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EXPERIMENTAL RESULTS ON SELF-SENSING AMB USING A THREE-STATE PWM AMPLIFIER

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ABSTRACT

This paper presents a self-sensing AMB using a three-state PWM amplifier. The amplifier is configured in such a way that the voltage ripple never becomes zero. Since the proposed approach is based on demodulating the current and the voltage, the amplifier setup guarantees that the position can be always estimated. Experimental results from a radial magnetic bearing are presented. The quality of the estimation signal is evaluated in terms of noise and bandwidth.

INTRODUCTION

The idea of the self-sensing AMB is to eliminate the position sensors and to estimate the position by measuring the current in the actuator coils. This represents a significant cost reduction. The additional advantages are a reduced number of components, an increased system reliability and a shorter rotor (i.e. higher natural frequencies).

Many different configurations of self-sensing AMB have been proposed in literature [1-5]. These configurations can be divided into two classes: linear control and modulation approaches. Morse, et al [6] have shown that the linear observer approach lacks robustness compared to the sensor case. On the other hand, Montie and Maslen [5] have shown that the system robustness is improved using a modulation approach.

This paper presents a self-sensing AMB based on a modulation approach using a three-state PWM amplifier. The amplifier is configured in such way that the voltage ripple never becomes zero. Since the proposed method is based on the voltage and current ripple measurements, the estimation of position can be guaranteed at any time.

BEARING COIL MODEL

Figure 1 shows the typical configuration of a magnetic bearing system with opposite coils. The coils are considered identical and there is no magnetic coupling between them. It is assumed that there is no flux leakage, no eddy current losses, and the behavior of the magnetic material is linear (constant permeability and no saturation). The air gaps can be written as $s_1 = x_0 + x$ and $s_2 = x_0 - x$, where x is the displacement with respect to the nominal air gap x_0.

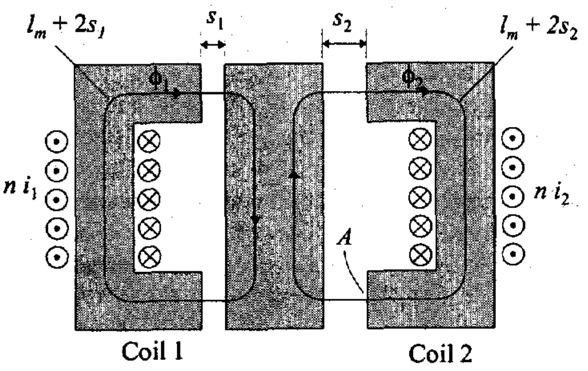

FIGURE 1: Simplified coil model.

Under the above-mentioned assumption, the relation among the current i_k, voltage u_k, and gap s_k of coil k (k=1, 2) is given by:

$$u_k = K\frac{d}{dt}\left(\frac{i_k}{2s_k+\frac{l_m}{\mu_r}}\right)+Ri_k \qquad (1)$$

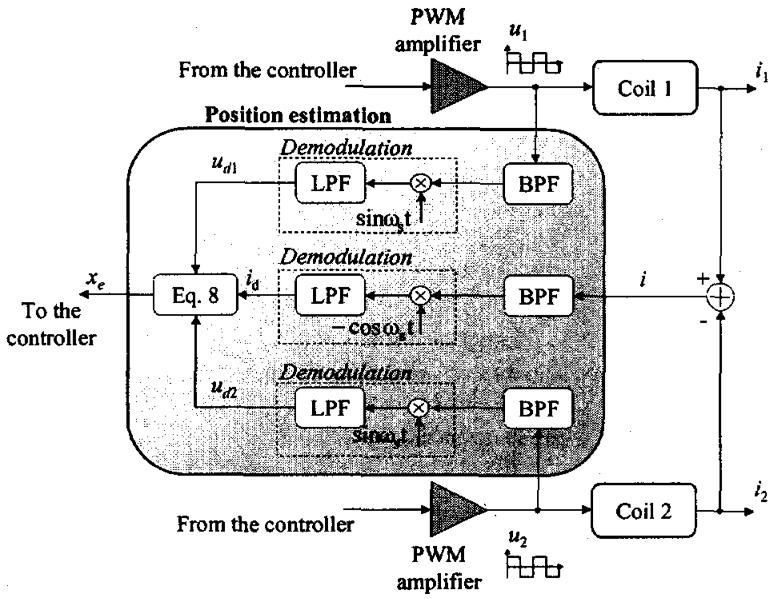

FIGURE 2: Estimation configuration using opposite coils

where $K=\mu_0 n^2 A$ is the magnetic bearing constant, with μ_r the relative permeability of the core material, μ_0 the magnetic permeability, and A the cross section of the iron core.

PWM SWITCHING AMPLIFIER

In the proposed self-sensing approach, the actuator coils are fed by a PWM switching amplifier. Figure 3 illustrates the operation principle of the three-state PWM amplifier used in this work.

The operation principle is similar to the amplifier presented in [7]. The voltage of each coil terminal is controlled separately by a pair of switches. The switches 1 and 2 control the voltage u_a and the switches 2 and 4 control the voltage u_b. The voltage u_a is switched between zero and the supply voltage u_z with a variable duty-cycle α. For this configuration, $\alpha=0$ represents $u_a=0$, and $\alpha=1$ means $u_a=u_z$. In the other terminal, the voltage u_b is switched with constant duty-cycle of 0.5. In this way, the control average voltage across the coil u (the difference u_a-u_b) can vary between $-u_z/2$ and $u_z/2$.

This type of PWM amplifier circuit presents an advantage for self-sensing technique compared to conventional two-state PWM amplifiers [8]. Since the voltage u_b has a constant duty-cycle and phase shift of 180 degrees with respect to u_a, the voltage ripple across the coil is never zero. This is true even at zero average control voltage. Since the proposed approach is based on the voltage and current ripple measurements, this amplifier set up guarantees that the position can be always estimated.

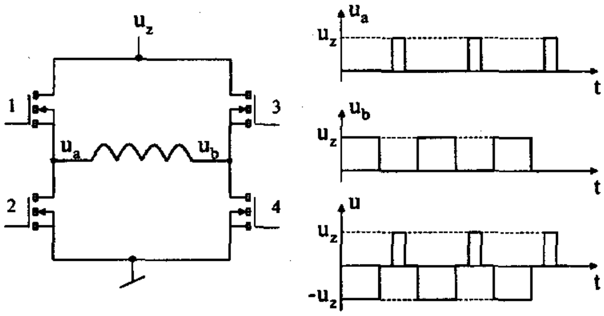

FIGURE 3: Three-state PWM switching amplifier

The voltage u across the coil can be approximated by Fourier series if the change of the duty-cycle is much slower than the switching frequency ω_s [8]. Considering only the first harmonic of voltage ripple, u can be expressed by equation (2):

$$u(t) \cong u_z(\alpha-0.5)+\frac{2u_z(1+\sin\pi\alpha)}{\pi}\sin\omega_s t \qquad (2)$$

POSITION ESTIMATION

Figure 2 shows the self-sensing configuration using opposite coils. The objective of using the information of opposite coils is to increase robustness with respect to noise and modeling errors. The difference of the currents $i=i_1-i_2$ is measured and filtered by a band-pass. The voltage is measured for each coil individually and also band-pass filtered. With this configuration, it is possible to obtain the information from opposite coils without measuring each current individually.

Assuming that for high frequencies $u_k >> Ri_k$ and both coils are identical, i_1 and i_2 can be written:

$$i_1 \cong \frac{1}{K}(2x + s_0)\int u_1 dt \qquad (3)$$

$$i_2 \cong \frac{1}{K}(-2x + s_0)\int u_2 dt \qquad (4)$$

where $s_0 = 2x_0 + l_m / \mu_r$. The cutoff frequency of the band-pass filter is selected such that only the first harmonic remains in the signal. In this way, the voltage u_k is written:

$$u_k = \hat{u}_k \sin \omega_s t \qquad (5)$$

where $\hat{u}_k = \frac{2u_z}{\pi}(1 + \sin \pi\alpha_k)$.

Using equations (3) and (4), the difference of the currents can be expressed:

$$i = \frac{1}{K}[2(v_1 + v_2)x + (v_1 - v_2)s_0] \qquad (6)$$

where $v_1 = \int u_1 dt$ and $v_2 = \int u_2 dt$.

Assuming that bandwidth of $\hat{u}_k$ is much smaller than the switching frequency ω_s, u_1 and u_2 can be substituted in equation (6) and the current difference i can be written as:

$$i \cong \frac{-1}{K\omega_s}[2(\hat{u}_1 + \hat{u}_2)x + (\hat{u}_1 - \hat{u}_2)s_0]\cos \omega_s t \qquad (7)$$

As shown in Figure 3, synchronous demodulation is used in order to obtain the real part of demodulated current i_d and demodulated voltages u_{d1} and u_{d2}. The synchronous demodulation consists of multiplying the signal by a sinusoidal signal with same phase. In this way, the estimated position x_e is obtained:

$$x_e = \frac{K\omega_s i_d - s_0(u_{d1} - u_{d2})}{2(u_{d1} + u_{d2})} \qquad (8)$$

As it was mentioned before, the PWM amplifier guarantees that u_{d1} and u_{d2} are never zero. This means that the division in equation (8) is well conditioned.

RESULTS

Experimental results of a radial bearing are presented in this section. The quality of the estimated position is discussed in terms of noise and dynamic performance.

Figure 4 shows the test rig used for the experimental measurements. The system is composed of two radial bearings, one axial bearing and a short rotor with first bending mode at 1.5 kHz. The reference sensors used on the test rig were eddy current type. The main characteristics of the system are the nominal air gap of 0.4 mm, and the maximum current of 1.2 A. The rotor has the length of 235 mm, and a weight of 2.78 kg.

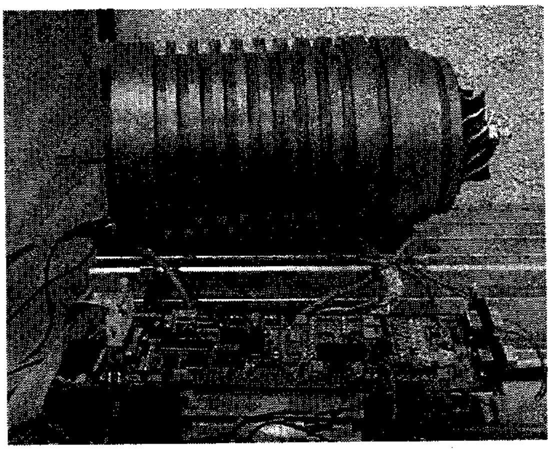

FIGURE 4: Test rig

The switching frequency of PWM amplifier was 20 kHz. To implement the position estimator, the filters and demodulators were built in analog circuit and equation (8) and the control algorithm was implemented in a DSP (digital signal processor).

The position estimator was applied to one radial axis (one degree of freedom). Measurements in time and frequency domain were realized in order to evaluate the estimator performance. These measurements were realized with the levitated rotor using the reference sensors. The estimated position was compared to the measured position obtained from the reference sensor.

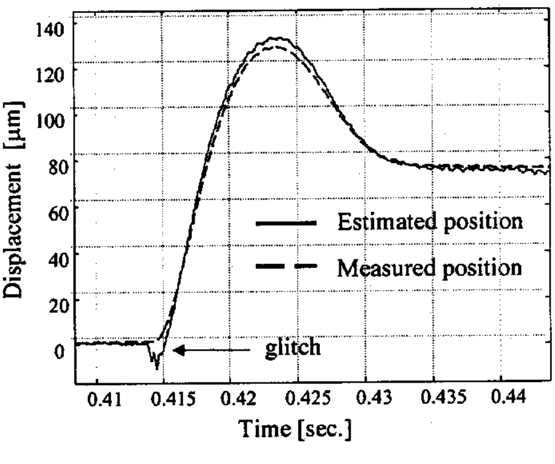

FIGURE 5: Step response of estimated and measured position.

Figure 5 shows the step response of the position. The estimated and measured positions are compared. The estimated signal presents a small error with respect to the measured signal. The largest error is observed in the beginning of the step response (glitch), but it

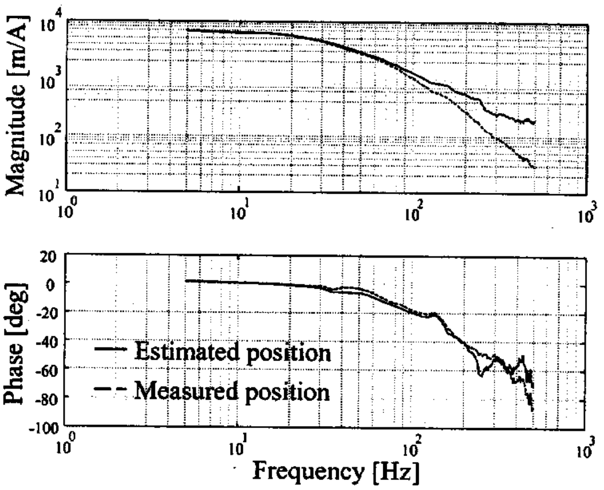

FIGURE 6: Frequency response of estimated position and the sensor position with respect to desired current.

remains lower than 5% of the nominal gap. The noise of the estimated position was relatively low. The measured peak-to-peak value was 3.7 μm, which represents 1.5% of the nominal gap. This value is similar to the noise of measured position.

Figure 6 shows the frequency response of the estimated position and the measured position with respect to the desired current. Since it is desired to have the estimated signal equal to the "true" position, ideally both frequency responses should be the same.

The figure shows that the response matches well for low frequencies. The phase difference between the curves remains small up to 500 Hz, but the difference of the amplitude becomes larger for higher frequencies. Additionally, further investigations have shown that the magnitude of the estimated signal depends on the excitation amplitude. Although not fully investigated, other measurements indicate that the estimated signal is sensitive to the low frequency component of the current. Despite this non-linear behavior, it was possible to levitate the rotor using the estimator.

CONCLUSIONS AND OUTLOOK

A self-sensing AMB was presented using a three-state PWM amplifier. Measurements in time and frequency domain were presented in order to evaluate the quality of estimated position.

It was possible to levitate one axis of an AMB without sensors with relatively low noise. The results show that estimated position has an excellent match with the "true" position and low noise for low frequencies. However, the quality of the estimation decreases for high frequencies.

To improve the dynamic performance, the next step is to compensate the influence of change in the low frequency component of the current. To achieve this, the influence of material non-linearity must be investigated in the future.

Although the position estimation was implemented mostly in hardware, the proposed self-sensing configuration has the potential to be implemented digitally. The next step in order to reduce the costs and hardware complexity is to implement the PWM amplifier on a DSP. In this case, the duty-cycle is generated by the DSP and there is no need to measure the voltages in the coil, only the current must be measured and demodulated.

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