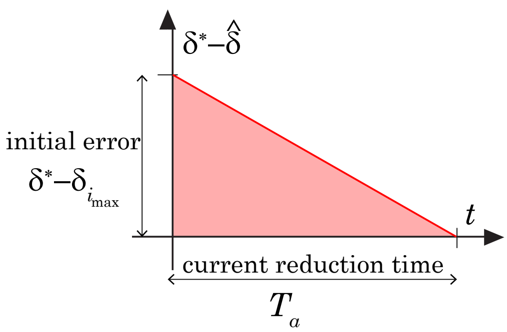

Fig. 4. Expected linearly decrease of the error between δ^* and $\hat{\delta}$

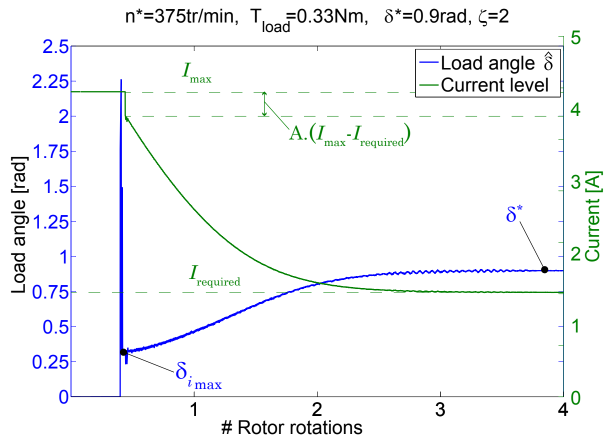

Fig. 5. Controlled current reduction and load angle optimization at specific operation point and PI settings based on linearized process dynamics

$$C_T.i.\delta = b\omega + T_{\text{last}} \qquad (12)$$

The product of $\delta_{i_{\max}}$ and $I_{\max}$ results in a ratio which contains information about the load and the machine:

$$I_{\max}.\delta_{i_{\max}} = \frac{b.\omega + T_{\text{last}}}{C_T} \qquad (13)$$

The left-hand part of (13) is constant for a given load and speed, so the required current I_{required} at the setpoint load angle δ^* can be determined by equating the right-hand sights of (13) and (14).

$$I_{\text{required}}.\delta^* = \frac{b.\omega + T_{\text{last}}}{C_T} \qquad (14)$$

In this way, (11) can be rewritten as:

$$\frac{K_p}{T_i} = \frac{I_{\max} - \left(I_{\max}.\frac{\delta_{i_{\max}}}{\delta^*}\right)}{0,5\left(\delta^* - \delta_{i_{\max}}\right)T_a} \qquad (15)$$

Based on the linear representation of the torque-load angle relation (12), this can simplified:

$$\frac{K_p}{T_i} = \frac{I_{\max}}{0,5.\delta^*.T_a} \tag{16}$$

Simulation results obtained by tuning the PI controller based on the linearized dynamics (Table 1) are useful to find the setting of the current reduction time T_a. Table 2 summarizes the time response and the required rotor rotations rotor to control the load angle at different operating points.

Table 2. Time response PI controller based on linearized dynamics

Speed [rpm]:	75	75	375	500	500
Load torque [Nm]	0	0,49	0,33	0	0,49
$\delta_{i_{\max}}$ [rad]	0,06	0,24	0,31	0,55	0,83
δ^* [rad]	0,20	0,67	0,80	1,05	1,18
# rotor rotations	0,5	0,4	1,9	3,1	3,3
Settling time [s]	0,4	0,32	0,30	0,37	0,4
Part A of K_p on reduction	0,1	0,11	0,1	0,21	0,38

These results show that a minimum of 0,4 rotor rotations is required to control the load angle. This limit is assumed as design constraint whereby the current reduction time T_a can be written with n the motor speed as:

$$T_a = 0,4\frac{60}{n} \tag{17}$$

On the one hand, the I-action is responsible for the current reduction. On the other hand, the P-action is responsible to react on sudden changes in the load angle so that the current level can be adjusted sufficiently fast. When the P-action has to reduce instantaneously the maximum current level $I_{\max}$ until the required current level I_{required} is reached, K_p can be written as:

$$K_p = \frac{I_{\max} - I_{\text{required}}}{\delta^* - \delta_{i_{\max}}} \tag{18}$$

Also this equation can be simplified in an analogue way as (16) is simplified:

$$K_p = \frac{I_{\max}}{\delta^*} \tag{19}$$

It is not realistic to expect that the P-action completely guarantees the reduction of the current level taking in mind the dynamics of the system and the estimator. Therefore, the parameter A is presented which indicates the part of the proportional gain K_p leading to the current reduction. The factor A is also illustrated in Fig. 5. K_p can then be calculated as follows:

$$K_p = A.\frac{I_{\max}}{\delta^*} \tag{20}$$

A too large A leads to a too large gain factor K_p which results in an unstable system. In order to find out how large the fraction A of the P-action may be, the fraction is calculated when the previous settings (Table 1) based on the linearized process dynamics are used. Table 2 shows that the maximum fraction of the P-action on the current reduction may not exceed 10 % if a reduction period of 0,4 rotor rotations is desired. Table 3 summarizes the settings of the PI controller based on the current reduction time which are much simpler than settings based on the linearized dynamics summarized in Table 1. The damping b, the inertia J, the torque constant C_{T}, the load angle $\delta_{i_{\max}}$ at maximum current level and the damping factor ζ are no longer required. In this paper, an adaptive PI controller is obtained (Fig. 3) which adjusts K_p and T_i in function of the motor speed n^*, the setpoint load angle δ^* and the maximum current level $I_{\max}$.

Table 3. Settings of the PI controller based on the current reduction time

K_p	$-A.\frac{I_{\max}}{\delta^*}$ for $A_{\text{opt}} = 0,1$
T_i	$-0,5.A.T_a$ for $T_{a_{\text{opt}}} = 0,4.n$

4. MEASUREMENTS

The presented design procedure is used to measure the load angle optimisation at unload, half and fully loaded motor operation. In this way, Figures 6, 7 and 8 are obtained.

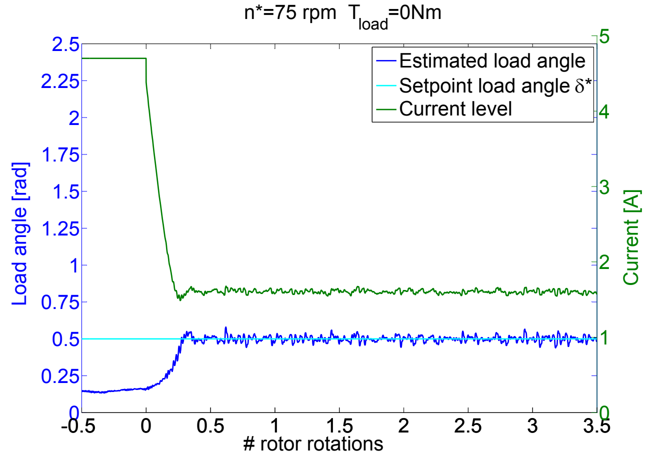

Fig. 6. Controlled current reduction and load angle optimisation at low speed and unloaded

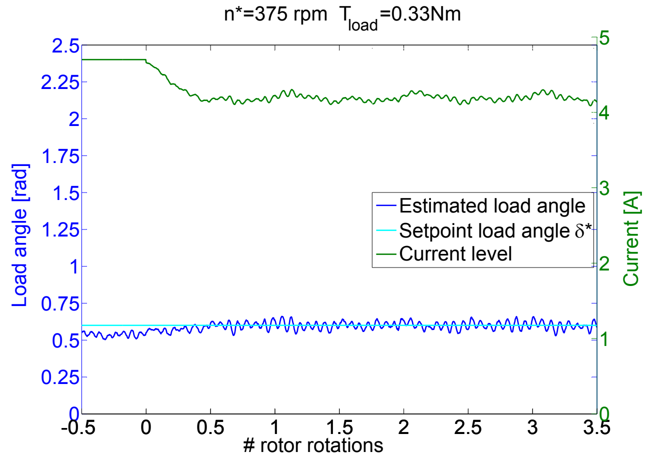

Fig. 7. Controlled current reduction and load angle optimisation at half loaded motor operation

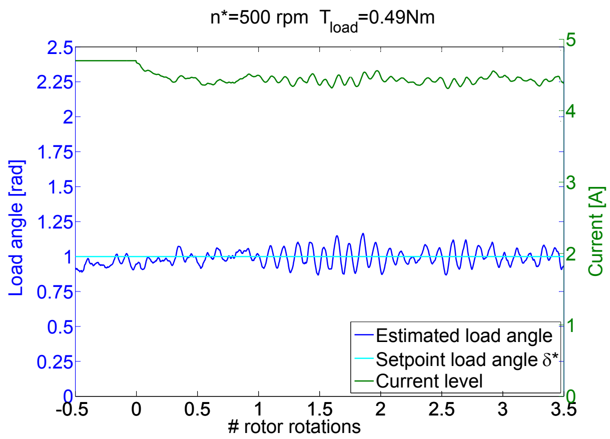

Fig. 8. Controlled current reduction and load angle optimisation at high speed and fully loaded motor

By presenting the Figures as a function of the rotor rotations, it is clear that the current reduction in Figs. 6-8 always follows the same pattern. A current reduction of 65, 13 and 6 % of the nominal current level is achieved respectively at unloaded, half and fully loaded motor operation if I_{max} is 2.4 A. The time response of the controller satisfies the design constraint because the load angle reaches the setpoint load angle each time after approximately 0.4 rotor revolutions.

5. CONCLUSION

The bulk of the stepping motor applications are driven in open loop, with maximum current to avoid step loss. These drive strategies result in very poor energy efficiency. In order to use stepping motors in a more optimal way, closed-loop control is needed. A previously described sensorless load angle estimation algorithm, solely based on voltage and current measurements, is used to provide the necessary feedback without using a mechanical position sensor. The load angle determines the torque/current ratio of the stepper motor drive.

The contribution of this paper is in the design of a control which adapts the stepping motor current level based on the estimated load angle. An important advantage is the fact that optimal performance can be obtained without changing the control architecture for the stepping motor user. The design method to achieve the optimal PI settings does not use the transfer function which describes the process dynamics. A novel more simple procedure for finding the PI settings is described. An adaptive PI controller is obtained which adjusts K_p and T_i in function of the motor speed, the setpoint load angle and the maximum current level. Measurements prove that this method satisfies the design constraint. The load angle setpoint is obtained in 0,4 rotor rotations at different motor operating points. The controller is able to reduce the current level up to 65 % of the nominal current level at low speed and no load.

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