Abstract—This paper describes a novel and compact design concept for contactless levitation and rotation of a wide annular rotor through the walls of a process chamber. In the proposed set-up, a homopolar magnetic bearing biased by permanent magnets is combined with a high-pole-number segment motor. The paper describes the functional principle of the motor focusing on the design and the behavior of the drive. An experimental system is presented along with a comprehensive set of measurement results verifying the theoretical considerations.

Index Terms—Magnetic bearing, bearingless motor, hollow shaft drive, process chamber.

I. INTRODUCTION

The reproducibility of many industrial processes can be improved by the use of hermetically sealed process chambers, since they facilitate proper control of crucial process parameters such as pressure, temperature, or composition of the gas or liquid surrounding an object that is to be processed. These chambers are used in biotechnological applications [1] as well as in the semiconductor industry [2], where they also prevent leakage of potentially harmful chemicals into the environment. State-of-the-art wafer production requires a device placed inside of the sealed chamber and which rotates the wafer during processing, thereby providing a uniformly distributed concentration and flow pattern. The design of existing process chambers for wafer production relies on a gas-tight feed through of the rotating drive axis (in the wall of the process chamber) and is, therefore, afflicted with two major disadvantages. Not only are the drive shaft, the shaft sealing, the bearings, and the motor rather bulky, sometimes occupying a significant portion of the space available inside of the chamber, but they also generate particles detrimental for the media inside of the chamber at every frictional contact between moving parts.

In this paper, a novel bearingless hollow shaft drive, which is based on the principle of the “Bearingless Motor” ([3] - [9]), is described. This drive allows for contactless levitation and rotation effectuated through the walls of the chamber on a rotor to which a wafer clamping device is attached (Fig. 1). Due to the entire integration of the drive and the magnetic bearing, the over-all size of the motor is significantly smaller than any design with the magnetic bearings separated from the drive. All elements necessary for the drive and bearings (coils, sensors and electronics) are placed outside the chamber and are integrated in the motor stator. Also the corresponding power and data processing electronics can be placed outside the process chamber, and is therefore not exposed to harmful process chemicals and can be replaced quickly in the case of a failure. The object to be treated is clamped in the middle of an annular rotor and the whole place in the chamber above and beneath the object remains free from any bearing or drive elements. Therefore, process sources, such as nozzles or radiation sources, can be placed on both sides of the object, which allows a simultaneous treatment of both object surfaces.

Fig. 1. Arrangement of a hermetically sealed process chamber with the proposed bearingless hollow shaft drive and two process sources. All electric parts of the system are placed outside the chamber, only the rotor floats inside the process chamber and is levitated by magnetic fields through its walls. Therefore, the rotor is easy accessible from both sides.
II. MOTOR FUNCTIONAL PRINCIPLE

A. Bearing

In principle, the rotor has six spatial degrees of freedom: linear motion and rotation along each of three axes in an xyz-system. While the rotor is held in place by the bearing, all three of its translatory modes and two of its rotatory modes of motion must be suppressed. The only remaining degree of freedom is the desired rotation around the z-axis.

In contrast to conventional magnetically levitated drives, the bearing forces in a “Bearingless Motor” are not built up in additional magnetic bearings placed along the axis of rotation, but in the motor itself: The active motor part generates the torque as well as radial magnetic bearing forces. Because the length of the rotor is small compared to its diameter, it is possible to stabilize three spatial degrees of freedom (z position, tilting moment in x and y direction) passively by attractive magnetic forces (minimization of reluctance, Fig. 2) so that only one active radial bearing (for x and y position) is needed. Although being stabilized only passively, the axial position of the rotor lies within a narrow range for small loads. For special applications, its accuracy could be improved by additional or stronger permanent magnets or by adjusting the axial position of the stator.

Fig. 3 shows the functional principle of the active radial bearing. Permanent magnets placed on the rotor and on the stator provide a bias flux in the air gap. Depending on the axial position of the rotor, the bearing windings are supplied with a current, which alters the flux density in the air gap, thereby generating a resulting Maxwell-force towards the target position. In order to maximize this Maxwell-force, both opposite bearing windings are connected against each other.

The use of high-energy permanent magnets to generate a magnetically biased bearing flux leads to a compact system: The neodymium iron boron magnets used in this set-up have coercive field strengths of more than 1000kA/m. As an example, a coil with 8’000 ampere turns generates the same field strength as a magnet with a height of only 8mm. However, a reasonable force on the rotor is obtainable with a comparatively low bearing current due to the square dependence of the Maxwell-force on the magnetic flux density.

It must be noted that, due to the relative permeability of a permanent magnet, which is approximately unity, a bearing design, where (additional) permanent magnets are placed horizontally in the direction of the air gap flux, is inadvisable. In such a design, the magnetic air gap would automatically be enlarged by the size of the magnets if the physical air gap is kept constant. An enlarged air gap leads to a reduced flux density, since the flux density is inversely proportional to the gap width. This will result in a drastically reduced bearing force, since the Maxwell-force scales with flux density squared.

B. Drive

For the applications described in the introduction, only a moderate torque is required to overcome the rotor torque resulting from inertia. A rapid acceleration is crucial to ensure a reasonable cycle time. A speed range from 1rpm (for allowing a more homogeneous exposure of a process to the object) up to 1500rpm (spin cycle) must be covered. For the rotation to be uniform even at a speed as low as 1rpm, a motor configuration with a high pole number is a necessary prerequisite.

In our configuration, which was introduced earlier [1], the bearing bias magnets on the rotor are used to generate a high-pole, modulated but rectified bias flux. For this purpose, the magnets mounted on the rotor are not placed next to each other, but in pole distance (Fig. 4). To retain the...
homopolarity of the bearing, the pole width of the bearing stator claws must be matched to the pole width of the rotor. This modulated bias flux in interaction with the motor segments of the same pole width on the stator generates the torque (synchronous rotating field machine). Obviously, the maximum possible torque is limited by the size of the air gap.

For the field orientated speed control, two hall sensors - placed on the stator - determine the rotor angle in a sine-cosine analysis. For a detailed information on the current and position control procedure we refer to [7] and [10] for the sake of brevity.

III. MOTOR DESIGN ASPECTS

In this section, the basic procedure of the drive design is described. The determination of the proper number of turns of the drive coils is depicted as well as the resulting description of the drive design is followed by an optimization procedure focused on the objective to reach a description of the features of two existing motor controllers A and B defined rotor speed as quickly as possible while considering variations in the amplitudes of the induced voltage and the motor torque, respectively, is just the alternating component of B. A as a result of the large leakage fields, a homogeneous distribution of the magnetic flux density B over the entire pole face cannot be assumed; dividing the measured value $u_{\text{ind}}(t)$ by A leads, therefore, to a derivative, $dB/dt$, averaged over the entire pole face.

Saturation effects in this set-up are of minor importance due to the big air gap and the comparatively small number of ampere turns, even if the drive coils are supplied with the maximal drive current.

The time-dependent behavior of $B(t)$ is virtually sinusoidal as a spectrum analysis of $u_{\text{ind}}(t)$ shows in Fig. 5:

$$B(t) = \dot{B} \cdot \sin(2\pi f_d \cdot t).$$

The frequency $f_d$ of the induced voltage $u_{\text{ind}}(t)$ is calculated from the rotor speed $n_R$ (in rpm) and the number of pole pairs $p$ of the rotor:

$$f_d = \frac{n_R}{60} \cdot p.$$  

From this, the induced voltage can be derived as:

$$u_{\text{ind}}(t) = 2\pi \frac{n_R}{60} \cdot p \cdot N \cdot \dot{B} \cdot A \cdot \cos(2\pi \frac{n_R}{60} \cdot p \cdot t).$$

Tab. 1 shows the values of the induced voltage measured in the experimental set-up. Due to the proportionality of $U_{\text{ind},\text{RMS}}$ and the rotor speed $n_R$ a speed independent and al-

<table>
<thead>
<tr>
<th>$n_R$ [rpm]</th>
<th>$f_d$ [Hz]</th>
<th>$U_{\text{ind},\text{RMS}}$ [V]</th>
<th>$k_{\text{ind}}$ [mV/rpm]</th>
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<tr>
<td>100</td>
<td>36.7</td>
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<tr>
<td>1200</td>
<td>440.0</td>
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Tab. 1. RMS-value of the induced voltage $u_{\text{ind}}(t)$ and $k_{\text{ind}}$ measured in the experimental set-up ($p = 22$) in one drive coil ($N = 190$) for different rotor speeds.
most constant factor $k_{\text{ind}}$ can be calculated by averaging:

$$k_{\text{ind}} = \frac{1}{i} \sum \frac{U_{\text{ind},RMS}}{n_k}. \quad (5)$$

### B. Equivalent Electrical Circuit of one Drive Phase

The equivalent electrical circuit consists of the above-derived induced voltage $U_{\text{ind}}$, of the coil resistance $R_C$, the coil inductivity $L_C$, and of the voltage source $U_i$. All of the $k$ drive coils can be connected in series (Fig. 6(a)) or, for example, the two coils that are attached in series connection on one drive lamination stack can be connected in parallel (Fig. 6(b)). Applying Kirchhoff’s law to the equivalent electrical circuit for the configuration with drive coils in series connection leads to

$$U_i = k \cdot R_C \cdot I_1 + k \cdot j \cdot \omega \cdot (L_C + L_k) \cdot I_1 + k \cdot U_{\text{ind}}, \quad (6)$$

or for the second configuration, respectively, to:

$$U_i = \frac{k}{4} \cdot R_C \cdot I_1 + \frac{k}{4} \cdot j \cdot \omega \cdot (L_C + L_k) \cdot I_1 + \frac{k}{2} \cdot U_{\text{ind}}. \quad (7)$$

The following calculations are carried out corresponding to the configuration depicted in Fig. 6(a). The induced voltage is in phase with the drive current $I_1$ and, together with the phasor chart (see Fig. 7(a)), the following equation can be derived:

$$(k \cdot U_{\text{ind}} + k \cdot R_C \cdot I_1)^2 + (\omega \cdot k \cdot \omega \cdot (L_C + L_k) \cdot I_1)^2 = U_i^2. \quad (8)$$

The resistance $R_C$ of one coil is calculated applying

$$R_C = \rho \cdot (1 + \alpha \cdot (T - T_0)) \cdot \frac{N \cdot l}{(d/2)^2 \cdot \pi}, \quad (9)$$

where $\rho$ is the resistivity, $\alpha$ the temperature coefficient, $T$ the operating temperature (e.g. 75°C), $N$ the number of windings, $l$ the average length of one winding, and $d$ the wire diameter.

The inductivity $L_C$ of one coil is given by

$$L_C = \frac{\mu \cdot N^2 \cdot A}{l}, \quad (10)$$

where $\mu = \mu_0 \cdot \mu_r$ is the permeability, $N$ the number of windings, $A$, the coil cross-sectional area, and $l$, the coil length. Using (10), the drive coil inductivity $L_C$ of this configuration cannot be calculated. However, the quadratic relation of $L_C$ with $N$ can be used to calculate $L_C$ for a given number of turns $N_1$ of the drive winding, starting from an initial coil inductivity measurement $L_0$ of a drive coil with a specified number of turns $N_0$:

$$L_C = \frac{N_1^2}{N_0^2} \cdot L_0. \quad (11)$$

### C. Drive Current and Induced Voltage Depending on the Rotor Speed and the Number of Turns of the Drive Coils

The drive current $I_1$ can be obtained from (8):

$$I_1 = \frac{-U_{\text{ind}} \cdot R_C \pm \sqrt{(R_C^2 + \omega^2 \cdot L_C^2) \cdot U_{\text{ind}}^2 - \omega^2 \cdot L_C^2 \cdot U_{\text{ind}}^2}}{2 \cdot (R_C^2 + \omega^2 \cdot L_C^2)}. \quad (12)$$

With the formulas derived above, $U_{\text{ind}}$ and $I_1$ (not listed) are calculated as a function of the rotor speed $n_k$ with the optimized number of turns $N_1$ as a parameter:

$$U_{\text{ind}} = k_{\text{ind}} \cdot N_k \cdot \frac{N_1}{N_0}. \quad (13)$$

Now, the phasor charts and the behavior of $U_{\text{ind}}$ and $I_1$ can be calculated (e.g. with MATLAB) for different numbers of turns. The current limit $I_{1,max}$ and the maximum output voltage $U_{\text{ind},max}$ given by the motor controller specifications can also be taken into consideration of the drive coil design.

Fig. 7 shows the phasor charts for $N_1 = N_0 = 190$ and $k = 4$ for a motor controller (type B) with a dc link voltage of $U_{\text{DC}} = 325V$ and the drive current limit of $I_{1,max} = 5.3A$ for a rotor speed of (b) 100rpm, (c) 500rpm, and (d) 1000rpm with in each case the maximal possible drive current $I_1$ and the maximal motor torque respectively.
A first prototype of the outlined hollow shaft drive has
been realized (see Fig. 10 and Fig. 9, and its key data in Tab. 2) and its rotor acceleration time from 0 up to 1200rpm successfully optimized according to the guidelines given in this section.

IV. CONCLUSIONS

This paper briefly describes a novel and compact design concept for contactless levitation and rotation of a rotor through the walls of a process chamber. In the proposed setup, a homopolar magnetic bearing biased by permanent magnets is combined with a high-pole-number segment motor.

The large width of the air gap and the drive configuration result in a weak magnetic coupling of rotor and stator. Thus, the drive design requires a compromise between a large induced voltage and a small number of turns of the drive windings, since a large inductance combined with a high number of poles leads to a major inductive voltage drop. Particularly, a motor supplied by a converter with a low dc-link voltage has a reduced maximum possible drive current already at a comparatively low rotor speed. An optimization with respect to the maximum achievable acceleration leads therefore to a moderate torque in the low-speed range.

REFERENCES


<table>
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<td>In one drive coil induced voltage</td>
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Tab. 2. Key data of the experimental set-up