Reduction of sensing cycle duration in self-sensing active magnetic bearings using voltage asymptote prediction

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Abstract
This work aims at reducing the required time to obtain a self-sensed position signal in active magnetic bearings. Specifically, the self-sensing concept of Direct Digital Inductance Estimation should be enhanced by a prediction algorithm. Ordinarily, the current slope as a result of a coil voltage step is measured via a transformer, whereby a single measurement of the transformer's secondary coil voltage is sampled after the voltage has converged towards its asymptote value. This paper proposes to predict the asymptote by fitting an exponential curve to multiple measurements sampled before the transformer voltage has converged. While a lower accuracy of the self-sensed position signal is to be expected with this approach, the shorter duration of the sensing phase allows for more of the amplifier voltage to be used to actuate the magnetic bearings. This in turn can be useful in situations where the magnetic bearings reach their dynamic limits in supplying the requested coil currents. A single-axis magnetic bearing test bench driven by a switching amplifier in differential current control is used to evaluate this approach at different flotor positions. Statistical evaluations of the position signals obtained by the conventional and by the prediction method are compared. Finally, an outlook is given on the feasibility of this approach and possible use cases.

Keywords : Active Magnetic Bearings, Self-Sensing, Position Estimation, Direct Digital Inductance Estimation, Current Slope, Transformers, Asymptote Estimation, Exponential Curve Fitting

1. Introduction

Dissemination of magnetic bearings is mostly limited by their complexity, size, and cost (Maslen and Schweitzer, 2009; Richter, 2020). Thus, self-sensing magnetic bearings are researched, which alleviate all these aspects by eliminating the dedicated position sensors. Instead, the rotor position is estimated only from the voltage and current signals available in the amplifier.

Various methods with different upsides and downsides have been established to achieve this goal (Maslen and Schweitzer, 2009; Schammass et al., 2005). A promising approach is Direct Digital Inductance Estimation, which leverages the bearings' current ripple as a response to the pulse width modulation (PWM) inherent to modern switching amplifiers, first shown in Glück et al. (2011). Richter (2020) has presented a method that determines the current slope with transformers that are integrated into the bearing amplifier. This method is the basis for this paper and explained in the following section.

2. Direct Digital Inductance Position Estimation

The self-sensing method described in Richter (2020) and used in this paper separates each PWM cycle into a sensing and an acting phase (Fig. 1). During the sensing phase, two voltage excitations are applied after another to each side of the opposing bearings of an axis. These excitations are not meant to control the bearing force but to provide a dependable sensing input, thus voltage high and low time are set equal to keep the coil currents i_1 and i_2 roughly constant. The acting phase uses the remaining time to alter the coil currents according to the output of the current controller. Opposing magnets are actuated asymmetrically and each of them is attached to a low-inductivity transformer (Fig. 2). The secondary-side transformer voltages $u_{TS,1}$ and $u_{TS,2}$ are measured with high impedance and used to evaluate the rotor position x.

To establish a mathematical connection between the transformer voltages and the rotor position, equations are formulated for the transformers (1) and bearings (2), exemplarily stated here for transformer 1 and bearing 1:

$$u_{TS,1} = \frac{\partial \Psi_{TS,1}}{\partial i_1}\frac{di_1}{dt} + \frac{\partial \Psi_{TS,1}}{\partial i_{TS,1}}\frac{di_{TS,1}}{dt} = N_{TS}\frac{\partial \phi_{TS,1}}{\partial i_1}\frac{di_1}{dt} \quad (1)$$

$$u_1 = R_1 i_1 + \frac{d\Psi_1}{dt} + \frac{d\Psi_{TP,1}}{dt} = R_1 i_1 + \frac{\partial \Psi_1}{\partial i_1}\frac{di_1}{dt} + \frac{\partial \Psi_1}{\partial s_1}\frac{ds_1}{dt} + \frac{d\Psi_{TP,1}}{dt} \quad (2)$$

Here, ψ represent the flux linkages, R the electric winding resistances, and N the winding numbers (Fig. 2). Equation (1)

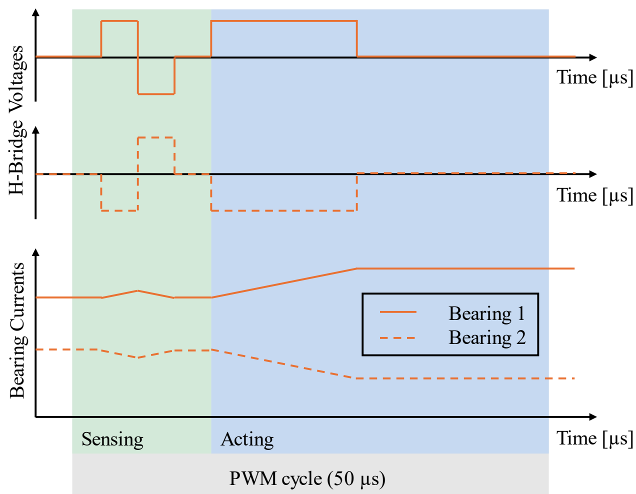

Fig. 1 Qualitative PWM cycle for two opposing magnets, divided into a sensing and an acting phase. The sensing phase remains equal between cycles, whereas the acting signal can vary in length and sign.

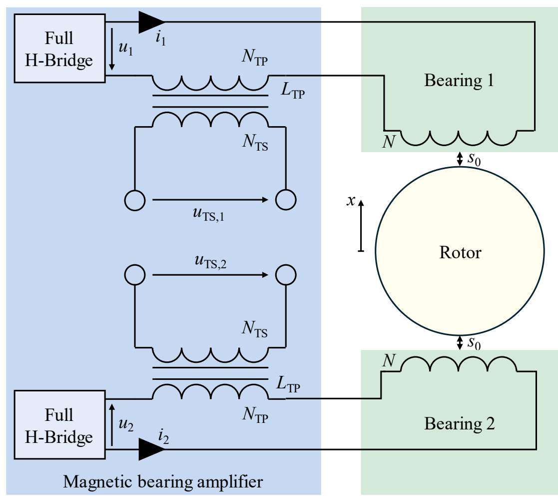

Fig. 2 System structure of two opposing magnets. Each magnet is driven by a single full H-bridge. The current slopes are measured by amplifiers.

exploits that the secondary transformer winding is open or rather measured with high impedance, so that its change in current is near zero. Equations (1) and (2) are combined under the assumptions of linearly acting transformers to obtain an equation for the transformers' secondary winding voltage depending on the flux linkages of the magnetic bearing. If the sum of the transformer voltages of opposing bearings is calculated and averaged over the high and the low flank of the sensing phase, a formula is obtained that shows linear dependence on the rotor position, L_0 representing the nominal bearing inductance, l the iron length:

$$u_{TM} = -\frac{N_{TS}}{N_{TP}}\frac{2L_{TP}}{L_0}\left(\frac{u}{1+\frac{2L_{TP}}{L_0}}\right)\left(\frac{2x}{2s_0+\frac{l}{\mu_r}}\right) \quad (3)$$

Previous experiments have shown that (3) can estimate the rotor position with an accuracy of only a few micrometers and with a bandwidth as high as the PWM frequency (Richter, 2020). Additionally, it was suitable for use in a control loop and did not negatively impact overall system sensitivity. A drawback of this method however is that the time allocated for the sensing pulses takes away a significant portion of the PWM cycle, thus limiting the effectiveness of the bearing. Addressing this problem is the aim of this paper and a solution approach is proposed in the following section.

3. Voltage Asymptote Prediction

Two reasons prevent reduction of the sensing pulse width: First, parasitic capacities lead to high current ripples directly after switching, which only gradually decrease over time and are difficult or impossible to model accurately. Thus, the first portion of the current response cannot be used for measuring (Phase 1 in Fig. 3). Second, the transformer voltages reach their end values asymptotically, in part because they are filtered analog to reduce switching ripples (Phase 2 in Fig. 3). Thus, depending on the bearing parameters, a lot of time can pass until the final transformer voltages can be measured during Phase 3 of Fig. 3.

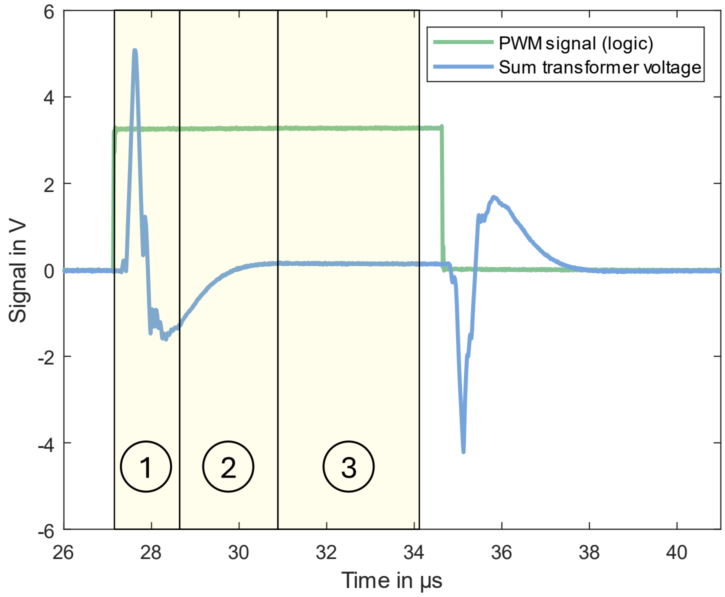

Fig. 3 Exemplary response of the sum transformer voltage $u_{TS} = u_{TS,1} + u_{TS,2}$ to a PWM pulse. The figure only shows the logic side of the pulse (3.3 V) as input to the gate driver. The pulse acting on the coil is 48 V. Phase 1: Switching ripple. Phase 2: Asymptotic progression to be used for prediction. Phase 3: Constant progression used to sample the true asymptote.

A possibility to expedite the self-sensing process is to estimate the asymptote of the transformer voltages based on high-frequency measurements while the voltages are still changing. The first idea would be to fit this voltage progression with an exponential equation in the form of (4) and extrapolate in time. a, b, and c are the parameters to be fitted in this case.

$$u(t) = a \cdot (1 - e^{-bt}) + c \tag{4}$$

However, a linear solver should be used due to computational restraints, thus b cannot be optimized. Instead, a differential equation in the form of a first order delay is applied in discretized form and set up as a matrix equation over k samples, see Eq. (5). Amongst others, Glück et al. (2011) have implemented a least-mean-squares fit to estimate bearing inductances from current slopes and thus show that this is possible in principle even with the demand for real-time results.

$$\tau \cdot \dot{u}(t) + u(t) = u_{\mathrm{asy}} \tag{5}$$

Here, $u(t)$ denotes the transformer voltage, which reaches its end value u_{asy}. τ is the exponential time constant to be fitted to the measurements. In discretized form, this becomes

$$\tau \cdot (u_k - u_{k-1}) \cdot f_{\mathrm{s}} - u_{\mathrm{asy}} = -u_k \tag{6}$$

for each time step k with sample frequency f_{s}. If n samples are measured, the following system of equations is obtained:

$$\begin{bmatrix} u_2 - u_1 & -1 \\ u_3 - u_2 & -1 \\ \vdots & \vdots \\ u_n - u_{n-1} & -1 \end{bmatrix} \cdot \begin{bmatrix} \tau \cdot f_{\mathrm{s}} \\ u_{\mathrm{asy}} \end{bmatrix} = \begin{bmatrix} -u_2 \\ -u_3 \\ \vdots \\ -u_n \end{bmatrix} \tag{7}$$

This system of equations is typically overdetermined and can be solved with the pseudo-inverse. While this approach works on ideal data, the substantial switching ripples lead to high estimation errors. To make the optimization more robust to noise, the trapezoidal integral of the voltage over k steps with sample time t_{s} is fitted instead of the absolute datapoints. (5) is integrated with integration constant γ and expressed in discretized form, again assuming n samples obtained at a frequency f_{s}:

$$\tau \cdot f_{\mathrm{s}} \cdot u_k + \sum_{i=1}^{k-1} u_i + \frac{u_k - u_1}{2} = u_{\mathrm{asy}} \cdot t_k \cdot f_{\mathrm{s}} + \gamma \cdot f_{\mathrm{s}} \tag{8}$$

Here, t_k is the time difference between the first and the k-th sample. Analogous to (7), this equation is evaluated for every sample k leading to the following system of equations:

$$\begin{bmatrix} u_2 & -t_2 & -1 \\ u_3 & -t_3 & -1 \\ \vdots & \vdots & \vdots \\ u_n & -t_n & -1 \end{bmatrix} \cdot \begin{bmatrix} \tau \\ u_{\mathrm{asy}} \\ \gamma \end{bmatrix} \cdot f_{\mathrm{s}} = \begin{bmatrix} u_1 - 0.5 \cdot (u_2 - u_1) \\ (u_1 + u_2) - 0.5 \cdot (u_3 - u_1) \\ \vdots \\ \sum_{i=1}^{n-1} u_i - 0.5 \cdot (u_n - u_1) \end{bmatrix} \tag{9}$$

It is solved by means of the pseudo inverse and the obtained value u_{asy} directly represents the asymptote of the transformer voltage.

4. Test Setup

The test bench is shown in Fig. 4 and Fig. 5. A flotor is suspended by wires with almost no friction. It is made from aluminum with the exception of laminated steel blocks to route the flux on each bearing side. The stator is located in between these steel blocks and holds the coils and the laminated steel yokes. Keyence LK-G32 laser distance sensors are used to determine the reference position for the bearing controller, which is calculated from their measurement difference to compensate for symmetric deformations of the flotor and stator due to bias currents and to improve accuracy.

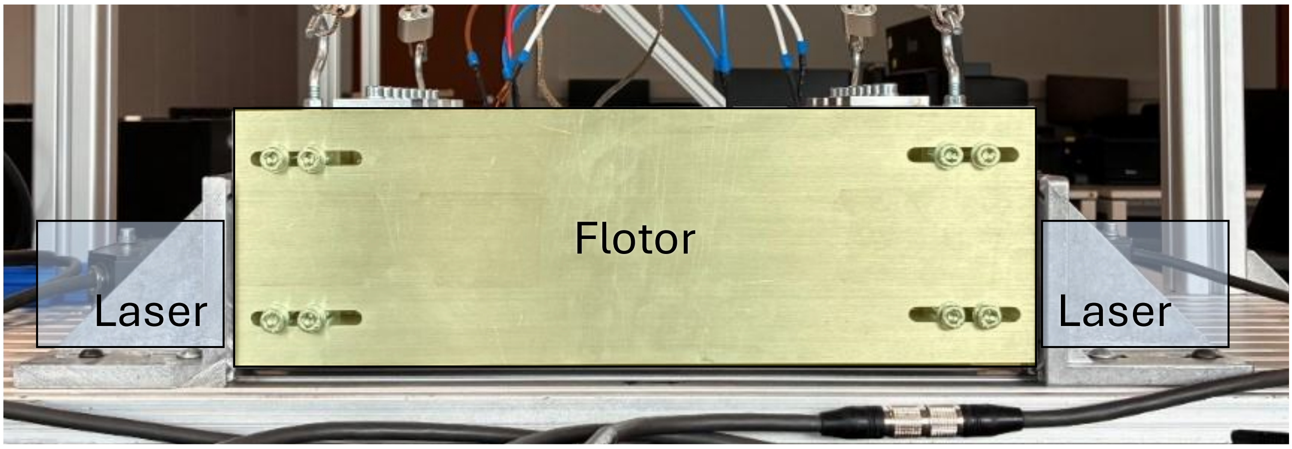

Fig. 4 Side view of the test bench. The flotor is suspended by hooks almost without friction and can move left and right. The lasers measure the flotor position differentially.

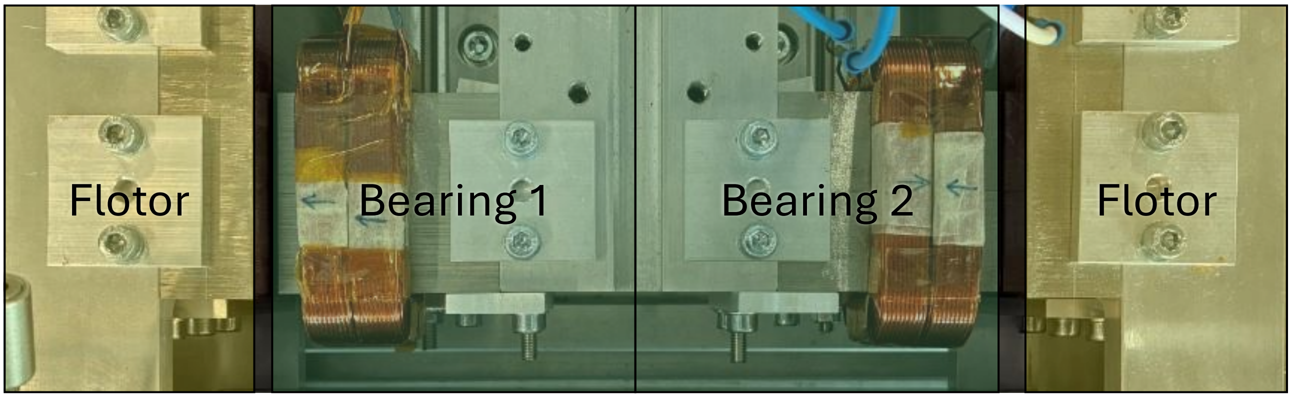

Fig. 5 Top view of the test bench. The flotor can move left and right. Each of the bearings is actuated by two winding packages connected in series so they act as one coil per side.

Each side of the bearing is actuated by a single coil in differential current control with a switching amplifier. The amplifier operates at 48 V and supplies a bias current of 1.5 A. The control loop operates at 20 kHz based on high-bandwidth current sensors in the amplifier and the laser sensors as position reference. While Fig. 2 shows the transformers as separated units to clearly show the relevant parameters, the test bench generates the sum of the transformer voltages analog as a summing transformer.

5. Measurements and Results

Within the scope of this paper, measurements are conducted with the flotor stationary. The reference position for the magnetic bearing control is swept in varying increments from –100 µm to +100 µm. After each position change, a short waiting time is passed before measuring so that any residual oscillations caused by the set position step can subside. Then, 20 PWM cycles are measured with the oscilloscope and saved for evaluation. The u_{TM} values based on asymptote prediction are generated by sampling 13 values of the summing transformer at 20 MHz and then solving Eq. (9) for u_{asy} before averaging over high and low flank of the sensing cycle. The resulting curves are shown in Fig. 6, which compares the estimated positions based on directly sampled asymptotes to those predicted with Eq. (9). Each point in Fig. 6 represents an average over 20 PWM cycles.

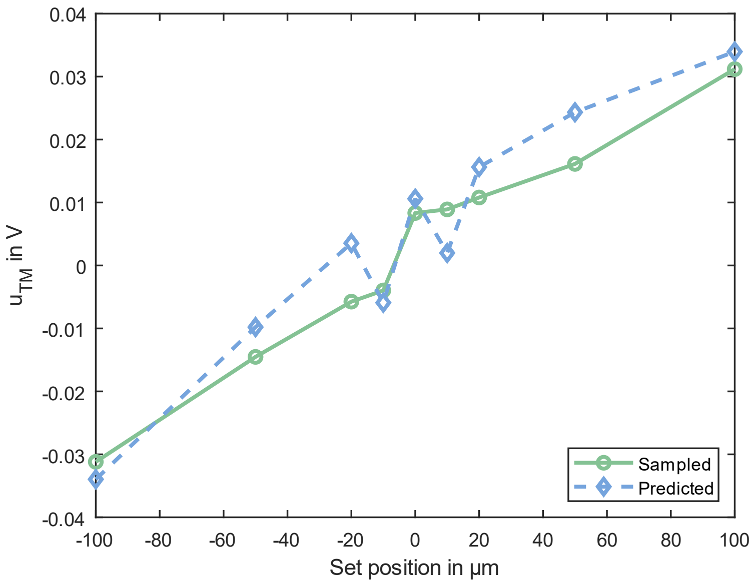

Fig. 6 Offset-corrected self-sensing signal u_{TM} over flotor set position. Each point is averaged over 20 measurements. Ideally, this signal is directly proportional to the flotor position. Errors are due to measurement noise, nonlinearities, and asymmetries in the bearing.

It can be seen that even the position estimation based on the sampled data is not ideal, which is attributed to measurement noise, nonlinearities, and asymmetries in the bearing. It is deemed good enough though to be used as feedback for the magnetic bearing control.

The curve for the predicted asymptotes is visibly worse, as expected to a degree, especially around the center position of the flotor, where errors in the asymptote prediction influence the position estimation to a greater extent. The main problem currently preventing the use of this signal as control input is that the standard deviation of u_{TM} with asymptote prediction is far higher than with direct measurement of the asymptote.

An advantage of the prediction is that the goal of reducing the sensing time could be achieved. Each sensing PWM pulse could be reduced by 2.2 µs, leading to a total increase of the acting phase of about 12 %. It is expected that larger magnetic bearings will benefit more.

6. Conclusion and Outlook

In this paper, an approach to reduce the time required for sensing the position in an active magnetic bearing with Direct Digital Inductance Estimation was tested. It was observed that a small reduction was possible for the specific bearing configuration of this test rig. For larger bearings, the time saving is expected to be more significant. However, while the mean values of the asymptote prediction are relatively close to the directly measured asymptotes of the sum transformer voltage, the standard deviation is likely too high for this exact method to be applied as an input signal for a controller for normal operating conditions around the center position.

While there is an argument that the prediction could be turned on in cases where the rotor is strongly eccentric and the bearing needs to deliver maximum voltage, i.e. the exact position is less relevant than the required direction of bearing force, this edge case likely does not justify the use of more expensive Analog-to-Digital-Converters.

Further research could be dedicated to more elaborate prediction methods, such as neural networks. If the sample frequency and noise requirements for the prediction could be reduced and its accuracy increased, the approach of asymptote prediction could indeed reduce sensing time.

10. References

Schweitzer, G. and Maslen, E., Magnetic Bearings: Theory, Design, and Application to Rotating Machinery (2009), Springer, New York.

Schammass, A., Herzog, R., Buhler, P., Bleuler, H., New results for self-sensing active magnetic bearings using modulation approach, Control Syst. Technol. IEEE Trans., vol. 13, no. 4 (2005), pp. 509–516.

Glück, T., Kemmetmüller, W., Tump, C., Kugi, A., A novel robust position estimator for self-sensing magnetic levitation systems based on least squares identification, Control Eng. Pr., vol. 19, no. 2 (2011), pp. 146– 157.

Richter, M., Erhöhung der Praxistauglichkeit selbst-sensierender aktiver Magnetlager (2020), Shaker Verlag, Düren.