Synchronous Three-Level PWM Power Amplifier for Active Magnetic Bearings

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Abstract: The availability of low cost and high performance power amplifiers is essential for a wide-spread application of active magnetic bearings (AMB). In this paper, a new switching control pattern, referred to as synchronous three-level PWM (Pulse Width Modulation) scheme, is recommended for switching power amplifiers used in digitally controlled AMBs. This PWM control pattern allows it to work at a low switching frequency with low current harmonics. At the same time, the disturbance response is improved.

Nomenclature

A	effective cross section area of air gap
B_0	bias magnetic flux density
F	magnetic force
f_t	normalized magnetic force
f_z	normalized disturbance force
$H_a(s)$	transfer function of a switching power amplifier
$H_c(s)$	transfer function of a current controller
$H_{xi}(s)$	transfer function between current reference and displacement in normalized form
$H_{xu}(s)$	transfer function between voltage reference and displacement in normalized form
$H_{xz}(s)$	transfer function between disturbance force and displacement in normalized form
I	controlled current
I_0	bias current of a magnet
i	normalized current
K_f	feedback coefficient of a current control loop
k_i	gain relating current to magnetic force
k_s	gain relating displacement to magnetic force
m	mass of a rotor
N	turns of coil winding around a pole
s	Laplace variable
$S_{1,2}$	Power switches 1 and 2
t	time
T_m	mechanical time constant, $2B_0\sqrt{\dfrac{\mu_0 m \delta_0}{A}}$
U_d	dc-link voltage
u	normalized voltage
X	displacement
x	normalized displacement
δ_0	mean air-gap
μ_0	permeability of free space, 1.257×10^{-6} H/m

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0	bias value
r	reference variable
z	disturbance variable

Reference values

$$\frac{X}{x} = \delta_0, \qquad \frac{I}{i} = \frac{2B_0\delta_0}{\mu_0 N}$$

$$\frac{F}{f_t} = \frac{4B_0^2 A}{\mu_0}, \qquad \frac{U}{u} = NAB_0$$

Introduction

Active magnetic bearings (AMBs) suspend a rotor with the magnetic force in the air-gap between a stator and a rotor. Due to the advantages against conventional ball bearings, active magnetic bearings are finding more and more application in industry.

Up to now the application of AMBs in industry is still thwarted by the high cost of the system, which are dominated by those of the power amplifiers: The requirement for the dynamical performance of the current control of AMBs is, in general, higher than for an electrical drive, the required bandwidth of the current control loop being typically about one or two kilohertz. In order to meet this requirement, a high voltage reserve is needed. On the other hand, the load of the power amplifier is a magnet coil. The high voltage is only required in the transient state. In the steady state, only a small voltage is required to overcome the voltage drop across the ohmic resistance of the magnet coil.

Although linear power amplifiers were used in the early development of AMBs with a small power [1][2], they are not suitable for the main part of AMB applications because of their inherently low efficiency. For this reason, AMBs are practically always supplied by switching power amplifiers that allow it to operate at high voltage and current with a high efficiency. The main disadvantage of a switching power amplifier is the larger content of current harmonics produced by the switching processes.

With a conventional switching power amplifier, the amplitude of the current ripple is inversely proportional to the switchng frequency and directly proportional to the dc-link voltage [2]. If a high dc-link voltage is required to achieve a steep current slope, it is necessary to switch at a high frequency in order to limit the current ripple. Up to now the switching frequencies of power amplifiers used in high dynamic and high precision AMBs range from about 25 to 100 kHz. This means that the power amplifier must be equipped with extra fast power elements which are not only expensive but also have considerably higher conduction losses. Furthermore, the eddy current losses in the iron core depends only on the dc-link voltage and cannot be influenced through the switching frequency.

An approach that aims at the reduction of the current harmonics without a reduction of the dynamical performance whth a hybrid power circuit has been published in [3] and [4], in which a switching power amplifier and a linear power amplifier are combined so that the advantages of the both types of power amplifiers can be fully used. The drawback of this power amplifier is the complexity of the circuit and control strategy.

In this paper, a new concept for the design of the switching power amplifier is recommended based on the combination of the advantages of a current control AMB and a voltage control AMB. A synchronous three-level PWM scheme is suggested to control the power devices of the switching power amplifier used in a digital current control AMB. With the application of such a PWM-scheme, the switching frequency can be reduced to half the sampling frequency of the digital controller at very low current harmonics. Furthermore, the dynamical performance of a digital current control AMB can also be improved.

Principle of Active Magnetic Bearings

In order to understand the new control strategy it is necessary to explain the two control strategies for active magnetic bearings.

The most commonly used principle is current control with differential driving mode as shown in *Fig. 1(a)*: Two counteracting electromagnets are used to generate a magnetic force and to control the position of the rotor in one axis. In order to obtain a good linear behaviour and a good dynamical performance of the force control, a preload is developed through a bias current. The relation between the total magnetic force in one axis F, current I and displacement X is given in Eq. (1):

$$F = \frac{\mu_0 A N^2}{4}\left[\frac{(I_0+I)^2}{(\delta_0-X)^2} - \frac{(I_0-I)^2}{(\delta_0+X)^2}\right] \quad (1)$$

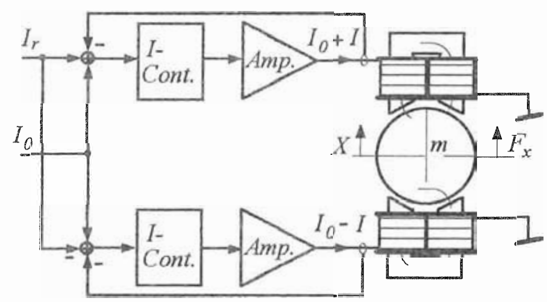
At the working point ($I=0$, $X=0$), the linearized model of the magnetic force is given in Eq. (2), in which the resulting force on the shaft is proportional to the control current I applied to both magnets and to the displacement of the shaft against the center point.

$$F = k_i \cdot I + k_s \cdot X \quad (2)$$

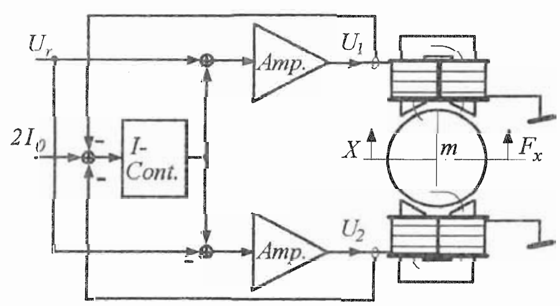
As an alternative to current control, voltage control strategy have been proposed in some publications [1][5]. The principle is shown in *Fig. 1(b)*. As the name suggests, not the current but the voltage is imposed to the coils of the AMB. As a defined bias current is necessary to maintain a stable working point, an additional bias current control loop is required. It grants for the sum of both currents always being equal to $2I_0$, which means that the bias flux density between the air gaps is B_0. If the the effect of the current control loop is not considered and furthermore, the ohmic resistances of the magnet coils are neglected, the relation between the magnetic force and the control voltage is given in Eq. (3).

$$F = \frac{4B_0}{\mu_0 N}\int_0^t U dt \quad (3)$$

It has been shown that there is a linear relation between the magnetic force and the control voltage. The magnetic force acting on the shaft is proportional to the integral of the control voltage. The drawbacks are the necessarity of an additional current control loop which affects the dynamical performance, and the difficulty to start operation.



(a)



(b)

Fig. 1 Basic construction of AMBs (a) current control AMB, (b) voltage control AMB

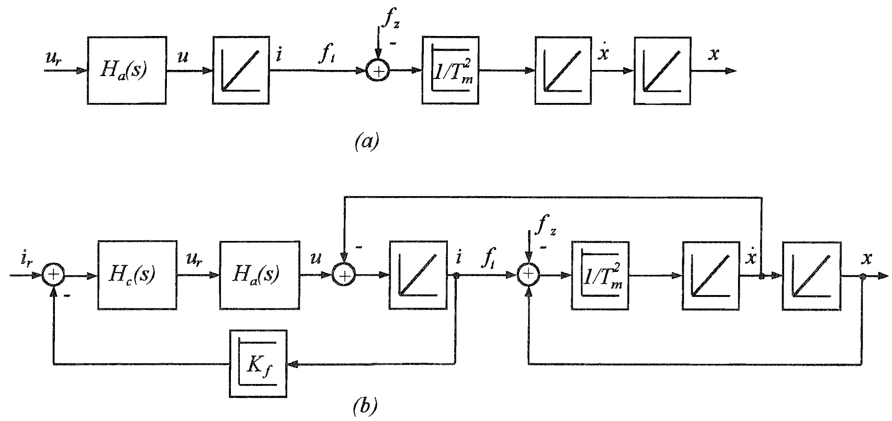


Fig. 2 Dynamical models of AMB with a power amplifier a) voltage control AMB, b) current control AMB

Dynamical model

The dynamical models of the voltage control AMB and current control AMB with their power amplifiers are shown in *Fig. 2 a)* and *b)*, respectively. In a voltage control AMB there are three integrators in cascade. In order to stabilize the system, a PD^2 position controller (or better, a state control) has to be used. The transfer functions between the output x and the control reference u_r on the one hand, and the disturbance force f_z on the other hand are given Eq. (4) and Eq. (5), respectively.

$$H_{xu}(s) = \frac{x(s)}{u_r(s)} = \frac{H_a(s)}{T_m^2 s^3} \tag{4}$$

$$H_{xz}(s) = \frac{x(s)}{f_z(s)} = -\frac{1}{T_m^2 s^2} \tag{5}$$

In the current control AMB, there are only two integrators in cascade and the position controller can be realized with a PD controller to stabilize the system. However, if the current controller is an ideal current controller, i.e. the controlled current is always equal to the reference value, the induced voltage due to speed of displacement is compensated in the current loop. As a consequence, only the positive feedback of the displacement remains, which has a negative effect on the disturbance response of the AMB, especially when the system is controlled with a digital controller at a low sampling frequency. The dynamical performance of a current control AMB is, therefore, worse than that of a voltage control AMB. The transfer functions for the control and disturbance dynamics are given in Eq. (6) and Eq. (7), respectively.

$$H_{xi}(s) = \frac{x(s)}{i_r(s)} = \frac{H_c(s)H_a(s)}{(T_m^2 s^2 - 1)[s + K_f H_a(s)H_c(s)] + s} \tag{6}$$

$$H_{xz}(s) = \frac{x(s)}{f_z(s)} = -\frac{s + K_f H_c(s)H_a(s)}{(T_m^2 s^2 - 1)[s + K_f H_c(s)H_a(s)] + s} \tag{7}$$

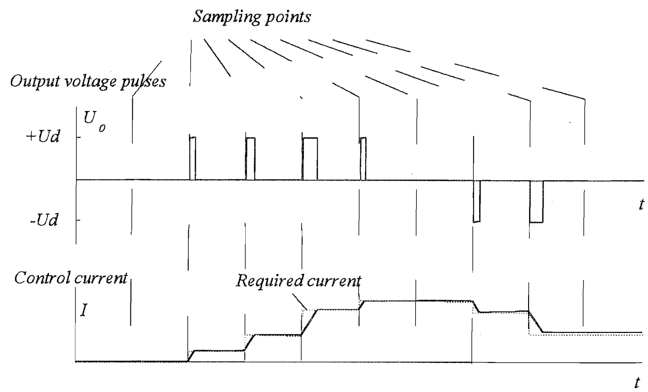


Fig.3 Coil voltage, required current and actual control current in an AMB with synchronous three level PWM control.

When the current controller is ideal, $K_f = 1$ and the transfer functions for the control and disturbance dynamics follow Eq. (8) and Eq. (9), respectively.

$$H_{xi}(s) = \frac{x(s)}{i_r(s)} = \frac{1}{T_m^2 s^2 - 1} \tag{8}$$

$$H_{xz}(s) = \frac{x(s)}{f_z(s)} = -\frac{1}{T_m^2 s^2 - 1} \tag{9}$$

In this case, the transfer function of the system contains a pole pair on the right half s-plane, which will cause a larger displacement due to a force disturbance than that of a voltage control AMB.

If the current control loop is open ($K_f = 0$), the transfer functions become to be as in Eq. (10) and Eq. (11). In this case, the dynamical performances of the current control AMB is the same as that of the voltage control AMB.

$$H_{xi}(s) = \frac{x(s)}{i_r(s)} = \frac{H_a(s)H_c(s)}{T_m^2 s^3} \tag{10}$$

$$H_{xz}(s) = \frac{x(s)}{f_z(s)} = -\frac{1}{T_m^2 s^2} \tag{11}$$

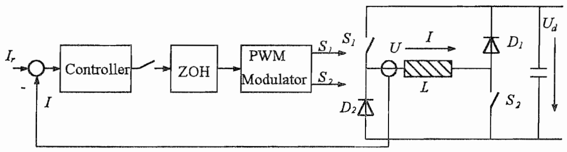


Fig. 4 Current control switching power amplifier

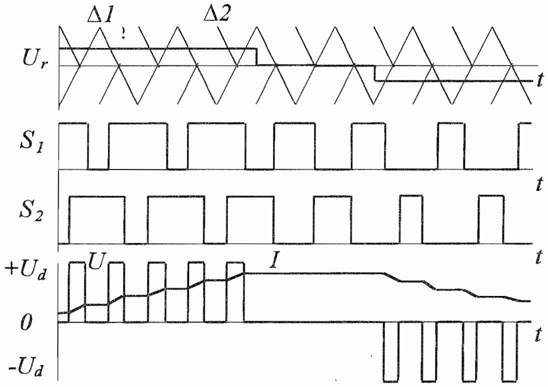


Fig. 5 Waveforms of three-level PWM scheme

The Principle of the Synchronous three-level PWM

In a digitally controlled AMB, the position control loop is open between two adjacent sampling times. According to the discussion above, the disturbance response is better if the current control loop is also left open during this period. This is realized with the control strategy proposed in this paper: The current loop is closed ($K_f = 1$) only at the sampling instances of the position controller. Between two adjacent sampling points, the current loop is left open ($K_f = 0$). During this period, the coil voltage performs a predifined cycle (see *Fig. 3*) that begins with a positive or negative voltage pulse directly after the sampling point that makes the current reach the required value, followed by zero voltage, where the current stays almost constant and the dynamical performance of the AMB is the same as that of a voltage control AMB, so that the displacemant due to a force disturbance during this time is smaller than that of a conventional current control AMB.

The amplifier has a conventional H-bridge architecture as shown in *Fig. 3*. The two switching devices S1 and S2 are switched independently from each other. This allows it to generate four switching states, two of which are equivalent:

+1	S1 and S2 closed	$U = +U_d$
0L	S1 open, S2 closed	$U = 0$
0H	S1 closed, S2 open	$U = 0$
-1	S1 and S2 open	$U = -U_d$

As long as the measured current equals the required current, S1 and S2 are commuted simultaneously at a fixed frequency so that the switching state changes directly between 0L and 0H. If the required value of the current is higher than the measured value, the closing of the open switch (either S1 or S2) is retarded by a delay time that is proportional to the current error. During this delay time, the switching state is +1 and the current rises until it reaches the required value. If the current error is negative, the closing is delayed and the switching state passes -1. As the delay is predefined, it is not influenced by an induced voltage so that the current loop is left open during the whole switching period. This is one of the three advantages of the synchronous three-level PWM in comparison to the conventional current controlled two-level PWM. The second advantage of the synchronous PWM is that the fact that the current reaches the required value as quickly as possible so that the delay time of the amplifier is minimal. In contrast to this, if the PWM signal is asynchronous as in *Fig. 5*, the current is increased in various steps so that it would not reach the required value before the end of the interval. At the same time, the switching frequency must be considerably higer than the sampling frequency of the control cycle in order to grant for a time-invariant behaviour of the amplifier. That is the reason why the synchronous three-level PWM shows a better dynamic performance at a lower switching frequency.

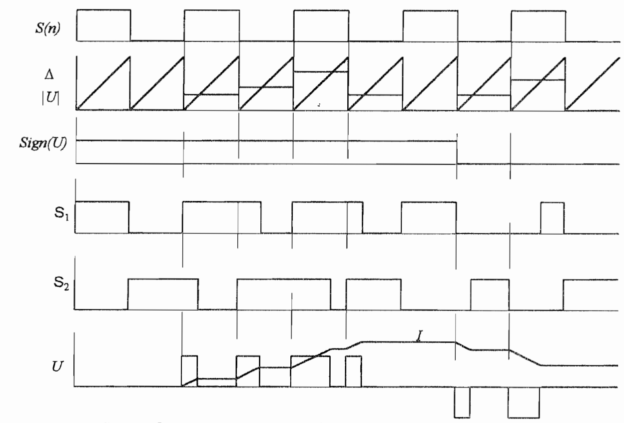


Fig. 6 Waveforms of a synchronous three-level PWM-scheme

Realization of a Synchronous three-level PWM

The synchronous three-level PWM signals are produced with a special hybrid circuit. The PWM scheme cannot be realized with a conventional PWM generator, in which the PWM signals are generated through the crossing between a triangular signal and a reference signal. It is possible to use a microcontroller to realize the current controller and to generate the PWM signal. There will always be a time delay of one sampling period due to the response delay of the timer on chip after a new word is written. The current controller and the PWM generator have therefore been realized with a special hybrid circuit in this work. The principle of the circuit is shown in the block scheme

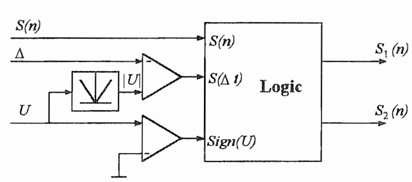


Fig. 7 Realization of the PWM generator

Fig. 6. The signal *S(n)* is the synchronization signal which is produced by the digital controller; the rising and falling edges of the signal correspond to the sampling points. The delay time is generated through comparison between the absolute amplitude $|U|$ of a reference voltage that is proportional to the current error and a sawtooth signal Δ. The sign of the error signal is evaluated by the logic chip that determines whether the "on" signal or the "off" signal should be delayed. The output signals of the logic block are the signals for the two switching devices of one H-bridge. The logic block has been realized with a GAL-chip.

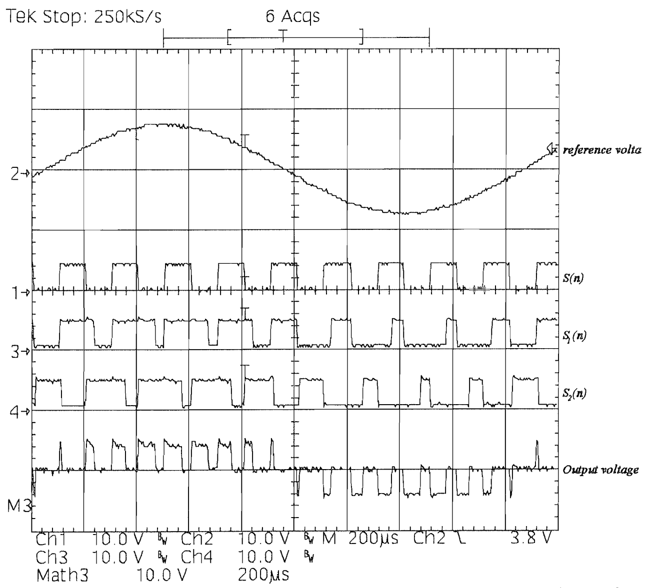


Fig. 8 Measured waveforms of the synchronous three-level PWM generator

Test and simulation results

The experimental circuit of the PWM-modulator has so far been realized and tested with a switching power amplifier and a magnet coil. *Fig. 8* shows the measured waveforms of the circuit, in which the reference voltage is a sinusoidal signal.

To compare the behaviors of digitally controlled magnetic bearings with various pulse modulation schemes and switching frequencies, *Fig. 8* shows the simulation results of the displacement response to a force step from zero to 50 % maximum magnetic force.

The parameters of the system are

Magnetic bearing:

$\delta_0 = 0.5$ mm, $I_0 = 2$ A, $m = 1$ kg, $N = 200$, $A = 4$ cm^2, $B_0 = 0.5$ T

Amplifier:

Maximum voltage	310	V
Maximum current	4	A

The position controller is a digital controller with a sampling rate of 10 kHz.

The waveform of position responses in the figure are:

a) voltage control with an ideal power amplifier;

b) synchronous three-level PWM current control at the synchronous switching frequency of 5 kHz and a dead-beat current controller;

c) asynchronous three-level PWM current control with a switching frequency of 50 kHz and a dead-beat current controller;

d) two-level PWM current control with a switching frequency 100 kHz and a dead-beat current controller;

The results are given in normalized form, the reference value is δ_0.

The response with the ideal voltage control is obviously better than the other three cases. This simulation result depicts, however, merely the the limitation that is inherent to the discrete time control; it does not consider either the current limitation or the saturation limit.

The differences between the behaviors of the three cases with the three (realistic) current control strategies are fairly small. Still the synchronous three level PWM shows the best result. The main advantage of this control strategy is, however, the fact that the switching frequency can be reduced by a factor ten or twenty, respectively. The response with an asynchronous three-level PWM at the same switching frequency would be notably worse, and a two-level PWM at the same frequency could not be realized at all because the peak-to-peak amplitude of the current ripple would reach 4 A. At the same time, the eddy current losses with the three-level PWM with respect to a two-level switching mode are lower by a factor that corresponds the ratio between the "on" time (+1 stat ore -1 state) to the sampling period. This factor can easily reach the order of twenty.

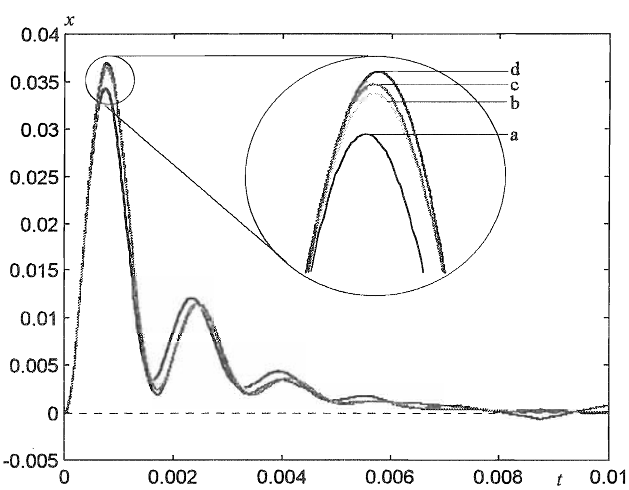


Fig. 9 Simulation results of the disturbance response of AMBs

Conclusions

The synchronous three-level PWM scheme is optimal for a digital current control AMB. The advantages of a voltage control AMB can entirely be exploited in a digital current control for an AMB that is supplied with switching power amplifiers in the sense that the disturbance response of the current control AMB is improved. The switching frequency can be reduced by a factor about twenty with respect to asynchronous switching modes. This means that no special high speed power elements or power circuit design are necessary. At the same time the current harmonics and the eddy current losses can be reduced considerably.

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