Design and Optimization of a Rotary Actuator for a Two Degree-of-Freedom $z\phi$-Module

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Abstract—The paper concerns the design and optimization of a rotary actuator of which the rotor is attached to a linear actuator inside a two degree-of-freedom $z\phi$-module, which is part of a pick-and-place robot. The rotary actuator provides $\pm180^\circ$ rotation, while the linear actuator offers a $z$-motion of $\pm5$ mm. In the paper, the optimal combinations of magnet poles and coils are determined for this slotless actuator with concentrated windings. Based on this analysis, of the rotary actuator is optimized using a multi-physical framework, which contains a coupled electromagnetic, mechanical and thermal model. Because the rotation angle is limited, both a moving-coil design with a double mechanical clearance and a moving-magnet design with a single mechanical clearance have been investigated and compared. Additionally, the influence of the edge effects of the magnets on the performance of the rotary actuator has been investigated with both 3D FEM simulations and measurements.

I. INTRODUCTION

Pick-and-place (P&P) actuators consist of a long-stroke robot arm, which is responsible for moving electrical components (such as Surface-Mounted-Devices or Ball-Grid-Array components) from a feeder over the Printed-Circuit-Board (PCB). Attached to this arm, a placement module picks up the components, orientates and places them on the PCB. This high-precision, short-stroke, two degree-of-freedom (2 DoF) actuator, therefore, needs to enable rotational and translational motion. Moreover, they require a more compact and light weighted design, due to higher accelerations and operational speeds. Such a module will be referred to as a $z\phi$-module.

Several types for $z\phi$-modules can be distinguished. In the first category and classic approach, the linear stroke actuator drives a trolley which holds a rotary actuator. As an alternative, in the second category the movers of the rotary and linear actuator are attached to the same shaft above each other. For example, in [1] a $z\phi$-module from the second category is discussed which uses two separate magnet arrays along the axial length of the rotor; one array for the rotary actuator and one array for the linear actuator. Instead of stacking two magnetization patterns, they can also be integrated in a single magnet array, as is done in the third category. Using a checkerboard magnet array and by appropriately commutating the current inside the coils, the design in [2] offers both degrees of freedom.

Based on the second category, a $z\phi$-module is designed, which has to pick-and-place 10.000 BGA-components in one hour. Figure 1 shows a 3D overview of this design, where the rotary actuator is placed at the top and the linear actuator is placed at the bottom of the module. The specifications of the $z\phi$-module are given in Table I. The rotary actuator is a slotless permanent magnet actuator and provides $\pm180^\circ$ degrees rotation. Because the total stroke of the linear actuator is only 10 mm, a non-commutated short-stroke linear actuator is selected instead of a three phase linear actuator, like is used in [1].

This paper concerns the design and optimization of the rotary actuator as part of the $z\phi$-module, while the design of the non-commutated short-stroke linear actuator is fixed and will be considered as load. First, the basic design regarding the rotary actuator is addressed and the test results of a non-
optimized pre-prototype are discussed. Prior to the optimization step, an optimum combination of the number of magnet poles and coils is selected. A multi-physical framework of the rotary actuator is presented and used to optimize the design, which is performed for both a configuration with moving magnets and a configuration with moving coils. Finally, the optimized designs are analyzed for the magnet edge-effects.

II. BASIC DESIGN CONSIDERATIONS AND PRE-PROTOTYPE TESTING

The basic design of the three-phase AC brushless permanent magnet rotary actuator is shown in Fig. 2. It has a slotless structure in order to provide a low torque ripple. To increase the magnetic loading inside the airgap, a two segmented quasi-Halbach magnet array is used. Concentrated windings are selected for their ease of manufacturing. Because the rotary actuator also needs to accommodate the translational motion of the rotor, the coils are elongated in order to deliver constant torque for any given axial position. Because the rotational angle of the rotary actuator is limited to ±180°, both a moving-magnet and a moving-coil configuration will be optimized and compared. In the moving-magnet configuration the coils are attached to the back-iron leaving only a single mechanical clearance between the coils and magnets. For moving coils, however, a double mechanical clearance is required; one additional clearance between the coils and back-iron. Figure 2 shows an overview of the rotary actuator for the moving-magnet and moving-coil configuration. In both cases, the rotor is coupled to the rotor of the linear actuator.

A non-optimized pre-prototype of the rotary actuator has already been designed, built and tested. Also a prototype of the non-commutated short-stroke actuator is built, but as the linear actuator is out of the scope of this work, its specifications are given in Table II. The design of the rotary actuator is based on a configuration with moving coils and its dimensions are given in Table III. Figure 3 shows the pre-prototype. Each coil in the slotless PM actuator contains 92 turns and the magnets are anisotropic sintered NdFeB, with a remanent flux density of 1.33 T and a relative recoil permeability of 1.1. The core and back-iron are made of steel N398 with a saturation flux density of 1.5 T. Using the 2D semi-analytical magneto-static model, as described in Section IV, a torque constant of 0.93 Nm A−1 is estimated, however, tests on the pre-prototype showed a reduced amplitude of 11.7 %. To eliminate the influence of friction, electro-motive force (EMF) measurements were performed, which showed a reduced amplitude of 11.7 % compared to the 2D semi-analytical model. This reduction is caused by the lower magnetic loading near the edges of the magnets, as will be discussed in Section V. The test also showed that the helical wound electrical wires (see Fig. 1) were insufficient to guarantee a long life operation.
### TABLE III
**Dimension of initial design rotary actuator.**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>p</td>
<td>8</td>
<td>Number of pole pairs</td>
</tr>
<tr>
<td>Q</td>
<td>12</td>
<td>Number of coils</td>
</tr>
<tr>
<td>r1 [mm]</td>
<td>18.0</td>
<td>Inner radius</td>
</tr>
<tr>
<td>ro [mm]</td>
<td>30.0</td>
<td>Outer radius</td>
</tr>
<tr>
<td>tsc [mm]</td>
<td>1.5</td>
<td>Radial length core</td>
</tr>
<tr>
<td>tm [mm]</td>
<td>5.7</td>
<td>Radial length magnets</td>
</tr>
<tr>
<td>lw [mm]</td>
<td>1.9</td>
<td>Radial length coils</td>
</tr>
<tr>
<td>lb [mm]</td>
<td>2.5</td>
<td>Radial length back-iron</td>
</tr>
<tr>
<td>tcl [mm]</td>
<td>0.2</td>
<td>Radial length mechanical clearance</td>
</tr>
<tr>
<td>Laclt [mm]</td>
<td>15.0</td>
<td>Active length actuator</td>
</tr>
<tr>
<td>α</td>
<td>0.67</td>
<td>Magnet arc to pole arc ratio</td>
</tr>
<tr>
<td>β0 [degree]</td>
<td>1.70</td>
<td>Coil opening angle</td>
</tr>
</tbody>
</table>

**Fig. 3.** Pre-prototype of the rotary actuator with (a) the test set-up and (b) the coil array.

### III. Selection number of magnet poles and coils

As a first step in the design process of the rotary actuator, the appropriate combinations of the number of magnet poles (2p) and coils (Q) are selected. First of all, only combinations resulting in a balanced structure will be evaluated during the optimization step. Next, the selection is further reduced by choosing magnet pole and coil combinations based on the winding factor, \( k_w \). The winding factor is a measure for the flux linkage of a certain winding layout and ranges from 0 (no linkage) to 1 (optimal linkage). The torque produced by a three-phase AC brushless PM actuator is related to the flux linkage according to

\[
T = \frac{3}{2\omega} I \hat{E} = \frac{3}{2\omega} I k_w \frac{d\Lambda_{\text{max}}}{dt},
\]

where \( I \) is the peak phase current, \( \hat{E} \) is the peak phase back-EMF, \( \omega \) is the angular velocity and \( \Lambda_{\text{max}} \) is the maximum possible flux linkage of a single phase. As the expression shows, the torque is directly linked to the winding factor, and it would be desirable to select a combination of magnet poles and coils resulting in the highest possible winding factor.

The winding factor can be split into

\[
k_w = k_p \cdot k_d \cdot k_{\text{skew}}, \tag{2}
\]

where \( k_p \) is the pitch factor, \( k_d \) is the distribution factor and \( k_{\text{skew}} \) is the skewing factor. Since no skewing is applied, the skewing factor is assumed to be unity. The winding factors can only be found for slotted structures, though ([3],[4],[5],[6]). Therefore, for slotted machines with concentrated windings, the fundamental winding factor is calculated for different combinations of the number of magnet poles and coils. Below, both the pitch and distribution factor will be determined, when only the fundamental harmonic of the magnetic flux density due to the magnets is considered.

#### A. Pitch factor

The pitch factor is a measure for the flux linkage of a single coil. Ideally, the total flux through one magnet pole is linked by all turns in a coil, which is achieved in slotted machines having one coil per magnet pole and per phase. Whereas the teeth in a slotted machine provide a low reluctance path for the flux through the coil and all turns in the coil link the same amount of flux, in slotted machines however, this low reluctance path is not provided and, hence, not all turns show the same flux linkage. To determine the pitch factor in a slotless machine as is shown in Fig. 4, the average flux linkage of a single turn is calculated, by varying the angular span, \( 2\alpha_t \), of a turn at radius \( r \)

\[
\Psi_{av} = \frac{1}{\alpha_c - \beta_o} \int_{\alpha_c}^{\alpha_o} \Psi_1(\alpha_t)d\alpha_t
= \frac{1}{\alpha_c - \beta_o} \int_{\alpha_o}^{\alpha_c} r L_{\text{act}} \hat{B}_1(r) \cos(p\alpha_t) d\alpha_t
= \frac{2r L_{\text{act}} \hat{B}_1(r)}{\alpha_c - \beta_o} \left[ \cos(p\beta_o) - \cos(p\alpha_c) \right], \tag{3}
\]

where \( 2\beta_o \) is the opening angle of a coil, \( 2\alpha_c \) is the angular span of a coil, \( L_{\text{act}} \) is the length in axial direction and \( \hat{B}_1 \) is the amplitude of the fundamental harmonic of the magnetic flux density. With the maximum flux linkage of a single turn being equal to the flux through one magnet pole, \( \Psi_{\text{max}} = \frac{2r L_{\text{act}} \hat{B}_1(r)}{p} \), the pitch factor, \( k_p \), for a slotless machine with concentrated windings becomes

\[
k_p = \frac{\Psi_{av}}{\Psi_{\text{max}}} = \frac{1}{p\alpha_c - p\beta_o} \left[ \cos(p\beta_o) - \cos(p\alpha_c) \right]. \tag{4}
\]

#### B. Distribution factor

The distribution factor is a measure of the electrical alignment of all coils in a single phase and can be calculated by writing the back-EMF of a single coil in phasor representation (in per unit)

\[
\overline{E}_{i,\text{pu}} = e^{j(\gamma_i)},
\]

where \( \gamma_i = \frac{2\pi i}{p} \), \( i \) is the electrical angle offset of coil \( i \).

Figure 5 shows an example of a phasor representation of the back-EMF for the individual coils and also the resulting phase back-EMF phasors, which are symmetrically shifted by 120° electrical degrees from each other. The minus sign for a phasors means that the windings of the coil are connected in the opposite direction. The phase back-EMF phasors (in per unit) can be calculated according

\[
\overline{E}_{\text{ph,pu}} = \sum_{Q/3} \overline{E}_{i,\text{pu}}.
\]
For both cases it can be noticed that machines with an angle of the coil, it is calculated for the case when the opening angle is zero and half the total coil span (\(2\alpha_t = \pi/Q\)). Combinations resulting in the same number of coils per phase, as given by

\[ n = \frac{Q}{p} \]

and, therefore, create the peaks in Fig. 6.

In the further analysis, only magnet poles and coils combinations resulting in \(q=1/4\) are considered because they high winding factor and can be obtained with every number of magnet poles which is a multiple of four. During optimization the number of magnet poles is varied and the number of coils is determined according to \(Q=6qp\).

IV. Multi-physical framework

A multi-physical framework, containing a 2D semi-analytical magneto-static, thermal and mechanical model, is derived to obtain an optimized design of the rotary actuator. In the following subsections these models are discussed.

A. 2D semi-analytical magneto-static model

The magnetic field distribution inside the slotless permanent magnet actuator, as depicted in Fig. 7, is obtained by using the method as described in [7]. This method provides an accurate and fast means of determining the torque capabilities and an estimate of the saturation inside the iron parts of actuator, while accounting for the relative recoil permeability, \(\mu_r\), of the magnets. In this subsection the models obtained from this technique are further extended to account for the two-segmented Halbach array with straight magnetization, which is shown in Fig. 8.

The method assumes two-dimensional fields in cylindrical coordinates, infinite permeable iron and linear demagnetization of the magnets in the second-quadrant. Due to the infinite permeability of the iron parts, only the magnetic fields inside the airgap \(\left(r_m \leq r \leq r_b, \text{ indicated by subscript } I\right)\) and the magnets \(\left(r_c \leq r \leq r_m, \text{ indicated by subscript } II\right)\) have to be determined. By introducing the vector potential according to

\[ \mathbf{B} = \nabla \times \mathbf{A}, \]

and assuming no current density inside the airgap, Ampère’s law for the airgap region can be rewritten into the Laplace
For both regions the following constitutive relations are used
\[ \nabla^2 \mathbf{A}_I = 0. \]  
Likewise, the Poisson equation for the magnet region can be obtained
\[ \nabla^2 \mathbf{A}_{II} = -\mu_0 \nabla \times \mathbf{M}_0. \]  
For both regions the following constitutive relations are used
\[ \mathbf{B}_I = \mu_0 \mathbf{H}_I, \]  
\[ \mathbf{B}_{II} = \mu_0 \mu_r \mathbf{H}_{II} + \mu_0 \mathbf{M}_0. \]  

where \( \mathbf{M}_0 \) is Fourier series representation of the remanent magnetization vector and is given by
\[ \mathbf{M}_0 = \sum_{n=odd}^{\infty} \left[ M_{nr} \cos(n \phi) \mathbf{e}_r - M_{n\phi} \sin(n \phi) \mathbf{e}_\phi \right]. \]

The coefficients \( M_{nr} \) and \( M_{n\phi} \) are obtained from the Fourier transformation of the magnetization waveform, which for the two-segmented Halbach array with straight magnetization is given by
\[ M_r = \frac{-B_{rem}}{\mu_0} \cos\left(\phi + \frac{\pi}{p}\right) \]
\[ M_\phi = \frac{B_{rem}}{\mu_0} \sin\left(\phi + \frac{\pi}{p}\right) \]
\[ M_r = \frac{B_{rem}}{\mu_0} \sin\left(\phi + \frac{\pi}{2p}\right) \]
\[ M_\phi = \frac{-B_{rem}}{\mu_0} \cos\left(\phi + \frac{\pi}{2p}\right) \]
\[ M_r = \frac{B_{rem}}{\mu_0} \cos\left(\phi \right) \]
\[ M_\phi = \frac{B_{rem}}{\mu_0} \sin\left(\phi \right) \]
\[ M_r = \frac{-B_{rem}}{\mu_0} \cos\left(\phi - \frac{\pi}{2p}\right) \]
\[ M_\phi = \frac{B_{rem}}{\mu_0} \sin\left(\phi - \frac{\pi}{2p}\right) \]
\[ M_r = \frac{B_{rem}}{\mu_0} \cos\left(\phi - \frac{\pi}{p}\right) \]
\[ M_\phi = \frac{-B_{rem}}{\mu_0} \sin\left(\phi - \frac{\pi}{p}\right) \]

Here, \( B_{rem} \) is the remanent flux density of the magnets and \( \alpha = \frac{\phi_m}{\frac{\pi}{p} + \phi_r} \) is the radial magnet arc to the total magnet pole arc ratio.

A description of the magnetic fields inside the the magnet and airgap region is obtained by finding solutions for the vector potential which are governed by the Laplace and Poisson equations and by applying the appropriate boundary conditions [7]. The analytical solutions to the flux density inside the airgap are simulated and show good agreement with finite element modeling (FEM) in 2D FLUX, as can be seen in Fig. 9.

For a machine with 1/4 coil per magnet pole per phase the winding arrangement for a machine with moving coils is
shown in Fig. 7. The torque is calculated by using the Lorentz' force law

$$T = \int_V \mathbf{r} \times (\mathbf{J} \times \mathbf{B}_f) dv = \int_V J_z B_r r dV \epsilon_e$$  \hspace{1cm} (12)$$

where $V$ is the volume occupied by the coils, $\mathbf{r}$ is the vector to the point about which the torque is computed, $\mathbf{J}$ is the current density vector and $\mathbf{B}_f$ is the airgap flux density vector, which is assumed to be constant along the axial length of the magnets.

### B. Thermal model

The amplitude of the current density inside the coils is constrained by the temperature distribution inside the actuator. A similar approach like is proposed in [8], is used and modified to predict the temperature distribution inside the rotary actuator. In this approach a thermal equivalent circuit (TEC) of the actuator is created, which offers a fast way to predict the thermal behavior. The actuator geometry is divided in lumped components and the thermal behavior of these components is presented in a network of thermal resistances, capacitances and heat sources.

While the transient thermal behavior of a machine can be simulated using TEC, in this case only the steady-state temperature distribution is of concern, and hence, only the thermal resistances and heat sources need to be determined. The thermal model is further simplified by considering one section of the actuator and assuming only radial heat flow. For the moving coil configuration, Fig. 10 shows the resulting radial heat flow by conduction, convection and radiation. Heat is produced inside the coils and is determined from the ohmic losses. The linear dependency of the electrical resistivity of copper on temperature is accounted for by modeling it as a negative thermal resistance. The conductive heat flow across the actuator is modeled by thermal resistances which are determined for the different cylindrical components. To account for radiation in the TEC, heat flow is linearized and modeled like convection with a heat transfer coefficient $h_r \approx \epsilon_r$, where $\epsilon$ is the emissivity. Convection at inner and outer radii of the actuator is modeled by a heat transfer coefficient, $h_{gap}$.

The resulting TEC is solved as described in [8] and using the dimensions of the initial design of the slotless actuator and the material properties from Table IV, the temperatures inside the coils and magnets as function of the heat transfer function, $h_{gap}$, for both the moving-coil and moving-magnet configuration are shown in Fig. 11. As the figure shows, the temperature drops significantly when $h_{gap}$ is increased from 5 to 30 W m$^{-2}$ K$^{-1}$. Although high values for the heat transfer function can be achieved by forced air cooling [10], $h_{gap}$ is set to a value of 15 W m$^{-2}$ K$^{-1}$ during the optimization procedure. This value was estimated from measurements on the pre-prototype with an air flow of 25 liters per minute through the airgap.

### C. Mechanical model

The performance of the actuator strongly depends on the mechanical load on it. When no external load is considered, the required torque and force to achieve a certain acceleration are linked to the inertia and total mass of the moving parts of both the rotary and linear actuator. While the design of the linear actuator is fixed, the inertia and moving mass of the rotary actuator are determined for the moving-magnet and moving-coil configuration.

### V. DESIGN OPTIMIZATION

The multi-physical framework is implemented in MATLAB and used in the optimization procedure, which is performed with sequential quadratic programming (SQP). In this section the optimization objective, design constraints and results are discussed.
A. Optimization objective

To create an efficient design of the zφ-module the rotary actuator is optimized with the objective to minimize the copper losses inside the rotary and linear actuator combined. These losses are calculated for a third order motion profile. By taking the losses inside the linear actuator into account, the moving mass of the rotary actuator is limited. The objective function can be written as

$$ f(\alpha, \beta, r_i, l_c, l_m, l_w, l_b, L_{act}) = P_{cu,rot} + I_{lin}^2 R_{lin} \tag{13} $$

where $\alpha$, $\beta$, $r_i$, $l_c$, $l_m$, $l_w$, $l_b$ and $L_{act}$ are the optimization variables, which are described in Table III and indicated in Fig. 2 and 7. $I_{lin}$ is the current and $R_{lin}$ is the resistance of the linear actuator, as is given in Table II.

The initial design of the rotary actuator already showed that the specifications given in Table I are easily met within the volume constraints, making this design oversized for the application. Therefore, the design is also optimized with the objective to minimize the volume of the rotary actuator. The objective function can expressed as

$$ f(\alpha, \beta, r_i, l_c, l_m, l_w, l_b, L_{act}) = \pi r_o^2 L, \tag{14} $$

where $r_o$ is the outer radius of the actuator and $L$ is the total length of the coils.

B. Constraints

The design of the rotary actuator is subject to several constraints. First, the volume of the actuator is constraint by the available height inside the zφ-module, the radius of the shaft and the outside radius. Furthermore, a lower bound is set to the radial length of the coils. To guarantee a linear response of the actuator to the injected current, saturation of the steel is avoided. Finally, to prevent damage of the winding insulation, possible irreversible demagnetization of the magnets or substantial loss of performance due to a lower value of the intrinsic magnetization, the coil and magnet temperatures are constrained. Table V lists all the constraints.

C. Optimization results

The optimization is performed for the moving-magnet configuration and moving-coil configuration and Fig. 12 shows the minimized copper losses for different magnet counts. Both configurations have a minimum at $p=14$ and are constraint by the volume. In Fig. 13, the results for optimization with the objective to minimize the volume of the rotary actuator are shown. The figure shows a minimum for $p=18$ and $p=16$ for the moving-magnet and moving-coil configuration, respectively. In this case the minimization is limited by the magnet temperature.

As is already mentioned in Section II, the reduced amplitude of the EMF measurements compared to the 2D semi-analytical magneto-static model, is caused by the magnet edge-effects. For the design of the pre-prototype, the peak airgap flux density along the axial length of the magnets is simulated with 3D FEM, and Fig. 14 shows a comparison to the flux density predicted with the 2D semi-analytical model. A reduction of 10.3 % of the flux linkage by the coils and, hence, a similar reduction of the EMF, is predicted from this analysis. Likewise, the magnet edge-effects are predicted for the optimized designs and the copper losses in the optimized designs are recalculated while the reduced magnetic loading is accounted for. The sizes and specifications of these designs are listed in Table VI for both minimized copper losses and minimized volume.

<table>
<thead>
<tr>
<th>Constraint</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$B_{core}&lt;1.4$ T</td>
<td>Saturation in core</td>
</tr>
<tr>
<td>$B_{back}&lt;1.4$ T</td>
<td>Saturation in back-iron</td>
</tr>
<tr>
<td>$r_i&gt;18$ mm</td>
<td>Minimum inner radius</td>
</tr>
<tr>
<td>$r_o&lt;30$ mm</td>
<td>Maximum outer radius</td>
</tr>
<tr>
<td>$L&lt;37$ mm</td>
<td>Maximum length actuator</td>
</tr>
<tr>
<td>$l_w&gt;0.9$ mm</td>
<td>Radial length coils (windings)</td>
</tr>
<tr>
<td>$T_{co}&lt;100$ °C</td>
<td>Maximum coil temperature</td>
</tr>
<tr>
<td>$T_{m}&lt;60$ °C</td>
<td>Maximum magnet temperature</td>
</tr>
</tbody>
</table>

Fig. 12. Minimized copper losses for different magnet pole pairs.

Fig. 13. Minimized volume for different magnet pole pairs.
With the objective to minimize the copper losses, the optimized design with moving coils produces the lowest losses, which is 30\% less compared to the moving-magnet configuration. This is due to a lower inertia, a smaller moving mass and a higher magnetic loading. On the other hand, from a practical point of view, the electrical wiring and additional mechanical clearance with the moving-coil configuration may be of a concern. As Table VI shows, these two designs occupy the maximum available space inside the \( z\phi \)-module. In the new designs with minimized volume, both the moving-coil and moving-magnet configuration occupy 50\% and 45\%, respectively, less volume than the first two designs. Because the magnet edge-effects were analyzed after the optimization step, the magnet temperature has slightly exceeded its thermal constraint. In this case the moving-magnet configuration is preferred because it has lower copper losses. In all four new designs, it can be noticed that the linear actuator is the dominating mechanical load and can account for 85\% and 87\% of the total inertia and moving mass, respectively.

**VI. Conclusions**

In this paper an optimized design of a rotary actuator, which is coupled to a linear actuator, has been obtained using a multi-physical framework. Only combinations of the number magnet poles and coils resulting in 1/4 coil per magnet pole per phase have been considered during optimization, because, for slotless machines, they have the highest winding factor. Within the available volume, a design with moving coils results in the lowest combined copper losses of the rotary and linear actuator. This design has 30\% less copper losses compared to a configuration with moving magnets. Although only a limited rotational stroke is required, electrical wiring with moving coils may form a point of concern. It is also shown that the same specifications for the \( z\phi \)-module can be met within a 50\% smaller design of the rotary actuator.