Abstract: The accuracy of manufacturing highly characterises the performance of industrial motion control. However, detrimental effects such as nonlinear coupling, model uncertainty and unknown external disturbances severely affect trajectory tracking performance. In this paper, we proposed an ADRC and feedback linearisation-based control algorithm for the high-precision trajectory tracking of feed drives. The controller was rigorously designed to ensure the convergence of observer states and tracking error. The applicability of the proposed approach was successfully demonstrated via high-fidelity simulation, and the numerical results based on different tracking methods were compared.

Keywords: feedback linearisation; ADRC; motion control

1. Introduction

With the increasingly high demand for industrial manufacturing, the control of feed drives keeps drawing attention from engineers and researchers [1–5]. Consisting of motor and transmission parts, the feed drive plays an important role in delivering translational movement during the machining process [2,4].

However, the control of feed drives is not trivial. The detrimental effects such as nonlinearities, modelling uncertainty and unknown disturbance can deteriorate the tracking performance and/or lead to system instability [6]. Although the proportional-integral-derivative (PID)-based cascaded control is still the most widely adopted control method in the industry, the increasing requirements on machining accuracy inspire the development of advanced control laws for industrial feed drives [7].

In real applications, field-oriented control is extensively used due to its simplicity [8]. It simplifies the design of controllers by achieving the separation between the flux and torque/force references [8,9]. However, strong nonlinear coupling exists between the direct and quadratic direction, which makes the control problem challenging.

To solve this problem, the feedback linearisation-based method can be investigated. With nonlinear couplings cancelled in the system, the well-known linear control strategies can be used [10]. The feedback linearisation-based control methods are widely used, and related applications on feed drives control can be found in [10–12]. However, to implement this algorithm, all the states of the system need to be measurable or additional observers are required for the partial feedback linearisation system. Considering the exact cancellation is unachievable since model mismatch exists, extra control parts are required to ensure the robustness of systems.

In addition to nonlinearities in the electrical subsystem, disturbance forces on feed drives such as friction, cogging force and the changes of the load strongly limit the tracking accuracy at high speed [13–15]. When a disturbance force is measurable, it is clear that a feed-forward structure or the pre-mentioned feedback linearisation method can attenuate or eliminate the influence of disturbance. However, for the cases where the external disturbance cannot be directly measured or is too expensive to measure, a possible way is
to estimate the disturbance from measurable variables and then design an observer-based controller for disturbance compensation [13,16,17].

Among numerous disturbance observer-based control methods, the active disturbance rejection control (ADRC) method is a popular choice and is widely adopted in the industry [18,19]. The general structure of ADRC is composed of three parts including the tracking differentiator (TD), the nonlinear weighted sum and the extended state observer (ESO). As a non-model-based disturbance rejection method, only the order and partial estimation of system parameters are required to estimate the external disturbance. The ease of use of ADRC has seen various applications including robotic systems [20,21], vehicle suspension system [22,23] and power electronics [24,25].

Although ADRC has been investigated in motion control and previous research work such as [26–28] can be found, the controller is not designed in a systematic way, and the convergence of the controller cannot be guaranteed. Different from other control algorithms for which the theoretical outcome promotes the development of applications, the theoretical work of ADRC was deficient at the time it was developed. Due to its popularity in applications, the theoretical research work of ADRC has increased [29–31], and it has become necessary to design an ADRC-based controller for feed drives in a rigorous format.

In this paper, a control architecture based on ADRC and feedback linearisation was proposed for the trajectory tracking of industrial feed drives. The ADRC was applied to tackle unknown/unmodelled disturbances in the mechanical subsystem, while the feedback linearisation-based controller was utilised in the electrical subsystem to further improve tracking performance. The effectiveness of the proposed control method was validated on a high-fidelity model of an industrial linear motor, and the tracking performance was compared with benchmark controllers.

This paper is organised as follows. The mathematical models of translational feed drives are described first. The detailed dynamics of the electrical and mechanical subsystem are derived. This is followed by the design of the feedback linearisation-based controller for the current loop and the ADRC-based controller for the mechanical subsystem. The results of trajectory tracking using the proposed method are presented in the Results Section, and numerical comparisons with the cascaded controller and the cascaded controller with the extended state observer are demonstrated after. Finally, the conclusions are given.

2. Modelling of Feed Drives

To achieve linear movement, linear permanent magnet synchronous motors and rotary permanent magnet synchronous motors with transmission parts are widely used. In this section, the dynamics of the electrical and mechanical subsystem of a general feed drive are derived.

2.1. Modelling of Electrical Subsystem

Generally, the d-q (direct and quadrature) axis model is frequently used for the design of control algorithms in sinusoidally excited permanent magnet synchronous motors. By doing so, time-varying parameters are eliminated, and all variables are expressed in orthogonal or mutually decoupled d and q axes [9].

For a general permanent magnet synchronous motor (PMSM), the governing equations of the current loop are:

\[ u_d = R_a i_d + \frac{d}{dt} \frac{\psi_d}{d} - \omega_e \psi_q \]

\[ = R_a i_d + L_d \frac{di_d}{dt} - \omega_e L_q i_q \]  

(1)
\[ u_q = Ra_iq + \frac{d\psi_q}{dt} + \omega_e\psi_d \]
\[ = Ra_iq + L_d \frac{di_q}{dt} + \omega_e L_di_d + \omega_e \psi_{PM} \]  
(2)

where \( R_a \) is the armature winding resistance, \( u_d, u_q \) and \( i_d, i_q \) are the terminal voltage and current of the \( d \) and \( q \) axes, \( \psi_d, \psi_q \) and \( L_d, L_q \) are the magnetic flux and resultant armature inductances on the \( d \) and \( q \) axes, respectively, \( \omega_e \) is the angular frequency of armature current and the flux linkage \( \psi_{PM} \) is a constant for the permanent magnet motor.

To get the torque/force equation of a rotary/linear motor, we start from the electromagnetic power of a three-phase PMSM as:

\[ P = \frac{3}{2} \omega_e (\psi_d i_q - \psi_q i_d) \]
\[ = \frac{3}{2} \omega_e [(L_d i_d + \psi_{PM}) i_q - L_q i_d i_q] \]
\[ = \frac{3}{2} \omega_e [\psi_{PM} + (L_d - L_q) i_d] i_q \]

(3)

For the most widely used PMSM, the surface mounted permanent magnet synchronous motor has the property \( L_d = L_q \). Without loss of generality, the electromagnetic force generated by a \( p \) pole pair linear permanent magnet linear synchronous motor (LPMSM) is:

\[ F_m = \frac{P}{v_m} = \frac{3}{2} p \frac{\pi}{\tau} \psi_{PM} i_q \]

(4)

where \( \omega_e = p \omega_m = p \pi v_m / \tau \) holds, \( v_m \) is the translational velocity of the motor, and \( \tau \) is the pole pitch.

Then, the dynamics of the electrical subsystem for the linear PMSM is:

\[ i_d = -\frac{R_a}{L_d} i_d + \frac{p \pi}{\tau} i_q v_m + \frac{u_d}{L_d} \]
\[ i_q = -\frac{p \pi}{\tau} i_d v_m - \frac{R_a}{L_q} i_q - \frac{p \pi \psi_{PM}}{\tau L_q} v_m + \frac{u_q}{L_q} \]

(5)
(6)

2.2. Modelling of the Mechanical Subsystem

In this subsection, the dynamics of the mechanical subsystem for direct driven motors and rotary motors with a rack and pinion are derived. The unified model of the mechanical subsystem for the PMSM is given after.

The mechanical subsystem of the direct-driven feed drives can be presented as:

\[ M_m \ddot{x}_m = k_1 i_q + F_d \]

(7)

where \( M_m \) is the mass of the translational movement part, \( x_m \) is the position of the motor, \( k_1 = \frac{3}{2} p \frac{\pi}{\tau} \psi_{PM} \) is the force constant of the linear motor and \( F_d \) is the lumped disturbance imposed on the translational movement part.

For feed drives driven by the rotary PMSM and transmission mechanisms, the dynamics of the mechanical subsystem can be converted into a similar form as (7). Here, the derivation of the mechanical dynamics of a rotary motor with a rack and pinion is demonstrated.

The schematic and block diagrams of a rotary motor with a rack and pinion mechanism are shown in Figure 1, where the gearbox between the motor and pinion is omitted. Then, the dynamics can be described as:
where $\omega_m = \frac{n_g}{R_r} \omega_m$ holds, $k_r$ is the torque constant of the rotary motor, $n_g$ is the gear ratio, $R_r$ is the radius of the pinion and $J_m$ is the moment of inertia of the rotary motor.

By rearranging (8), we have:

$$\left( M_r + \frac{J_m n_g^2}{R_r^2} \right) \dot{v}_m = n_g k_r i_q + F_d$$

(9)

Then, the uniform format of the mechanical subsystem can be represented by:

$$M_e \ddot{x}_m = k_v i_q + F_d$$

where $M_e$ is the equivalent mass, which is $M_e = M_m$ for the linear motor and $M_e = M_r + J_m n_g^2 / R_r^2$ for the rotary motor with the rack and pinion; $k_v$ is the equivalent force constant, which is $k_v = k_I$ for the linear motor, and $k_v = n_g k_r / R_r$ for the rotary motor with the rack and pinion.

### 3. Controller Design

In this section, the design of the proposed controller for the linear PMSM is demonstrated. The controller design procedure for the rotary PMSM can be conducted by a similar procedure; thus, it is omitted.

Inspired by the idea that ADRC can be used to estimate and compensate unknown external disturbance and model mismatch in the velocity loop while feedback linearisation takes care of the nonlinear coupling in the electrical subsystem, a trajectory tracking architecture based on ADRC and feedback linearisation was proposed. The entire system including the controllers and plant is demonstrated in Figure 2. Note that $v_m, i_d$ and $i_q$ in the dashed lines represent the nonlinear coupling in (5) and (6).

![Figure 2](image-url)
To cancel the nonlinear coupling between the $d$ and $q$ axes, feedback linearisation is utilised in the current loop for the controller design. The $d$ axis controller is designed as:

$$u_d(i_d, i_q, v_m) = L_d \left(-\frac{p\pi}{\tau} i_q v_m - k_d i_d\right)$$  \hspace{1cm} (10)

where $k_d$ is the tuning parameter of the $d$ axis current controller. The dynamics of the $d$ axis current error becomes:

$$\dot{e}_d = -\left(\frac{R_d}{L_d} + k_d\right) e_d$$  \hspace{1cm} (11)

where $e_d = i_d^* - i_d$, $i_d^* = 0$ is the desired value of $i_d$ and $k_d$ determines the speed of error convergence.

By defining $e_q = i_q^* - i_q$, the dynamics of $e_q$ is:

$$\dot{e}_q = i_q^* - i_q = i_q^* + \frac{p\pi}{\tau} i_d v_m + \frac{R_d}{L_q} i_q + \frac{p\pi \psi_{PM}}{\tau L_q} v_m - \frac{u_q}{L_q}$$  \hspace{1cm} (12)

The $q$ axis current controller is thus designed based on the feedback linearisation method to cancel the nonlinearity in (12) as:

$$u_q = L_q \left(i_q^* + \frac{p\pi}{\tau} i_d v_m + \frac{p\pi \psi_{PM}}{\tau L_q} v_m + \frac{R_a}{L_q} i_q + k_q e_q\right)$$  \hspace{1cm} (13)

where $k_q$ is the tuning parameter. By applying (13), the dynamics of (12) becomes:

$$\dot{e}_q = -k_q e_q$$  \hspace{1cm} (14)

where the value of $k_q$ determines the convergence rate of $e_q$.

Then, the objective of outer-loop ADRC is to generate the reference $i_q^*$ and $i_d^*$ for the current loop. To get the derivatives of the position reference with less noise amplification, the tracking differentiator (TD) is utilised and designed as:

$$\dot{x}_m = \hat{x}_m^*$$
$$\dot{\hat{x}}_m^* = \hat{\dot{x}}_m^*$$
$$\hat{\dot{x}}_m^* = -\gamma^3 (\hat{x}_m^* - x_m^*) - 3\gamma^2 \hat{\dot{x}}_m^* - 3\gamma \hat{\ddot{x}}_m^*$$  \hspace{1cm} (15)

where $\hat{x}_m$, $\dot{x}_m$ and $\ddot{x}_m$ are the estimated values of the desired position, velocity and acceleration, respectively; $\gamma$ is the tuning parameter of the tracking differentiator. When $\hat{x}_m^*$ converges to $x_m^*$, $\dot{x}_m^* \rightarrow \dot{x}_m$ and $\ddot{x}_m^* \rightarrow \ddot{x}_m$ hold.

If the time scale of the current loop is set fast enough by tuning the controller $k_d$ and $k_p$, the time scale separation between the electrical and mechanical subsystem is achieved, and $i_q \approx i_q^*$. Then, the mechanical subsystem becomes:

$$\dot{x}_m = v_m$$
$$\dot{v}_m = \frac{k_f}{M_m} i_q^* + \left(\frac{k_i}{M_m} - \frac{k_i'}{M_m'}\right) i_q + \frac{F_d}{M_m}$$
$$= b_0 i_q^* + \xi_m$$  \hspace{1cm} (16)

where $M_m$ and $k_f'$ are the nominal values of the motor mass and force constant, $\xi_m \triangleq \left(\frac{k_i}{M_m} - \frac{k_i'}{M_m'}\right) i_q + \frac{F_d}{M_m}$ is the lumped disturbance including model mismatch and unknown disturbance and $b_0 \triangleq \frac{k_f'}{M_m'}$. 
In order to acquire the feedback signals with less noise and cancel the impact from lumped disturbance, an extend state observer (ESO) is designed to estimate the value of $x_m$, $v_m$ and $\zeta_m$ as:

\[
\dot{\hat{x}}_m = \hat{v}_m + \frac{3}{\rho} (x_m - \hat{x}_m) + \rho \Phi \left( \frac{x_m - \hat{x}_m}{\rho^2} \right) \\
\dot{\hat{v}}_m = \hat{\zeta}_m + \frac{3}{\rho^2} (x_m - \hat{x}_m) + \frac{k'_t}{M_m} i^*_q \\
\dot{\hat{\zeta}}_m = \frac{1}{\rho^3} (x_m - \hat{x}_m)
\]

where $\rho$ is the tuning parameter of the ESO and $\hat{x}_m$, $\hat{v}_m$ and $\hat{\zeta}_m$ are the estimated outputs. The convergence of the observer ensures the stability of the proposed controller. The asymptotic stability of the designed observer was rigorously discussed in [32], and thus omitted here. The nonlinear function $\Phi: \mathbb{R} \rightarrow \mathbb{R}$ is chosen as (18) to satisfy the conditions described in [32,33]:

\[
\Phi(r) = \begin{cases} 
-\frac{1}{4r}, & r \in (-\infty, -\frac{\pi}{2}], \\
\frac{1}{2}\sin r, & r \in \left(-\frac{\pi}{2}, \frac{\pi}{2}\right], \\
\frac{1}{4}, & r \in \left(\frac{\pi}{2}, +\infty\right),
\end{cases}
\]

with the estimated lumped disturbance from the ESO, the control law of the $q$ axis is set:

\[
i^*_q = \frac{1}{b_0} \left( \varphi((\hat{x}_m - \hat{x}^*_m), (\hat{v}_m - \hat{v}^*_m)) + \hat{\zeta}_m - \hat{\xi}_m \right)
\]

where $\hat{x}^*_m$, $\hat{v}^*_m$ and $\hat{\xi}_m$ are the outputs of the tracking differentiator, $\hat{x}_m$, $\hat{v}_m$ and $\hat{\zeta}_m$ are the outputs of the ESO and $\varphi(u_1, u_2) = -2u_1 - 4u_2 - \Phi(u_1)$.

Then, the control inputs of the system are $u_d$ and $u_q$, which are governed by (10) and (13) using the feedback linearisation-based method. At the outer loop, the ADRC is utilised to generate the current reference $i^*_q$ based on (19).

4. Results

4.1. High-Fidelity Model and System Identification

Due to the hardware limitation of the plant, the effectiveness of the proposed control architecture was validated by a high-fidelity model-based trajectory tracking. The system identification was conducted on a linear permanent magnet synchronous tubular motor, as shown in Figure 3.

The values of the motor parameters including resistance $R_a$, inductance $L_d$, $L_q$ and permanent magnet flux $\psi_{PM}$ were identified by applying the least squares method based on (6) with adequate feedback of $i_q$, $v_m$ and $u_q$ using two reference profiles. The value of identified parameters was validated by comparing the measured $i_q$ with the estimated $i_q$ computed by the system model, as shown in Figure 4. The value of motor mass $M_m$ and pole pitch $\tau$ were found in the product manual. The value of the linear motor is summarised in Table 1. In order to test the robustness of the proposed method, fifteen percent model mismatch was considered.
Table 1. Parameters of the linear motor.

| Name of Parameters          | Value | Unit |
|-----------------------------|-------|------|
| Resistance ($R_a$)          | 12.5  | Ω    |
| Inductance ($L_d = L_q$)    | 10.7  | mH   |
| Weight ($M_m$)              | 6     | kg   |
| Pole pitch ($\tau$)         | 180   | mm   |
| Permanent magnet flux ($\psi_{PM}$) | 1.01  | Wb   |

Figure 3. Linear permanent magnet synchronous tubular motor used for system identification.

The friction force is considered as the disturbance forces in the simulation and is governed by the following equation:

$$F_f(v_m) = f_v v_m + \left(f_c + (f_s - f_c)e^{-\frac{|v_m|}{v_s}}\right)\text{sgn}(v_m)$$  \hspace{1cm} (20)

where $f_v$ is the coefficient of viscous friction, $f_c$ represents the minimum level of Coulomb friction, $f_s$ stands for the level of static friction and $v_s$ and $\delta_s$ are the empirical parameters for the Stribeck effect. Values $f_v = 0.8$, $f_c = 10$, $f_s = 20$, $v_s = 0.001$ and $\delta_s = 2$ were chosen.

4.2. Tracking Results and Comparison

Conventionally, the point-to-point movement or sinusoidal reference are used for performance validation in motion control considering that most of the paths are consist
of a straight line or a circle [7]. The tracking error is used as the performance metrics to evaluate the tracking accuracy [34,35].

In this paper, the simulated reference tracking was conducted on the identified high-fidelity model for reference with \( x_m^* = 0.02 \sin(8\pi t) \). The effect of discretisation was considered, and the sampling rates for the position, velocity and current loop were 4 kHz, 4 kHz and 16 kHz, respectively.

The commissioning effort of the proposed controller was relatively trivial compared to the conventional cascaded controller. The parameters included the gain of the ESO \( \varrho \), the tuning parameter in the tracking differentiator \( \gamma \) and the proportional gain in the current loop \( k_d \) and \( k_q \). If we investigate the dynamics of the ESO (17), tracking differentiator (15) and current error on the \( d \) axis (11) and \( q \) axis (14), we can see that the values of \( 1/\varrho \), \( \gamma \), \( k_d \) and \( k_q \) determine the speed of states' convergence. In order to ensure the high-speed convergence of the system states while avoiding the sharp spike caused by high gains, a reasonable value of 1000 was chosen for \( 1/\varrho \), \( \gamma \), \( k_d \) and \( k_q \). The values of the tuning parameters are summarised in Table 2.

Table 2. Tuning parameters of the proposed controller.

| Method                  | Control Parameters | Value |
|-------------------------|--------------------|-------|
| ADRC                    | \( \varrho \)      | 0.001 |
|                         | \( \gamma \)      | 1000  |
| Feedback linearisation  | \( k_d \)         | 1000  |
|                         | \( k_q \)         | 1000  |

The profile of the moving trajectory and the estimation of the lumped disturbance based on the proposed method are shown in Figure 5a,b, respectively. It can be seen that the position of the motor can follow the reference precisely. The estimated lumped disturbance converged fast to the periodic form to approximate the disturbance forces caused by unmodelled friction force and model mismatch. If we look into the control inputs \( u_d \) and \( u_q \) generated by the proposed controller, we can see from Figure 6 that both control inputs were within the 480 V voltage range, and there was no saturation for the actuators.

![Figure 5](image-url)  
**Figure 5.** Reference tracking based on the proposed controller: (a) trajectory; (b) estimation of lumped disturbance.
Based on the fact that the cascaded PI controller and extended state observer-based controller are still widely used in the industry, they served as good candidates to provide benchmarks for assessing the performance of the proposed method. Thus, the cascaded controller in the structure in Figure 7 and state observer-based controller in the structure in Figure 8 were used for performance comparison.

For the cascaded control structure, the time scale of the current loop was the fastest. The bandwidth of the inner loop needed to be 5–10 times larger than the outer loop. Considering that the bandwidth of the position loop is usually set as 70 Hz in industrial applications, the cascaded controller was tuned to achieve 70 Hz, 350 Hz and 1750 Hz of the position, velocity and current loop bandwidth to ensure the feasibility of cascaded control. The proportional gain and integral time constant of the current loop were chosen as $k_{pc} = 358 \, \text{V/A}$, $T_{ic} = 0.0283 \, \text{ms}$. The proportional gains of the integral time of the velocity loop were $k_{pv} = 310 \, \text{As/m}$ and $T_{iv} = 0.0453 \, \text{ms}$. The proportional gain of the position loop was chosen as $k_{pp} = 200 \, 1/\text{s}$. As for the ESO-based controller, the same gains were chosen for the PI controller, and the same value for $\varrho$ in the proposed method was selected.

The reference tracking based on the same reference was conducted on the benchmark controllers. The tracking errors using three different controllers are presented in Figure 9. By comparing the tracking errors using the cascaded controller and cascaded controller with the ESO, we can see that the inclusion of the ESO part in the control architecture...
improved the tracking accuracy compared to the classical cascaded controller. Among the three control methods, the proposed controller achieved the best tracking performance with less position lag and higher accuracy compared to the conventional cascaded controller and ESO-based controller. The maximum tracking error was less than 2 mm using the proposed approach.

![Figure 9](image)

**Figure 9.** Tracking error comparison based on three different controllers.

Since model mismatch exists in practical applications, the Monte Carlo scheme was used to test the performance of different control algorithms when the given range of model uncertainties was considered in the simulation. By randomly varying the parameters in a ±15% range and repeating the simulation 100 times, we obtained the numerical performance comparison results of the investigated control algorithms in Table 3.

**Table 3.** Trajectory tracking performance of different control algorithms.

| Method                  | \(\max|e(t)|\) * | RMSE * | \(\int |e|\,dt\) * |
|------------------------|-----------------|--------|-------------------|
| Cascaded control       | 2.7655          | 1.9207 | 0.8630            |
| Cascaded PI with ESO   | 2.6867          | 1.8707 | 0.8415            |
| ADRC with FL           | 1.9234          | 1.2834 | 0.5794            |

* \(t \in [0, T_f]\), where \(T_f\) is the process time; the units are mm.

Based on the numerical results in Table 3, we showed that the proposed control method achieved the best tracking performance in both transient and steady states, which can be seen from the maximum and accumulated tracking error. In addition, by comparing the results achieved by the cascaded controller and cascaded controller with the extended state observer, we can see that the tracking performance can be improved if the external disturbances are attenuated by the observer.

5. Conclusions

In this paper, we presented a trajectory tracking method based on active disturbance rejection control and feedback linearisation for industrial feed drives. With the derivation of the system dynamics, the proposed controller was presented in a rigorous way. The non-linearity in the current loop was cancelled by the feedback linearisation scheme, and ADRC was capable of eliminating the impact from unknown disturbance.

The proposed algorithm was then tested in a simulation on a high-fidelity linear motor. It was demonstrated that the proposed controller can achieve decent tracking results and relatively better performance compared to the conventional cascaded controller and cascaded controller with the extended state observer.

**Author Contributions:** Writing—original draft preparation, M.Y.; writing—review and editing, Z.X.; funding acquisition, Z.X. Both authors read and agreed to the published version of the manuscript.
Funding: This research was funded by the research funding of the National Natural Science Foundation of China grant number 61973085.

Acknowledgments: We would like to acknowledge and thank Chris Manzie, Iman Shames and Malcolm Good for their discussions and suggestions at the University of Melbourne. We would like to thank ANCA Motion, Australia, for the support and supply of the experiment facilities.

Conflicts of Interest: The authors declare no conflict of interest.

Abbreviations
The following abbreviations are used in this manuscript:

ADRC  Active disturbance rejection control
FL    Feedback linearisation
TD    Tracking differentiator
ESO   Extended state observer
PMSM  Permanent magnet synchronous motor
PID   Proportional-integral-derivative
PI    Proportional-integral

References
1. Koren, Y.; Lo, C. Advanced Controllers for Feed Drives. CIRP Ann. 1992, 41, 689–698. [CrossRef]
2. Van Brussel, H.; Van den Braembussche, P. Robust control of feed drives with linear motors. CIRP Ann. 1998, 47, 325–328. [CrossRef]
3. Erkorkmaz, K.; Hosseinkhani, Y. Control of ball screw drives based on disturbance response optimization. CIRP Ann. Manuf. Technol. 2013, 62, 387–390. [CrossRef]
4. Altintas, Y.; Verl, A.; Brecher, C.; Uriarte, L.; Pritschow, G. Machine tool feed drives. CIRP Ann. Manuf. Technol. 2011, 60, 779–796. [CrossRef]
5. Erkorkmaz, K.; Altintas, Y. High speed CNC system design. Part III: High speed tracking and contouring control of feed drives. Int. J. Mach. Tools Manuf. 2001, 41, 1637–1658. [CrossRef]
6. Yao, J.; Jiao, Z.; Ma, D. Adaptive robust control of DC motors with extended state observer. IEEE Trans. Ind. Electron. 2014, 61, 3630–3637. [CrossRef]
7. Yuan, M.; Manzie, C.; Good, M.; Shames, I.; Gan, L.; Keynejad, F.; Robinette, T. A review of industrial tracking control algorithms. Control Eng. Pract. 2020, 102, 104536. [CrossRef]
8. Shyu, K.-K.; Lai, C.-K.; Tsai, Y.-W.; Yang, D.I. A newly robust controller design for the position control of permanent-magnet synchronous motor. IEEE Trans. Ind. Electron. 2002, 49, 558–565. [CrossRef]
9. Gieras, J.F.; Piech, Z.J.; Tomczuk, B. Linear Synchronous Motors: Transportation and Automation Systems; CRC Press: Boca Raton, FL, USA, 1999.
10. Ilic’-Spong, M.; Marino, R.; Peresada, S.; Taylor, D. Feedback linearizing control of switched reluctance motors. IEEE Trans. Autom. Control 1987, 32, 371–379. [CrossRef]
11. Castillo-Berrio, C.F.; Feliu-Batlle, V. Vibration-free position control for a two degrees of freedom flexible-beam sensor. Mechatronics 2015, 27, 1–12. [CrossRef]
12. Aguilar-Avelar, C.; Moreno-Valenzuela, J. New feedback linearisation based control for arm trajectory tracking of the Furuta pendulum. IEEE/ASME Trans. Mechatron. 2016, 21, 638–648. [CrossRef]
13. Jamaludin, Z.; Van Brussel, H.; Swevers, J. Friction compensation of an XY feed table using friction-model-based feedforward and an inverse-model-based disturbance observer. IEEE Trans. Ind. Electron. 2009, 56, 3848–3853. [CrossRef]
14. Hu, C.; Yao, B.; Wang, Q. Coordinated adaptive robust contouring control of an industrial biaxial precision gantry with cogging force compensations. IEEE Trans. Ind. Electron. 2010, 57, 1746–1754. [CrossRef]
15. Li, S.; Xia, C.; Zhou, X. Disturbance rejection control method for permanent magnet synchronous motor speed-regulation system. Mechatronics 2012, 22, 706–714. [CrossRef]
16. Ohnishi, K.; Shibata, M.; Murakami, T. Motion control for advanced mechatronics. IEEE/ASME Trans. Mechatron. 1996, 1, 56–67. [CrossRef]
17. Umeno, T.; Hori, Y. Robust speed control of DC servomotors using modern two degrees-of-freedom controller design. IEEE Trans. Ind. Electron. 1991, 38, 363–368. [CrossRef]
18. Han, J. From PID to active disturbance rejection control. IEEE Trans. Ind. Electron. 2009, 56, 900–906. [CrossRef]
19. Feng, H.; Guo, B.Z. Active disturbance rejection control: Old and new results. Annu. Rev. Control 2017, 44, 238–248. [CrossRef]
20. Huang, Y.; Luo, Z.; Svinin, M.; Odashima, T.; Hosoe, S. Extended state observer based technique for control of robot systems. Proceedings of the 4th World Congress on Intelligent Control and Automation. Proc. IEEE 2002, 4, 2807–2811. [CrossRef]
21. Su, J.; Qiu, W.; Ma, H.; Woo, P.Y. Calibration-free robotic eye-hand coordination based on an auto disturbance-rejection controller. IEEE Trans. Robot. 2004, 20, 899–907. [CrossRef]
22. Pan, H.; Sun, W.; Gao, H.; Hayat, T.; Alsaadi, F. Nonlinear tracking control based on extended state observer for vehicle active suspensions with performance constraints. *Mechatronics* **2015**, *30*, 363–370. [CrossRef]

23. Xia, Y.; Fu, M.; Li, C.; Pu, F.; Xu, Y. Active Disturbance Rejection Control for Active Suspension System of Tracked Vehicles With Gun. *IEEE Trans. Ind. Electron.* **2018**, *65*, 4051–4060. [CrossRef]

24. Sun, B.; Gao, Z. A DSP-based active disturbance rejection control design for a 1-kW H-bridge DC–DC power converter. *IEEE Trans. Ind. Electron.* **2005**, *52*, 1271–1277. [CrossRef]

25. Wu, G.; Sun, L.; Lee, K.Y. Disturbance rejection control of a fuel cell power plant in a grid-connected system. *Control Eng. Pract.* **2017**, *60*, 183–192. [CrossRef]

26. Su, Y.; Zheng, C.; Duan, B. Automatic disturbances rejection controller for precise motion control of permanent-magnet synchronous motors. *IEEE Trans. Ind. Electron.* **2005**, *52*, 1271–1277. [CrossRef]

27. Sira-Ramirez, H.; Linares-Flores, J.; García-Rodriguez, C.; Contreras-Ordaz, M.A. On the control of the permanent magnet synchronous motor: An active disturbance rejection control approach. *IEEE Trans. Control Syst. Technol.* **2014**, *22*, 2056–2063. [CrossRef]

28. Tian, G.T.G.; Gao, Z.G.Z. Benchmark tests of Active Disturbance Rejection Control on an industrial motion control platform. In Proceedings of the 2009 American Control Conference, St. Louis, MO, USA, 10–12 June 2009; pp. 5552–5557. [CrossRef]

29. Guo, B.Z.; Zhao, Z.L. On convergence of tracking differentiator. *Int. J. Control* **2011**, *84*, 693–701. [CrossRef]

30. Guo, B.Z.; Zhao, Z.L. On Convergence of the Nonlinear Active Disturbance Rejection Control for MIMO Systems. *SIAM J. Control Optim.* **2013**, *51*, 1727–1757. [CrossRef]

31. Zhao, Z.L.; Guo, B.Z. On Convergence of Nonlinear Active Disturbance Rejection Control for SISO Nonlinear Systems. *J. Dyn. Control Syst.* **2016**, *22*, 385–412. [CrossRef]

32. Guo, B.; Zhao, Z. On the convergence of an extended state observer for nonlinear systems with uncertainty. *Syst. Control Lett.* **2011**, *60*, 420–430. [CrossRef]

33. Guo, B.Z.; Wu, Z.H.; Zhou, H.C. Active Disturbance Rejection Control Approach to Output-Feedback Stabilization of a Class of Uncertain Nonlinear Systems Subject to Stochastic Disturbance. *IEEE Trans. Autom. Control* **2016**, *61*, 1613–1618. [CrossRef]

34. Makkar, C.; Hu, G.; Sawyer, W.; Dixon, W. Lyapunov-based tracking control in the presence of uncertain nonlinear parameterizable friction. *IEEE Trans. Autom. Control* **2007**, *52*, 1988–1994. [CrossRef]

35. Yuan, M.; Manzie, C.; Good, M.; Shames, I.; Gan, L.; Keynejad, F.; Robinette, T. Error-Bounded Reference Tracking MPC for Machines with Structural Flexibility. *IEEE Trans. Ind. Electron.* **2019**, *67*, 8143–8154. [CrossRef]