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Implementation of a Linear Quadratic Bumpless Transfer Method to a Magnetically Levitated Planar Actuator with Moving Magnets

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Abstract—A linear quadratic bumpless transfer technique for a six-degree-of-freedom planar actuator with integrated active magnetic bearing is proposed to solve the lift-off and landing problem by switching between different controller sets. The need for multiple controller operation states results from the difference between robustness and wind-up concerns during lift-off and landing and the high performance tracking requirements for accurate operation. A procedure for tuning the plant input and error matrices used in the Linear Quadratic Bumpless Transfer method based on new observations is presented. The procedure has been applied to synthesize a bidirectional bumpless switching mechanism that was successfully applied and tested on a 6 DOF laboratory setup.

I. INTRODUCTION

A. Planar actuator

Magnetically levitated planar actuators are developed as alternatives to xy drives constructed of stacked linear motors. The translator of these actuators is suspended above the stator with no support other than magnetic fields. The gravitational force is fully counteracted by the electromagnetic force. The translator of these ironless planar actuators can move over relatively large distances in the xy plane only, but it has to be controlled in six degrees of freedom (6-DOF) because of the active magnetic bearing. Magnetically levitated planar actuators can operate in vacuum, for example, in extreme-UV lithography equipment. Planar actuators can be constructed in two ways. The actuator has either moving coils and stationary magnets [1] or moving magnets and stationary coils [2]–[3]. The last type of planar actuator does not require a cable to the moving part. In this paper, the Herringbone Pattern Planar Actuator [4] (HPPA), a 6-DOF controlled long-stroke moving-magnet planar actuator, is considered. Fig. 1 shows a photo of this actuator. The stator of this motor consists of 84 coils with concentrated windings of which only 24 are simultaneously energized. The moving part is a permanent-magnet array having a Halbach magnetization.

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Fig. 1. HPPA developed at Eindhoven University of Technology.

During movements in the xy plane, the set of active coils changes with the position of the translator because only the coils below and near the edge of the magnet array can produce significant force and torque. The nonlinear nature of the plant and the high performance control requirements necessitate advanced control methods to be implemented.

B. Problem statement

The moving magnet planar actuator consists of a permanent magnet array levitated in a coil-current controlled magnetic field. Earnshaw[5] proved that passive levitation with static magnetic fields is not possible. Therefore, feedback control needs to be applied to stabilize the magnet plate position. The systems highly nonlinear position dependence is decoupled by feedback linearization which results in a set of six meta stable LTI-SISO systems [6]. Those can be stabilized by six SISO LTI controllers, illustrated in Fig. 3.

The plant has multiple states of operation each imposing different constraints and requirements for the controller.

For the HPPA three modes of operation can be distinguished:

1) startup or lift-off,
2) tracking or performance mode
3) shutdown or landing.

Transition among the states of operation was realized with a method that is limited to PID control only. It involves a smooth change of parameters that realizes the different controllers. The current controller system consists of a low bandwidth PD lift-off controller, a high bandwidth tracking
PID controller and a landing PD controller. The switching between these controllers is bumpless because the tracking controller can smoothly vary its bandwidth and modifies the integral action by pole zero cancellation. Therefore, at the moment of switching, the parameters of both controllers are equal, which guarantees a bumpless transition [7].

To apply advanced control the need arises to switch bumpless between controllers of higher order and with preferably equal, which guarantees a bumpless transition [7].

Many different bumpless switching solutions exist, however the majority of them is restricted to the PID controller type [8], [9]. These methods however, as well as the dynamic bumpless transfer method restrict the controllers to be bi-proper. The linear quadratic approach of Turner and Walker [11] appeared to be a good candidate, since it has all the properties that are required.

This paper deals with the implementation of this LQ approach to the switching among operation states of the moving magnet planar actuator. Numerical problems were encountered and a design rule to overcome them is proposed.

II. METHOD

Many techniques are based upon the Hanus method [8], which requires bi-proper controllers which is undesired for actual implementation. Moreover, most techniques are controller specific, so changing the controllers that are switched, requires the bumpless switching to be redesigned. In this paper the bumpless switching technique of Turner and Walker [11] is selected for implementation. To give some understanding of this method a short summary of [11] is given in this section.

To avoid confusion some terminology will be introduced. Throughout the paper we will distinguish between the controller that is actively controlling the plant and the non-active controller we want to switch to. They will be called on-line plant controller and off-line plant controller, respectively. The label plant controller is used because to switch bumpless another controller is introduced, the tracking controller, $F$. This tracking controller forces the off-line plant controller to track the output of the on-line plant controller. This is also depicted in Fig. 4.

Although the method presented in [11] allows the plant controllers to be strictly proper, it still posses other requirements. The controllers have to be finite dimensional, linear and time invariant (FDLTI), and the off-line controller states should be available at all times. The off-line controller realization has to be completely controllable and observable. The inputs and outputs of the on-line controller have to be available for the off-line controller. A quite modest assumption as most controllers are realized in software.

The bumpless switching problem is treated as a requirement on the controller outputs, whose values should not differ too much. However, the controller that is switched to could also exhibit unwanted dynamics due to a bump in its input. Therefore, its input should be close to the error signal at the time of switching. These requirements can be formulated in the linear quadratic context. One may express this problem in terms of the following cost function,

$$
J(u, \alpha, T) = \frac{1}{2} \int_0^T z_a(t)^T W_a z_a(t) + z_e(t)^T W_e z_e(t) dt + \frac{1}{2} z_a(T)^T P z_a(T),
$$

(1)

where

$$
z_a(t) = u(t) - \bar{u}(t),
$$

(2)

$$
z_e(t) = \alpha(t) - \bar{e}(t),
$$

(3)

with $\bar{u}(t)$ and $\bar{e}(t)$ the on-line control signal and error signal respectively, $u(t)$ the off-line control signal and $\alpha(t)$ the signal produced by the feedback gain which drives the off-line controller. $W_a$ and $W_e$ are constant positive-definite plant input difference and plant output difference weighting matrices, respectively, which give flexibility in the design. Fig. 4 illustrates the bumpless switching scheme for a one degree of freedom (1-DOF) controller.

Only the most important steps that are required to synthesize the mechanism are presented. The off-line controller driven by the signal $\alpha$ can be represented in state space form as,

$$
\dot{x} = Ax + B\alpha,
$$

(4)

$$
u = Cx + D\alpha.
$$

(5)
Substituting $u$ from (5) in the performance index (1) and using (4) as the constraint, one arrives at a standard minimization problem which can be reformulated by introducing a dynamic Lagrange multiplier $\lambda$. This leads to equations in terms of the state $x$ and the co-state $\lambda$:

$$
\begin{pmatrix}
\dot{x} \\
\dot{\lambda}
\end{pmatrix} = 
\begin{pmatrix}
\tilde{A} & \tilde{B} \\
\tilde{C} & \tilde{D}
\end{pmatrix}
\begin{pmatrix}
\dot{x} \\
\dot{\lambda}
\end{pmatrix} + 
\begin{pmatrix}
-B\Delta W_u \\
-C^TW_e\Delta W_e
\end{pmatrix}\tilde{e} + 
\begin{pmatrix}
\tilde{C} \Delta W_u(I+D\Delta^\top W_e) \\
\tilde{C}^\top W_e(I\Delta W_u)
\end{pmatrix}\tilde{u},
$$

(6)

where

$$
\begin{align*}
\tilde{A} &= A + B\Delta D^\top W_u C, \\
\tilde{B} &= B\Delta \tilde{B}, \\
\tilde{C} &= C^\top W_u(I + D\Delta D^\top W_u)C, \\
\Delta &= -(D^\top W_u D + W_e)^{-1}.
\end{align*}
$$

The system matrix defined in (6) will be denoted in this paper as $H$. This matrix contains the stable and anti-stable eigenvalues of the closed loop consisting of the off-line controller and the static feedback gain $F$. By choosing the weighting matrices $W_u$ and $W_e$ one can influence the location of these eigenvalues. Differential equation (6) is of the form which often arises in LQ minimization and can be solved by the method of sweep [12]. This leads to the differential Riccati equation, which, considering only strictly proper off-line controllers, results in the following algebraic Riccati equation,

$$
\Pi \tilde{A} + \tilde{A}^\top \Pi + \Pi \tilde{B} \Pi + \tilde{C} = 0.
$$

(8)

The numerical conditioning of the above Riccati equation depends on the distribution of the eigenvalues of the $\tilde{A}$, $\tilde{B}$ and $\tilde{C}$ matrices. Consequently, this influences the numerical conditioning of the synthesis of the bumpless mechanism, which will be discussed in the following section.

Finally, it is shown in [11] that, when assuming the exogenous signals to be constant over a certain period of time, the signal driving the off-line controller $\alpha$ can be expressed as a function of the signals $x$, $\tilde{u}$ and $\tilde{e}$:

$$
\alpha = F \begin{pmatrix} x \\ \tilde{u} \\ \tilde{e} \end{pmatrix},
$$

(9)

where $F$ is given in terms of the controller state space representation $(A, B, C, D)$, $\Pi$ and the weighting matrices $W_u$ and $W_e$.

$$
F = \Delta \left( \begin{pmatrix} \Pi + D^\top W_u C \\ \Pi B \end{pmatrix}^\top \begin{pmatrix} B^\top \Pi + D^\top W_u C \\ \Pi B \end{pmatrix} + \Pi D \Pi^\top \begin{pmatrix} B^\top \Pi + D^\top W_u C \\ \Pi B \end{pmatrix}^\top \begin{pmatrix} B^\top \Pi + D^\top W_u C \\ \Pi B \end{pmatrix}^\top \right)^\top,
$$

(10)

with $M = (\tilde{A}^\top + \Pi \tilde{B})^{-1}$.

III. IMPLEMENTATION

To apply this method to the planar actuator, it has to be implemented on the modular multiprocessor dSPACE system containing two DS1005 PPC boards and several interfacing boards which is used to control the setup. The solution to the LQ problem, will render the feedback $F$ which needs to be implemented in the realtime application. Although solving the LQ problem is computationally intensive, it can be solved off-line. The matrix manipulations and the solution to the Algebraic Riccati Equation have been obtained using Matlab with the Control System toolbox are implemented on the modular multiprocessor dSPACE system and $\tilde{H}$ conditioning of the synthetization of the bumpless mechanism, which will be discussed in the following section.

Finally, it is shown in [11] that, when assuming the exogenous signals to be constant over a certain period of time, the signal driving the off-line controller $\alpha$ can be expressed as a function of the signals $x$, $\tilde{u}$ and $\tilde{e}$:

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\alpha = F \begin{pmatrix} x \\ \tilde{u} \\ \tilde{e} \end{pmatrix},
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where $F$ is given in terms of the controller state space representation $(A, B, C, D)$, $\Pi$ and the weighting matrices $W_u$ and $W_e$.

$$
F = \Delta \left( \begin{pmatrix} \Pi + D^\top W_u C \\ \Pi B \end{pmatrix}^\top \begin{pmatrix} B^\top \Pi + D^\top W_u C \\ \Pi B \end{pmatrix} + \Pi D \Pi^\top \begin{pmatrix} B^\top \Pi + D^\top W_u C \\ \Pi B \end{pmatrix}^\top \begin{pmatrix} B^\top \Pi + D^\top W_u C \\ \Pi B \end{pmatrix}^\top \right)^\top,
$$

(10)

with $M = (\tilde{A}^\top + \Pi \tilde{B})^{-1}$.

Fig. 5. Eigenvalues of the matrix $H$, using the balanced statespace representation of (12) and $W_u = 50 \cdot 10^{-6}$ and $W_e = 3.5 \cdot 10^{6}$
ill conditioning, however proper selection of the weighting matrices is needed to resolve this problem.

The eigenvalues of $H$ can be influenced by the selection of the weighting matrices. A first selection should be based upon the physical quantities, the errors in the order of micrometers and the plant inputs in the order of decaNewtons. Fig. 5 shows the eigenvalues of the Hamiltonian constructed for the balanced statespace representation of the controller $C_2 (12)$ used in simulations of section IV with $W_u = 50 \cdot 10^{-6}$ and $W_c = 3.5 \cdot 10^3$. To resolve the numerical problems an engineering approach is applied to tune $W_u$ and $W_c$ to their final values:

1) Choose the weighting matrices so that the order of magnitude of the eigenvalues of $\hat{A}$, $\hat{B}$ and $\hat{C}$ are similar.
2) Perform simulations and compare controllers input and output. Increase $W_u$ until the controllers outputs match satisfactorily.

IV. SIMULATIONS

To evaluate the performance of the bumpless switching mechanism, which has been synthesized according to the newly proposed rules, simulations were carried out. The model of the system has been constructed on the basis of [7] [4] [13]. The nonlinear model is computational intensive and the dynamics are well covered by the linear model, therefore simulations where performed on the linearized plant model. The linearized plant is that of 6 DOF rigid body, it consists of three masses of 8.225 kg and three inertias of $0.062159$ kgm$^2$, $0.061899$ kgm$^2$, and $0.121619$ kgm$^2$ around the $x$-, $y$-, and $z$-axis respectively. The conceptual simulation diagram is presented in Fig. 6 where, to allow for bi-directional transfer, the switching mechanism was doubled. The input to the plant is the wrench vector $\vec{w}$ and the output is the position vector $\vec{q}$. The saturation block included in the plant dynamics models the limitation of the force and torque that can be supplied to the actuator due to the limited current provided to the coils by the power amplifiers.

The original $3^{rd}$ order controller[7] has a bandwidth of 28 Hz and consists of a gain, a lead filter, an integrator and a $1^{st}$ order low-pass filter, whereas the newly designed $4^{th}$ order controller has a bandwidth of 20 Hz and consists of a gain, a lead filter, an integrator and a $2^{nd}$ order low-pass filter.

$$C_1(s) = 3.18 \cdot 10^7 \frac{(s+12)(s+20)}{s(s+180)(s+3000)}$$

$$C_2(s) = \frac{4850.4 \cdot M}{s} \frac{x+12\pi}{s+160\pi} \frac{x+2\pi}{s^2+3600s+12400\pi^2} \frac{(240\pi)^2}{2}$$

where $M$ is the mass or inertia of the corresponding SISO plant. Tracking controllers $F_1$ and $F_2$ are different because they were calculated for controllers $C_1$ and $C_2$, respectively.

The performance of the switching mechanism can be evaluated in two different ways. Firstly, steady state operation of the system is obtained and then bi-directional switching between the controllers is performed.

Without the bumpless switching mechanism, the measurement noise and the disturbances acting on the system and the mismatch of initial conditions, could cause controllers output difference.

Fig. 7 shows the plant input $u$ when the controllers were switched at times $t = 7$ s and $t = 10$ s. The measurement noise used in the simulations was equal to $2\mu$m peak-to-peak value, which correspond to the resolution of the eddy current sensors used in the measurement system of the planar actuator. One can conclude, that the bump at the switching times is within the signal noise.

Another way of evaluating the performance of the designed bumpless switching mechanism, is to apply a challenging trajectory to the system and investigate how well the controllers outputs track each other. The planar actuator model was accelerated rapidly in the $z$-direction, which simulated a fast lift-off. Fig. 8 shows both controllers outputs for that case. One can see that the feedback controllers generate fast changing control signal. The off-line controller tracks the output of the on-line controller very well. The mean square error between the controller outputs equals 0.0110 N.

V. EXPERIMENTS

The performance of the planar actuator can be improved by increasing its accuracy and speed of operation. This implies that only the modification of the performance mode controller is necessary. Therefore, the focus is on switching during the performance mode of the planar actuator. Consequently the performance of the LQ bumpless switching mechanism was tested on the performance mode controller only. The current operation of the plant requires the lift-off controller to lift the translator and switch over to the originally designed $3^{rd}$ order performance mode controller. Then, it is possible to switch over to the newly designed $4^{th}$ order controller. The transfer has to be bi-directional because only the original $3^{rd}$ order controller can smoothly decrease the bandwidth and switch over to the landing controller. This approach requires minimum changes to the existing control system providing maximum safety.

The controllers used in the experiments are the same as $C_1$ and $C_2$ in section IV. The new performance mode controller, $C_2$, does not improve the performance of the planar actuator. To test the LQ bumpless transfer mechanism, it was decided to switch over to a controller with a lower bandwidth. This design was deliberately chosen to ensure a
stabilizing controller so instability of the closed loop is only possible due to switching and not due to plant uncertainty. Secondly, the order of both controllers is different. The choice to switch to a controller of higher order was made to test if the switching method can handle switching between controllers with different structure.

This section presents the experimental results of the LQ bumpless transfer method applied to the planar actuator system. In section V-A switching in steady state is considered. Section V-B deals with the performance analysis based on the difference in controllers outputs.

A. Switching in steady state

The planar actuator was lifted and switched over to the original performance mode controller, $C_1$, using the original switching method explained in I-B. After the system has reached steady state, bi-directional switching between both controllers was executed. Fig 9 elucidates the plant input $u$ with manual switching at approximately $t = 5.2$ s and $t = 6.5$ s. The switching was bumpless. Moreover, one can observe that the noise, which is mainly caused by measurement noise, is lower when the 4th order controller is active. Due to a lower bandwidth and a high frequency roll-off with a steeper slope of controller $C_2$.

B. Performance of the tracking loops

A demanding reference trajectory profile was applied to the $x$– and $y$–directions of the planar actuator simultaneously (acceleration equal to $10^2 \text{m/s}^2$). Fig. 10 and Fig. 11 present the off-line and on-line controller outputs. It is necessary to explain why the oscillations occur. An external $xy$ positioning system consisting of three linear motors (H-drive) moves along with the planar actuator. It is used for safety reasons and prevents damaging the actuator. During normal operation of the plant, there is no contact between the translator and the measurement frame. The position and orientation of the translator of the planar actuator is constructed from the sensors mounted on the external positioning system and the long stroke linear encoders of the H-drive. High accelerations of the H-drive excite oscillations, which transfer to the position measurement of the translator [7].

One observes that the off-line controller output tracks the on-line controller output very well. The mean square error between the controller outputs equals 0.0399 N for the $x$–axis and 0.0419 N for the $y$–axis, respectively. During
the synthetization of the tracking controller, which is used for bumpless switching, numerical issues have been encountered. A newly developed correct selection procedure for the values of the weighting matrices in the LQ cost function, resolves this ill conditioning.

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