Flux-controlled Hybrid Reluctance Actuator for High-precision Scanning Motion

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Abstract—To achieve highly precise and linear scanning motion by a hybrid reluctance actuator (HRA), this paper proposes a flux-controlled mode that uses regulated magnetic flux as the control input for the actuator operation and evaluates its performance in comparison with the conventional current-controlled mode. In the conventional case, HRAs exhibit magnetic nonlinearities (e.g. hysteresis) and position-dependent force that can make the system unstable. A model-based analysis reveals that they are included in the variable magnetic flux of a HRA. Thus, they are estimated by flux estimation and rejected by flux feedback control for high quality scanning motion. For the estimation, sensor fusion with a current monitor and a search coil is used. PI controllers are used for the flux feedback control, as well as for current feedback control of the benchmarking current-controlled mode. During scanning, feedforward control is used to compensate linear dynamics. When sine motions are experimentally tested at 60-300 Hz, the current-controlled mode exhibits a nonlinearity between 6% and 23%, which is decreased to less than 5% by the flux-controlled mode. For a ±75 μm triangular motion at 100 Hz, the flux-controlled mode decreases the tracking error by a factor of 19 to 3.2 μm, successfully demonstrating its high-quality linear scanning motion.

Index Terms—Actuators, Nanopositioning, Magnetic variables control, Motion control.

I. INTRODUCTION

High-precision actuators are used in many motion systems. For example, micro/nanopositioners are installed in manufacturing and imaging systems, such as 3-D printers [1], atomic force microscopes [2], and optical profilers [3]. Another example is fast steering mirrors, which are utilized to scan optical beams for free-space optical communication [4] and 3-D imaging [5]. For compact size with a relatively large motion range (more than tens of micrometers and a few milliradians [6]), Lorentz actuators are typically selected and guided by flexures in these high-precision motion systems.

Lorentz actuators, including voice coil actuators, use the Lorentz force [7], which is bidirectional unlike the electrostatic force [8] and the reluctance force [9]. Lorentz actuators are highly linear for the proportionality between the input coil current and the resulting force [10], which is independent of the mover position within a uniform flux density (i.e. zero-stiffness property [7]). Such characteristics are ideal to generate linear motion in comparison with other actuators (e.g. piezoelectric actuators with hysteresis [11]). Additionally, when Lorentz actuators are guided by flexures for compactness, their closed-loop control bandwidth is significantly higher than the suspension mode’s eigenfrequency by design [12]. This enables a large actuation range [13] and good vibration rejection [12]. As a result, flexure-guided Lorentz actuators realize even nanometer positioning resolution in vibrational environments [14]. However, Lorentz actuators have a relatively small motor constant (i.e. force-to-current ratio) [7] and need a large coil current and cooling mechanism to generate a high force, wasting energy in the form of heat [10], [15]. Thus, for better energy efficiency, actuators with a high motor constant and high linearity are desired in high-precision systems.

Hybrid reluctance actuators (HRAs) have a high potential for the desired performance. They use both coils and permanent magnets to create flux in ferromagnetic yokes [16]. HRAs generate a bidirectional force that is relatively proportional to the coil current and are usually regulated in the current-controlled mode [16], [17]. The motor constant of HRAs can be higher than comparable Lorentz actuators [7], and HRAs are utilized in fast steering mirrors [16], [18], fast tool servers [19], and nanopositioners [17]. However, HRAs create position-dependent force (i.e. no zero-stiffness property). It is regarded as a negative stiffness, and guiding mechanisms such as flexures with a sufficient stiffness are required for open-loop stability [17]. More critically, these actuators exhibit nonlinearities such as hysteresis and eddy currents [20], [21], in addition to the variation of the negative stiffness [17]. Such nonlinearities can be compensated by motion control (e.g. position feedback control, repetitive control, and learning control [17], [22]). However, the nonlinearities deform the scanning trajectories, resulting in a tracking error at high frequencies [23]. Consequently, the motion control’s bandwidth needs to be significantly higher than the frequency components of the motion reference. This is problematic particularly for high-frequency scanning motions because the achievable bandwidth can be limited by mechanical internal modes. Furthermore, when a position sensor is used for the motion control, the positioning resolution is degraded by feeding its noise back to the actuator at high frequencies.

For highly linear scanning motion of a HRA without the above problems, this paper proposes a new operating mode for...
HRAs: the flux-controlled mode, where controlled magnetic flux is used as the control input of a HRA. The model-based analysis in this paper reveals that the magnetic nonlinearities and the position-dependent flux are included in the variable flux selected as the control input. Consequently, they can be compensated by a flux feedback controller without a position sensor. For the feedback control, the variable flux is estimated at high resolution by utilizing sensor fusion with a search coil. The proposed flux-controlled mode is intended to replace the conventional current-controlled mode. To analyze the effectiveness of the flux-controlled mode clearly, it is experimentally compared with the current-controlled mode, without any other remedy to handle the actuator nonlinearities. The results demonstrate that even a simple PI controller significantly improves the linearity of the mover motion and stabilizes the HRA by compensating for the position-dependent force.

Flux measurement, estimation, and control have been themselves utilized for electromagnetic actuators [24]–[26]. In the case of reluctance actuators, their force is proportional to the squared flux, and this nonlinearity complicates the analysis and motion control design, or a biasing current is required, impairing the energy efficiency [7], [27], [28]. In contrast, this paper reveals that the proposed flux-controlled HRA generates a force proportional to the variable flux without a biasing current, which is desired for highly precise and accurate motion with high energy efficiency.

This paper is organized as follows. Section II introduces the proposed flux-controlled HRA. It is modeled in comparison with the conventional current-controlled mode in Section III. Section IV presents flux estimation for flux feedback control in Section V, which also discusses current feedback for comparison. The flux and current reference signals are designed to generate scanning motions in Section VI for experiments in Section VII. Section VIII concludes the paper.

II. HARDWARE ARCHITECTURE

Figure 1(a) and (b) show the employed HRA. The stator consists of a ferromagnetic yoke with two identical actuation coils and a Nd-Fe-B permanent magnet. The ferromagnetic mover is guided by leaf-spring aluminum flexures. The ferromagnetic stator and mover are made of laminated electrical steel sheets (EN10025-S235JR) to reduce eddy current losses for energy efficiency. The mover carries a cube-corner retroreflective (43-305, Edmund optics, Barrington, USA) for measuring its position with an interferometer with a resolution of 1.25 nm/bit (10899A, Agilent Technologies, Santa Clara, USA).

The actuation coils are connected in series and used with a custom-made voltage amplifier that includes a current monitor with a shunt resistor to measure the coil current. Additionally, a search coil [29] used with a preamplifier is attached to the stator for flux estimation. The current monitor and the preamplifier are connected via 16-bit ADCs (DS2004, dSPACE), a rapid prototyping system (DS1005, dSPACE), and an FPGA (DS2023, dSPACE), respectively. Fig. 1(a) additionally shows a Hall sensor (CYTHS124, ChenYang, Finsing, Germany) with a spacer in the left variable gap. Since the sensor in the gap interferes with the mover and limits the motion range, it cannot be used for real-time control, but for system identification and calibration of the flux estimator. During scanning, the Hall sensor and spacer are removed.

III. MODELING

The HRA utilizes unbalanced flux for actuation. To derive the flux for modeling, it is assumed that the yokes’ permeability is sufficiently large. The magnetic reluctances of the left variable gap $R_l$, the right variable gap $R_r$, the fixed gap $R_f$, and the permanent magnet $R_m$ shown in Fig. 1(b) are

$$R_l = \frac{x_g - x}{\mu_0 A}, \quad R_r = \frac{x_g + x}{\mu_0 A}, \quad R_f = \frac{1}{\mu_0 A}, \quad R_m = \frac{1}{\mu_0 A_m},$$

where $\mu_0$ and $A$ are the vacuum permeability and the cross section area of the flux paths, respectively, and $A_m$ denotes the magnet’s cross section area. The left and right variable gaps are denoted by $x_g$ when the mover is at the center, which the mover position $x$ is measured from. The length of the fixed gap and the permanent magnet is denoted by $l_f$ and $l_m$, respectively. Table I lists the nominal design parameters.
Because the magnet is regarded as a large air gap in (1), the flux generated by the actuation coils’ current $I$ does not go through the magnet and is given by

$$\Phi_c = 2NI/(R_t + R_c) = \mu_0 AN I/x_g,$$  \hspace{1cm} (2)

where $N$ is the number of windings of each actuation coil. Since $l_m$ is way larger than $l_f$ and $x_g$, the flux given by the magnet is approximated as

$$\Phi_m = \frac{H_m l_m}{R_m + R_c + R_c R_t/R_m} \approx \frac{H_m l_m}{R_m}, \hspace{1cm} (3)$$

where $H_m$ is the coercive force of the magnet. By superimposing $\Phi_c$ and $\Phi_m$, the flux through the left variable gap $\Phi_l$ and the right variable gap $\Phi_r$ are given by

$$\Phi_l = \Phi_c + \Phi_m R_c/(R_t + R_c) = \Phi_c + \Phi_m/2,$$ \hspace{1cm} (4)$$

$$\Phi_r = \Phi_c - \Phi_m R_c/(R_t + R_c) = \Phi_c - \Phi_m/2,$$ \hspace{1cm} (5)

where $\Phi_c$ is the variable flux defined by

$$\Phi_c = \Phi_c + \Phi_m x/(2x_g). \hspace{1cm} (6)$$

As illustrated in Fig. 1(b), $\Phi_l$ and $\Phi_r$ consist of the constant flux $\Phi_m/2$ and the variable flux $\Phi_c$, which depends on $x$ and $I$.

**A. Flux-controlled HRA**

Eq. (4) and (5) indicate that the flux in the left variable gap and in the right variable gap is unbalanced by $\Phi_c$, resulting in the actuation force $F$. In the flux-controlled mode, the control input is flux, and $F$ is written as a function of $\Phi_c$ by using the Maxwell stress tensor [21] as follows

$$F = \frac{(\Phi_c + \Phi_m/2)^2 - (\Phi_c - \Phi_m/2)^2}{2\mu_0 A} = K_m f^{\mu_0}, \hspace{1cm} (7)$$

where $K_m f^{\mu_0} = \Phi_m/\mu_0 A$ is the motor constant of the flux-controlled HRA. Unlike conventional modeling [30], Eq. (7) clearly indicates the proportionality between $F$ and $\Phi_c$. This is an important property of the proposed flux-controlled HRA for bidirectional force and linearity, in contrast to reluctance actuators, the force of which is proportional to squared flux.

Since the mover is guided by the flexures, it is modeled by a damped-mass spring system [17], resulting in the equation of motion

$$F = K_m f^{\mu_0} \phi_c = m \ddot{x} + c \dot{x} + k x, \hspace{1cm} (8)$$

with the mover mass $m$ and the damping $c$ and stiffness $k$ of the flexures. Its Laplace transform gives a transfer function from $\phi_c$ to $x$

$$P_f(s) = \frac{x(s)}{\phi_c(s)} = \frac{K_m f^{\mu_0}}{ms^2 + cs + k}. \hspace{1cm} (9)$$

**B. Comparison with current-controlled mode**

For comparison with the current-controlled mode, (7) is rewritten as a function of $I$ by using (2)-(6) as follows [16], [19]

$$F = K_m c I + k_a x, \hspace{1cm} (10)$$

where $K_m c$ and $k_a$ are the motor constant and the actuator stiffness of the current-controlled HRA. The equation of motion of the lumped mass model with (10) gives the transfer function [17]

$$P_c(s) = \frac{x(s)}{I(s)} = \frac{K_m c}{ms^2 + cs + k - k_a} \hspace{1cm} (11)$$

Fig. 1(c) visualizes the derived models. The problem in the current-controlled mode is that the nonlinear dynamics due to the B-H hysteresis and the residual eddy currents [21] occur between $I$ and $\phi_c$. They cannot be compensated by the current feedback controller $G_c(s)$ since they are outside of the current control loop. As a result, $K_m c$ in (11) varies in reality, making the current-controlled actuator nonlinear. Furthermore, current control cannot compensate for the position-dependent force $k_a x$ in (10), and this out-of-control force cancels $k$ in (11).

When the mover is far away from the center, $k_a$ increases due to the flux leakage [31] and the magnet’s flux that are influenced by $x$ [16]. In the extreme case of $k_a > k$, an unstable pole occurs in (11), and the actuator is unstable [17]. In contrast, the control input $\phi_c$ of the proposed flux-controlled mode includes the nonlinearities and the position-dependent flux in (6). Consequently, they are compensated by monitoring and controlling $\phi_c$, as discussed in the following sections. Note that the switch in Fig. 1(c) is only for the comparison of the two modes and is not toggled during the scanning operation.

**IV. Flux estimator**

**A. Estimator model**

For system identification, the Hall sensor is temporary installed to measure the flux ($\phi_c + \phi_m/2$) in the left variable gap. Since $\phi_m/2$ is ideally invariant, by removing the DC offset (zeroing the sensor), its output $y_h$ approximates $g_h \phi_c$, where the sensor gain $g_h$ is 2.5 V/T. Since the sensor in the gap reduces the motion range, the Hall sensor is removed, and $y_s$ is estimated by a flux estimator $G_e(s)$ based on the search coil and the current monitor during real-time control for high-quality motion (Fig. 1(c)). In this section, $G_e(s)$ is designed, as shown in Fig. 2.

The output $y_h$ of the search coil used with its preamplifier is given by Lenz’s law [29], and it is $y_s = g \phi_c$, where $g$ is a gain determined by the preamplifier gain and the number of the search coil’s windings that is 16. From the Laplace
transform of $y_h$, an estimator $G_{ec}(s)$ to estimate $y_h$ from $y_s$ is given by

$$G_{ec}(s) = \frac{\hat{y}_h(s)}{y_s(s)} = \frac{g_{ec}}{s}$$  
(12)

where $g_{ec}$ is an estimator gain. While the above integrator filters the high-frequency measurement noise for precise estimation, its high gain degrades the signal-to-noise ratio at low frequencies. Furthermore, $G_{ec}(s)$ is not bounded-input bounded-output (BIBO) stable, and its output signal drifts with $s$ if $x$ is sufficiently small in that frequency domain, $\Phi_e$ approximates $\Phi_s$ in (6), and an estimator is given from (2) for the current monitor output $y_t$ by

$$G_{ec}(s) = \frac{\hat{y}_h(s)}{y_t(s)} \approx \frac{2\Phi_e}{\Phi_s} \frac{g_{ec}}{s + \omega_c}$$  
(13)

where $g_o$ and $g_{ec}$ are the current monitor gain (1 V/A) and an estimator gain, respectively. The current-based flux estimator cannot capture the magnetic nonlinearities but is BIBO stable.

For the sensor fusion of $G_{ec}(s)$ and $G_{es}(s)$, a complementary filter that consists of first-order low-pass and high-pass filters [25], [31] is used (Fig. 2), and the flux estimator is given from (12) and (13) by a dual-input single-output transfer function $G_c(s)$:

$$\hat{y}_h(s) = \frac{g_{ec}\omega_c}{s + \omega_c} y_s(s) + \frac{g_{ec}}{s + \omega_c} y_t(s),$$  
(14)

where $\omega_c$ is the cut-off frequency to be determined. Notice that the above filtering is BIBO stable.

B. Parameter determination

To correctly identify $g_o$ and $g_{ec}$ in (12)-(14), frequency response functions (FRFs) are measured from the voltage output, reference search coil $V_r$ to $y_r$, $y_s$, and $y_h$, individually. The results are used to select $G_{ec}(j\omega)$ and $G_{es}(j\omega)$ by calculating

$$G_{ec}(j\omega) = \frac{G_{ec}(j\omega)}{V_r(j\omega)} \quad G_{es}(j\omega) = \frac{G_{es}(j\omega)}{V_r(j\omega)}$$  
(15)

as shown in Fig. 3. By tuning $g_{ec}$ and $g_{es}$, the transfer function $G_{ec}(s)$ and $G_{es}(s)$ are fit to the measured FRFs. The results with $g_{es} = 27.4\, \text{dB}$ and $g_{ec} = -13.9\, \text{dB}$ are also shown in Fig. 3. It is visible that $G_{ec}(s)$ and $G_{es}(s)$ capture their FRFs at least up to 1 kHz and 20 Hz, respectively. The mismatch at frequencies higher than a few kilohertz may be due to the parasitic dynamics of the actuation coils [32].

Fig. 4(a) shows the power spectral density (PSD) of the flux $\hat{y}_h(s)$ estimated by the measured $y_r$, and $G_{es}(s)$ and by the measured $y_s$ and $G_{es}(s)$ when $V_r$ is set to zero. For a fair comparison, the PSD is scaled to $\text{mT}/\sqrt{\text{Hz}}$. The noise of the flux estimated by $G_{es}(s)$ and $y_s$ is higher than that by $G_{es}(s)$ and $y_s$ below 10 Hz. To filter the low-frequency noise of $y_s$, $\omega_c$ is set to 10 Hz. Fig. 4(a) also shows the measured PSD of the resulting estimator $G_c(s)$ in (14). It validates that the estimator noise is successfully reduced to the level of $G_{ec}(s)$ below 10 Hz. The effectiveness is significant in the time domain, as shown in Fig. 4(b), where $\hat{y}_h$ estimated by $G_{ec}(s)$ shows less noise of 8.7 $\mu\text{T}_{\text{rms}}$ than that by $G_{ec}(s)$ ($57\, \mu\text{T}_{\text{rms}}$) without drift. Additionally, measured Bode plots from $V_r$ to $y_h$ and to $\hat{y}_h$ estimated by $G_{es}(s)$ are compared in Fig. 5, which clearly shows that $G_{es}(s)$ is capable of estimating $y_s$ up to about 3 kHz. Note that the dynamics in Fig. 5 have low-pass characteristics due to the impedance of the actuation coils denoted by $Z_o(s)$ in Fig. 1(c).

V. FEEDBACK CONTROL FOR FLUX- AND CURRENT-CONTROLLED MODES

Since the flux estimation error increases above 3 kHz in Fig. 5, a flux feedback controller $C_f(s)$ (see Fig. 1(c)) for the flux-controlled mode is designed for an open-loop cross-over frequency of 3 kHz with a sufficient phase margin (PM) of 55 deg. For this purpose, a PI controller is implemented by

$$C_f(s) = \frac{K_{pf} + j\omega_c}{s},$$  
(16)

The P gain $K_p$ and the corner frequency $\omega_c$ are set to 50.8 dB and 300 Hz to satisfy the requirements by using a simulated open-loop transfer function based on the measured response from $V_r$ to $\hat{y}_h$ in Fig. 5. Fig. 6(a) shows the simulated plot, indicating that the requirements are satisfied.

Similarly, another PI controller $C_c(s)$ (see Fig. 1(c)) is designed for the current-controlled mode. The red dashed line...
Fig. 5. Bode plots from $V_c$ to $y_h$ measured by the Hall sensor and estimated by the implemented $C_c(s)$ for validation.

Fig. 6. Simulated open-loop transfer function: (a) for flux feedback control design with the measured Bode plot from $V_c$ to $y_h$ as the plant and (b) for current feedback control design with the measured Bode plot from $V_c$ to $y_h$ as the plant. The arrows indicate the open-loop cross-over frequency and the phase margin.

in Fig. 6(right) shows a Bode plot from $V_c$ to $y_h$ as the plant. Notice that the gain increases at high frequencies due to the parasitic dynamics. To decrease the open-loop gain at high frequencies for stability, a low-pass filter is included in $C_c(s)$ as follows

$$C_c(s) = \frac{1}{s} \left( k_{pm} + k_{pc} \omega_c^2 s^2 / \alpha_c s^3 \right),$$

where $k_{pm}$, $\omega_c$, and $\alpha_c$ are the P gain, the desired cross-over frequency, and a parameter to tune PM, respectively. For a fair comparison with the flux control, $k_{pc}$, $\omega_c$, and $\alpha_c$ are tuned to 31.2 dB, 3 kHz, and 2.1, respectively, such that the open-loop cross-over frequency and PM are also 3 kHz and 55 deg, as shown in Fig. 6(b).

For evaluation, the complementary sensitivity functions from $r_s$ to $y_h$ and from $r_f$ to $y_h$ (see Fig. 1(c)) are measured as shown in Fig. 7(a). The current-controlled and flux-controlled modes show similar behavior as desired, with a (-3 dB)

bandwidth of 5.4 kHz and 6.0 kHz, respectively. Additionally, 100 Hz sinusoidal references are used to evaluate Lissajous curves. Fig. 7(b) demonstrates that the flux estimator $G_c(s)$ captures the hysteresis of the current-controlled mode. It is compensated by the flux-controlled mode in Fig. 7(c). When the trend lines in Fig. 7(b)(c) are used to calculate nonlinearity (mapping errors divided by scan ranges [33]), the nonlinearity due to the hysteresis is 6.8% and 1.1% for the current-controlled and flux-controlled modes, respectively.

VI. MOTION CONTROL

A. Trajectory

To evaluate scanning motions, sine waves are used, as well as a band-limited triangular wave, which is commonly used for imaging systems (e.g. [14]). These waves are periodic and given in the form of Fourier series

$$r_x(t) = \sum_{k=1}^{\infty} (c_k e^{j k \omega_s t} + c_k^* e^{-j k \omega_s t}),$$

where $t$, $l$, $c_k$, and $c_k^*$ are the time, the highest harmonic, the complex Fourier coefficient of the $k$-th harmonic, and its conjugate, respectively. As discussed in Section IV-A, the scanning frequency $\omega_s$ needs to be sufficiently higher than $\omega_c = 10$ Hz of the flux estimator $G_c(s)$, where the current-based flux estimation dominates without capturing the nonlinearities. Note that (18) has no DC component for the scanning motion evaluation. For sine waves with an amplitude of $A_x$, $r_x(t)$ consists of $c_1 = -j A_x/2$ and $c_k^*$ only. In the case of the triangular wave, the coefficients are given by

$$c_k = -j A_x / (\pi k^2) \sin \left( \frac{k \pi}{2} \right).$$

For the triangular scanning at 100 Hz, $A_x$ and $l$ are set to 75 $\mu$m and 7, respectively.
When a sine wave is used as the reference \( r_f \), the mover motion of \( A_f \) is increased to 65 Hz, which is about one third of the flux-controlled actuator’s resonance (193 Hz), the nonlinearity is 4.7 % only, demonstrating the improved linear motion. It is clear that the tracking error improvement by the flux-controlled mode is more significant for the 30 \( \mu \)m trajectory than for the 1 \( \mu \)m. To analyze this amplitude dependency, the normalized error in rms \( e_x/A_f \) is measured for \( \omega_e = 100 \) Hz when \( A_f \) is increased from 1 \( \mu \)m to 200 \( \mu \)m. Fig. 9(c) shows that the current-controlled mode significantly increases the normalized error as \( A_f \) increases. Particularly for \( A_f = 200 \) \( \mu \)m, the mover collides with the stator. Such phenomena can be explained by the actuator stiffness \( k_0 \) of the current-controlled HRA in (10). When the mover goes away from the origin, \( k_0 \) increases and changes the plant model (11) [17], resulting in a mismatch with \( P_{rc} \) of the feedforward control (21) and in a larger tracking error. Eventually, the actuator can be unstable due to the uncontrollable positive flux feedback loop in Fig. 1(c).

As discussed in Section III, the problematic position-dependent flux is regarded as a disturbance in the case of the flux-controlled mode and can be rejected by the flux feedback controller. The experiments in Fig. 9(c) verify the model-based analysis. In the the flux-controlled mode, the normalized error is less than 8.3 % over the entire amplitude range, and the stable actuator operation is realized even for \( A_f = 200 \) \( \mu \)m.

B. Triangular scanning motion

The linear sine motions of the flux-controlled actuator in Section VII-A imply its capability to realize an arbitrary periodic scanning motion by superimposing them. This capability is demonstrated by using the triangular trajectory designed in Section VI-A. Fig. 10(a)(b) shows that the current-controlled mode generates a significantly larger motion than the desired motion of ±75 \( \mu \)m. The tracking error \( e_x \) changes between -116 \( \mu \)m and 93 \( \mu \)m, resulting in an rms value of 60 \( \mu \)m. The flux-controlled mode decreases this large rms error by a factor of 19 to 3.2 \( \mu \)m.

More importantly, the spectrum of the tracking error in Fig. 10(c) shows that the current-controlled mode’s \( e_x \) has frequency components up to about 2 \( kHz \), which is far higher than the highest harmonic frequency 700 Hz of the reference \( r_e \). This is because the nonlinearities deform the scanning motion and create the higher harmonics. They are unwanted as they may excite mechanical internal modes. In contrast, the error spectrum in the flux-controlled mode is up to 800 Hz and stays at the noise floor level at higher frequencies, due to its high linearity. This property is more important when the scanning motion is further improved by combining motion.
control with a position sensor (e.g., position feedback control, repetitive control, and learning control [17], [22]). For the compensation of the residual tracking error, the motion control needs to have a bandwidth of 2 kHz in the current-controlled mode, which is significantly decreased to 800 Hz in the flux-controlled mode. This relaxed requirement in the flux-controlled mode is beneficial for high-precision motion, reducing the measurement noise that is fed back to the HRA by the motion control.

In summary, the proposed flux-controlled HRA realizes more stable operation, higher linearity, and smaller tracking error by regulating the variable flux, in comparison with the conventional current-controlled mode.

VIII. CONCLUSION

This paper proposes a flux-controlled HRA by using its variable flux as the control input. A model-based analysis reveals that the position-dependent flux and the magnetic nonlinearities are included in the control input. To capture them, sensor fusion is used with a current monitor and a search coil to precisely and accurately estimate the variable flux without drift in a targeted scanning frequency band. The captured nonlinearities are rejected by flux feedback control, which is implemented by a PI controller. For the benchmarking current-controlled mode, a low-pass-filtered PI controller is designed for fair comparison. Furthermore, the reference signals are individually designed to compensate for the linear dynamics. The experiments demonstrate that the nonlinearity of sine scanning motions is up to 23% in the current-controlled mode while it is less than 5% in the flux-controlled mode. When the ±75 μm triangular reference is used for 100 Hz scanning, the current-controlled mode results in the large distorted tracking error of 60 μm. The flux-controlled mode decreases it to 3.2 μm, demonstrating its improved linearity achieved by the regulated flux control input. Future work includes accurate flux estimation at low frequencies, in order to expand the applications of the highly linear HRAs in the flux-controlled mode.

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Fig. 9. Measured nonlinearity of x and tracking error $e_x$ for varying the flux-controlled and current-controlled modes, for sine motions with an amplitude $A_x$ of (a) 1 μm and (b) 30 μm. The plot (c) shows the relation between $A_x$ and the normalized error in rms $e_{x/A_x}$ for the scanning frequency $\omega_x$ of 100 Hz. The tracking error for $A_x = 200$ μm cannot be measured in the current-controlled mode because of a mover collision with the stator.

Fig. 10. Measured 100 Hz triangular motions in the flux-controlled mode and in the current-controlled modes: (a) measured position $x$, (b) tracking error $e_x$, and (c) the spectrum of $e_x$.
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