Research on the Hoisting Motor Drive System's Active Disturbance Rejection Control and Energy Consumption for a Crane

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ABSTRACT To allow the hoisting motor drive system of a crane to track a load torque quickly, a linearization method was used to transform a motor nominal dynamics model into two decoupled linear rotor speed and flux linkage subsystems. The method based on the theory of differential geometry was a precise feedback method. Two active disturbance rejection controllers (ADRCs) with identical structures were designed for the rotor speed and flux linkage subsystems. The extended state observer of the ADRC could estimate the unmodeled dynamics of the motor, the variation of motor parameters due to heating, and the unknown disturbances of the motor system to determine the total disturbances of the system. A closed-loop system with ADRC and an open-loop system were compared. The motor's full-load starting time was reduced by about 50%. When the motor operated smoothly at different load rates and the rated load was suddenly applied, the electromagnetic torque fluctuation range did not exceed 20 N · m. The rotor flux was always stable at the reference value. The motor speed decreased, but the amount of decrease did not exceed 7 rad/s. The closed-loop system had a significant energy-saving effect during the motor's starting process. The power saving rate was about 55%–59% if the motor started with a light load. The power saving rate could reach 71% if the motor started with a heavy load. The ADRC system could accurately estimate the unknown model of the rotor speed and flux linkage subsystems, and adapt to parameter variations of the motor stator and rotor resistance in the range of $\pm 10\%$.

INDEX TERMS Crane, hoisting motor drive system, active distance rejection control, energy consumption, energy-saving control.

NOMENCLATURE

Multi-Input Multi-Output (MIMO) Affine Nonlinear System:

$\boldsymbol{x}, \boldsymbol{u}, \boldsymbol{y}$	State variables, input and output vectors.
$\boldsymbol{f}(\boldsymbol{x}), \boldsymbol{g}(\boldsymbol{x})$	Sufficiently smooth vector fields.
$\boldsymbol{g}_j(\boldsymbol{x})$	The *j*-th column vector.
$f_i(\boldsymbol{x})$	The *j*-th scalar function.
$h_j(\boldsymbol{x})$	The *j*-th output function.
$\boldsymbol{\gamma}$	Relative order vector.
γ_j	The *j*-th relative order number.

Motor Drive System of Lifting Mechanism for the Crane:

J_e, J_1, J_2, J_r	Moments of inertia.
n_1, n_2, n_r	Rotational speeds.
j_1, j_2, j_r	Reduction ratios.
$\eta_1, \eta_2, \eta_r, \eta_L$	Transmission efficiencies.
D	Diameter of the lifting drum.
m_L	Payload.
m_0	Quality of the hook.
F_L	Wire rope's static tension.
g	Gravitational acceleration.

Motor:

T_e, T_L	Electromagnetic and load torques.
T_N	Nominal torque.

E_T	Electromagnetic torque tracking error.
n_p	Motor's pole logarithm.
J_e	Rotational inertia of the rotor.
n	Rotor speed.
ω	Electric angular velocity.
ω_r	Mechanical angular velocity.
α, β	Static two-phase coordinate axis.
A, B, C	Three-phase coordinate axis.
$u_{s\alpha}, u_{s\beta}, u_{r\alpha}, u_{r\beta}$	Stator and rotor voltages in α-β coordinate system.
$i_{s\alpha}, i_{s\beta}, i_{r\alpha}, i_{r\beta}$	Stator and rotor currents in α-β coordinate system.
$\psi_{s\alpha}, \psi_{s\beta}, \psi_{r\alpha}, \psi_{r\beta}$	Stator and rotor flux linkages.
R_s, R_r	Stator and rotor resistances.
L_s, L_r, L_m	Stator, rotor and mutual inductances.
u_A^*, u_B^*, u_C^*	PWM's three-phase modulation voltages.
u_A, u_B, u_C	Motor's three-phase control voltages.
$I(t), U(t), P(t)$	Effective value of the stator current, voltage and power.
W_{sk}, W_{sb}	Starting energy consumption.
t_s	Starting time.
ΔE	Power saving rate.

ADRC:

$\boldsymbol{\beta}^{(ESO)}, \boldsymbol{\beta}^{(ENC)}$	Parameter vector of ADRC.
$e^{(TD)}, e^{(ESO)}, e_1^{(ENC)}, e_2^{(ENC)}$	Tracking error.
w_1, w_2	TD's outputs.
z_1, z_2, z_3	State variables, and expanded state variable.
$\hat{z}_1, \hat{z}_2, \hat{z}_3$	Estimated state variables.
b	Compensation factor of the control strength.
T	Integral step.
r	Speed factor.
v_1, v_2	Virtual control input.
y_1^*, y_2^*	Reference input.

I. INTRODUCTION

A crane is an important type of logistics equipment to realize material handling mechanization in factory workshops or warehouses, railway freight yards, port wharfs, and other locations [1], [2]. Three phase AC (alternating current) asynchronous motors (hereinafter referred to as motors) serve as the main power devices for a crane's cart running mechanism, trolley running mechanism, and spout lifting mechanism [3], [4]. The motor and drive system of the running or lifting mechanism constitute the motor drive system of the crane. In the life cycle of the crane, 97%–98% of the operating cost is used to maintain the electric charge of the crane motor drive system [5], [6]. Thus, effective control of the crane motor drive system is an important means to improve the crane automation level, energy consumption, and crane operation efficiency [7], [8].

Crane systems mainly include a cart motor drive system, trolley motor drive system and a hoisting (lifting) motor drive system. As a main factor of improvement in the energetic efficiency of the crane, particular attention should be paid on the hoisting motor drive system [9]. For the crane hoisting motor drive system, Chen *et al.* [10] proposed an energy loss system, derived the calculation formula of the energy recovery late, and developed the control strategy of the energy-saving system. But Chen's research was based on the mechanical connection with the winch-motor-synchronous motor. The hoisting motor drive system has the distinct characteristics of a continually applied torque and variable working conditions [11], [12]. The load torque is a braking torque that hinders the movement, and the braking torque includes the continually applied torque of the driving wheel resistance of the cart or trolley and the torque from the potential energy of the lifting reel [13]. For the crane's lifting motor drive system, the load torque is often uncertain, due to the variability of the load. Therefore, the energy-saving principles and control measures adopted for the crane motor drive system are different from those of a motor drive system with a constant power load, such as a fan or pump [14].

In general, there are three kinds of energy-saving control and implementation measures of a motor drive system. The first requires optimization of the motor's internal structure and its magnetic material selection [15]. Reducing the copper and iron losses from the stator and rotor windings reduces internal losses of the motor [16], [17]. The second utilizes control strategies based on the steady-state operation of the motor. The third involves vector control based on a dynamic mathematical model of the motor to realize real-time control of the motor's electromagnetic torque [18]–[20].

Steady-state control strategies include variable voltage speed regulation, variable pole speed regulation, and variable frequency speed regulation. Energy feedback utilization technology is needed to utilize the slip energy in the variable voltage speed regulation process, and the motor power factor is low. The variable pole speed regulation cannot achieve stepless speed regulation smoothly, and the speed regulation range is limited because of the motor structure and manufacturing technology. Variable frequency speed regulation exhibits no additional slip loss, a high efficiency, and a wide speed range, and it is especially suitable for operating under low loads most of the times, which is the typical operating condition of a crane, and for frequent starting and stopping operations [21], [22].

Vector control theories and methods mainly include adaptive control [23]–[25], back stepping control [26], inverse dynamic control [27], passive control [28], sliding mode control [29], [30], and fuzzy logic control [31], [32]. Owing to the coupling terms of the state variables in the motor dynamics equations, the torque and excitation components of the motor stator current influence each other, and affect the dynamic performance of the control system. The control methods described previously require complex controller design [23]–[32], and they cannot influence the decoupled

control of the output variables. In addition, when motor parameters (e.g., self-inductance) change due to factors such as temperature rise, the control system's performance may suffer. The uncertainty of the load rate also creates higher requirements for the controller design [33], [34]. If the motor's dynamic model can be decoupled while devising an equivalently simple model of the motor, the controller design will be straightforward.

For affine nonlinear systems, such as motors [35], [36], the nonlinear systems can be linearized using the method of state variable feedback, which is a nonlinear feedback linearization method based on the theory of differential geometry [37]. Unlike local linear approximations using Taylor series expansions, nonlinear feedback linearization does not ignore any higher-order nonlinear terms in the linearization process, so the linearization is accurate and global. Using this nonlinear feedback linearization method, the coupled and complex nonlinear control system of an AC induction motor is transformed into two completely decoupled and simple second-order subsystems: linear rotor speed and flux linkage. On this basis, the complete decoupling control of these subsystems can be realized. This decoupling control is based on the premise that the motor has an accurate mathematical model. However, the motor parameters and the crane's load torque often have some a degree of uncertainty. In addition, the model does not account for all the dynamics in the system, leading to inaccuracies in the system dynamics model, resulting in a nominal dynamic model of the motor.

The active disturbance rejection control (ADRC) theory and methods do not rely heavily on the controlled object's mathematical model [38], [39], so the ADRC can solve the control problem of uncertainty well [40]–[42]. Sun *et al.* [43] used the ADRC method to control the open cathode proton exchange membrane fuel cell. Deng *et al.* [44] researched a vector control strategy of the permanent magnet synchronous motor based on ADRC technology. Patelski *et al.* [45] designed an ADRC controller for uncertain robot hand models. Zhonghua [46] detailed how an ADRC was applied to the naval gun servo system. Huang *et al.* [47] explored a self-searching optimal ADRC for the pitch angle control of an underwater thermal glider under vertical plane motion. Xie *et al.* [48] applied an ADRC to solve the speed fluctuation problem of permanent magnet synchronous motors resulting from parameter variation and load disturbance. Lotufo *et al.* [49] used an ADRC for UAV attitude control. Li *et al.* [50] designed an ADRC for the path following of underactuated marine surface ships. Si *et al.* [51] found that the ADRC approach can control pollutant release in a long-distance tunnel via longitudinal ventilation. To address the precise locating control problem of a polar crane with center of gravity shifting, cross-coupling, and external disturbance, the ADRC system was utilized in the paper [52].

In this study, two ADRCs with identical structures were designed for the two decoupled linear subsystems. The unmodeled dynamics of the motor, the variation of motor parameters due to heating, and the unknown disturbances

of the motor system were estimated by an extended state observer (ESO) in the ADRC. When comparing with the classical PID control method, the control system of this study exhibited stronger robustness for the variations of the motor parameters and load torque. As such, the essence of the theory and method for the affine nonlinear system's feedback linearization and precise decoupling and ADRC are fully absorbed. First, the nonlinear feedback linearization method was used to transform the complex, nonlinear, and coupled nominal motor dynamics model into two simple, linear, and decoupled subsystems: rotor speed and flux linkage. Second, two ADRC controllers with identical structures were designed for the two subsystems, which allows the motor to start quickly and shortens the transition process of the electromagnetic torque when the motor is suddenly loaded. Finally, the energy consumption during the starting and transition processes for the control system was examined with the power saving rate compared against the open-loop system.

There are eight parts in this paper. Part I serves as the literature review. In part II, we described a condition where a multi-input multi-output (MIMO) affine nonlinear system could be linearized. In part III, we established the dynamic equation of the hoisting motor drag system for the lifting mechanism of the crane. Part IV details the mathematical model of the motor. Based on the theory and method of the part II, we obtained the AC induction motor's two completely decoupled second-order linear subsystems (rotor speed and flux linkage) in the part V. In part VI, two ADRCs with identical structures were designed for the rotor speed and flux linkage subsystems. In parts VII and VIII, important conclusions were drawn through experimental study.

II. RELATIVE ORDER VECTOR AND RELATIVE ORDER OF AFFINE NONLINEAR SYSTEM

Consider a multi-input multi-output (MIMO) affine nonlinear system,

$$\begin{cases} \dot{\boldsymbol{x}} = \boldsymbol{f}(\boldsymbol{x}) + \boldsymbol{g}(\boldsymbol{x})\boldsymbol{u} \\ \boldsymbol{y} = \boldsymbol{h}(\boldsymbol{x}), \end{cases} \quad (1)$$

where $\boldsymbol{x} \in \mathrm{R}^n$, $\boldsymbol{u} \in \mathrm{R}^m$, and $\boldsymbol{y} \in \mathrm{R}^m$ are the n-dimensional state variables and the m-dimensional input and output vectors, respectively. Moreover, $\boldsymbol{x} = [\, x_1 \; x_2 \cdots x_n \,]^{\mathrm{T}}$, $\boldsymbol{u} = [\, u_1 \; u_2 \cdots u_m \,]^{\mathrm{T}}$, and $\boldsymbol{y} = [\, y_1 \; y_2 \cdots y_m \,]^{\mathrm{T}}$. The functions $\boldsymbol{f}(\boldsymbol{x}) = [f_1(\boldsymbol{x})f_2(\boldsymbol{x}) \cdots f_n(\boldsymbol{x})]^T$ and $\boldsymbol{g}(\boldsymbol{x}) = [\, \boldsymbol{g}_1(\boldsymbol{x}) \; \boldsymbol{g}_2(\boldsymbol{x}) \cdots \boldsymbol{g}_m(\boldsymbol{x}) \,]^{\mathrm{T}}$ are n-dimensional and m-dimensional sufficiently smooth vector fields, respectively, where $f_i(\boldsymbol{x})(i = 1, 2, \cdots n)$ is scalar function and $\boldsymbol{g}_j(\boldsymbol{x})(j = 1, 2, \cdots m)$ is an n-dimensional column vector. The function $\boldsymbol{h}(\boldsymbol{x}) = [\, h_1(\boldsymbol{x}) \; h_2(\boldsymbol{x}) \cdots h_m(\boldsymbol{x}) \,]^{\mathrm{T}}$, where $h_j(\boldsymbol{x})$ $(j = 1, 2, \cdots m)$ is a smooth scalar function.

Each output in the MIMO system given by Eq. (1), $y_j = h_j(\boldsymbol{x})(j = 1, 2, \cdots m)$, has a corresponding relative order γ_j. When $\boldsymbol{x}$ is approximately the neighborhood of $\boldsymbol{x}(\boldsymbol{0})$, if the following conditions are true, $\boldsymbol{\gamma} = (\gamma_1 \; \gamma_2 \cdots \gamma_m)$ is called the relative order vector of Equation (1). The relative order

number is $\gamma_j (j = 1, 2, \cdots m)$, e.g., $k_i < \gamma_i - 1$,

$$L_{g_j} L_f^{k_i} h_j(\boldsymbol{x}) = 0 \quad (j = 1, 2, \cdots m), \tag{2}$$

where $L_f^{k_i} h_j(\boldsymbol{x})$ is the r-th order Lie derivative of the j-th output function $h_j(\boldsymbol{x})$ with respect to the vector field $\boldsymbol{f}(\boldsymbol{x})$, $L_{g_j} L_f^{k_i} h_j(\boldsymbol{x})$ is the Lie derivative of $L_f^{k_i} h_j(\boldsymbol{x})$ with respect to the j-th column n-dimensional column vector $\boldsymbol{g}_j(\boldsymbol{x})$.

The following matrix is defined:

$$\boldsymbol{D}(\boldsymbol{x})$$
$$= \begin{bmatrix} L_{g_1} L_f^{\gamma_1-1} h_1(\boldsymbol{x}) & L_{g_2} L_f^{\gamma_1-1} h_1(\boldsymbol{x}) & \cdots & L_{g_m} L_f^{\gamma_1-1} h_1(\boldsymbol{x}) \\ L_{g_1} L_f^{\gamma_2-1} h_2(\boldsymbol{x}) & L_{g_2} L_f^{\gamma_2-1} h_2(\boldsymbol{x}) & \cdots & L_{g_m} L_f^{\gamma_2-1} h_2(\boldsymbol{x}) \\ \vdots & \vdots & \vdots & \vdots \\ L_{g_1} L_f^{\gamma_m-1} h_m(\boldsymbol{x}) & L_{g_2} L_f^{\gamma_m-1} h_m(\boldsymbol{x}) & \cdots & L_{g_m} L_f^{\gamma_m-1} h_m(\boldsymbol{x}) \end{bmatrix}$$
$$\tag{3}$$

where $L_f^{\gamma_m-1} h_m(\boldsymbol{x})$ is the $(\gamma_m - 1)$-th order Lie derivative of the scalar function $h_m(\boldsymbol{x})$ with respect to the vector field $\boldsymbol{f}(\boldsymbol{x})$, i.e., the rate of change of $h_m(\boldsymbol{x})$ in the $\boldsymbol{f}(\boldsymbol{x})$ direction.

III. EQUIVALENT LOAD TORQUE AND SYSTEM DYNAMIC EQUATION OF CRANE'S HOIST MOTOR DRAG SYSTEM

The hoisting motor drive system's lifting mechanism of the crane is shown in Fig. 1(a). The motor output electromagnetic torque is T_e. The motor rotor speed n is transferred to the hoisting drum through the intermediate deceleration mechanism, and then the load ($m_L + m_0$, where m_0 is the mass of spreader) is lifted through the pulley block with speed v_L. The rotation radius and mass of the pulley block are not considered. When the lifting load falls, the motor operates in a state of power generation, and then the motor no longer consumes the electric energy from the power grid. In practice, energy-consumption or mechanical-brake braking is generally used to consume electric energy generated by the motor, so electric energy consumption by the motor is only considered when the lifting load rises.

To facilitate the calculation, the single-shaft motor drive system shown in Fig. 1 (b) is equivalent to the multi-shaft motor drive system shown in Fig. 1 (a), and the positive direction of each physical quantity is specified in Fig. 1 (c). Based on the principle that the kinetic energy of the transmission system is invariable, the equivalent moment of inertia of the motor's rotor shaft is obtained as follows:

$$J = J_e + J_1 \frac{1}{j_1^2} + J_2 \frac{1}{j_2^2} + J_r \frac{1}{j_r^2} + m\left(\frac{v_L}{n}\right)^2 + \frac{W_{fd}}{n^2}, \tag{4}$$

where J_e is the moment of inertia of the motor rotor; J_1, J_2, and J_r are the moments of inertia of intermediate shaft 1, intermediate shaft 2, and the lifting drum, respectively; $j_1 = \frac{n}{n_1}, j_2 = \frac{n}{n_2}$, and $j_r = \frac{n}{n_r}$ are the reduction ratios of the inertia of intermediate shaft 1, intermediate shaft 2, and the lifting drum; n is the motor's rotor speed; n_1, n_2, and n_r are the rotational speeds of intermediate shaft 1, intermediate shaft 2, and the lifting drum, respectively; $m = \left(\frac{30}{\pi}\right)^2 (m_0 + m_L)$ is an
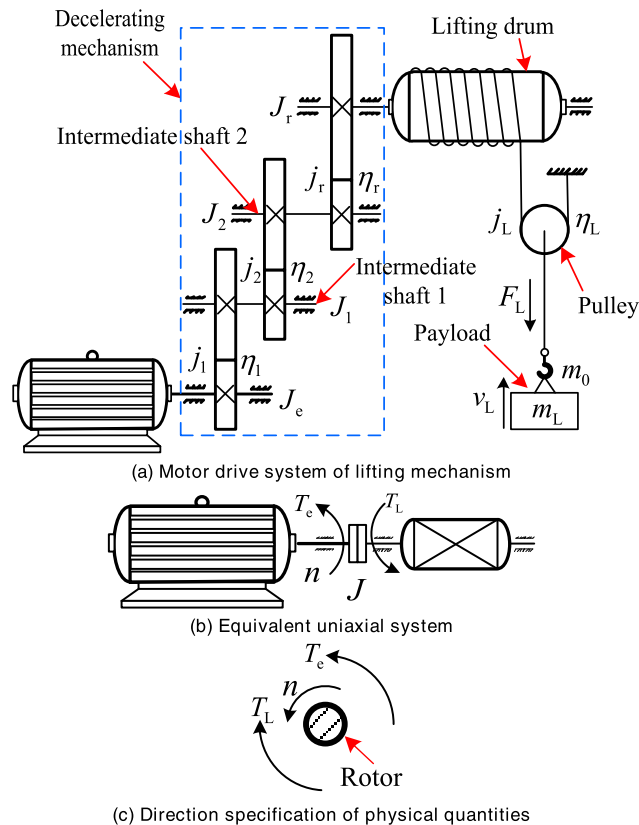

(a) Motor drive system of lifting mechanism

(b) Equivalent uniaxial system

(c) Direction specification of physical quantities

FIGURE 1. Equivalent single-shaft motor drive system of the crane.

equivalent mass; $W_{fd} = 2 \times \left(\frac{30}{\pi}\right)^2 W_f$ is an equivalent loss; and W_f is the loss caused by friction and other factors.

In general, $j_1^2 \gg 1, j_2^2 \gg 1, j_r^2 \gg 1$, and $\frac{1}{n^2} \approx 0$. According to Eq. (4), the equivalent rotational inertia J of the transmission mechanism can only depend on the rotational inertia J_e of the motor rotor, so $J \approx J_e$.

The equivalent load torque on the motor rotor shaft is referred to as the load torque,

$$T_L = \frac{DF_L}{2j_1 j_2 j_r j_L \eta_1 \eta_2 \eta_r \eta_L}, \tag{5}$$

where D is the lifting drum's diameter, $j_L = \frac{n}{n_L}$ is the pulley's deceleration ratio, $F_L = (m_0 + m_L)g$ is the wire rope's static tension, g is the gravitational acceleration, and η_1, η_2, η_r, and η_L are the transmission efficiencies of intermediate shaft 1, intermediate shaft 2, the lifting drum, and the pulley, respectively.

According to Eq. (5), when the structure of the transmission system is known, T_L is only related to F_L and has no relation to the motor rotor speed, which is the constant torque load characteristic of the crane. The analysis shows that the equivalent conversion method of the horizontal motion mechanism of the crane's cart and trolley is the same as that expressed by Eqs. (4) and (5).

According to Fig. 1 (b), the dynamic equation of the motor drag system can be obtained as follows:

$$J \frac{d\omega_r}{dt} = T_e - T_L, \tag{6}$$

where $\omega_r = \frac{\pi}{30}n$ is the mechanical angular velocity of the motor rotor, which is also known as the rotor mechanical angular frequency or rotor angular velocity. The electric angular velocity of the rotor is $\omega = n_p\omega_r$, and n_p is the motor's pole logarithm.

According to Eq. (6), when the motor's electromagnetic torque T_e is equal to the load torque T_L, the crane's lifting mechanism can operate smoothly. However, the load torque is often uncertain, which requires the electromagnetic torque T_e to quickly track the load torque T_L. The process of the electromagnetic torque T_e stabilizing at the load torque T_L is defined as the electromagnetic torque transition process of the motor, denoted as $[t_1, t_1 + t_s]$. In the electromagnetic torque transition process of the motor, t_1 is the moment of sudden load on the motor, and t_s is the time required for the transition process. The time is known as the motor electromagnetic torque stabilization time and will henceforth be referred to as the stabilization time. The motor's starting moment is $t_1 = 0$, with the corresponding t_s as the starting time. When t_s is smaller, the motor's stabilizing process is shorter, and the motor has a good load tracking ability. Therefore, the key for the lifting motor's control is to control the motor's electromagnetic torque T_e, so the electromagnetic torque T_e can be stabilized at the load torque T_L in a short time. Simply the tracking and adjusting time of T_e to T_L can be shortened and the stabilization time t_s can be reduced.

IV. MATHEMATICAL MODEL FOR THE THREE-PHASE AC INDUCTION MOTORS

The stator windings and rotor windings of the three-phase induction motor can be transformed into the static two-phase coordinate system α-β as shown in Fig. 2 by coordinate and rotation transformation. In Fig. 2, $u_{s\alpha}$ and $u_{s\beta}$ are stator voltages, whereas $u_{r\alpha}$ and $u_{r\beta}$ are rotor voltages. Considering that the cage rotor is short circuit, generally $u_{r\alpha} = u_{r\beta} = 0$. In addition, $i_{s\alpha}$ and $i_{s\beta}$ are the stator currents, whereas $i_{r\alpha}$ and $i_{r\beta}$ are the rotor currents. From Fig. 2, the voltage equation of the motor can be written as

$$\begin{bmatrix} u_{s\alpha} \\ u_{s\beta} \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 & 0 \\ 0 & R_s & 0 & 0 \\ 0 & 0 & R_r & 0 \\ 0 & 0 & 0 & R_r \end{bmatrix} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix}$$
$$+ \frac{d}{dt} \begin{bmatrix} \psi_{s\alpha} \\ \psi_{s\beta} \\ \psi_{r\alpha} \\ \psi_{r\beta} \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ \omega_r\psi_{r\beta} \\ -\omega_r\psi_{r\alpha} \end{bmatrix}, \tag{7}$$

where R_s and R_r are the stator and rotor resistances, $\psi_{s\alpha}$ and $\psi_{s\beta}$ are stator flux linkages, $\psi_{r\alpha}$ and $\psi_{r\beta}$ are rotor flux linkages.
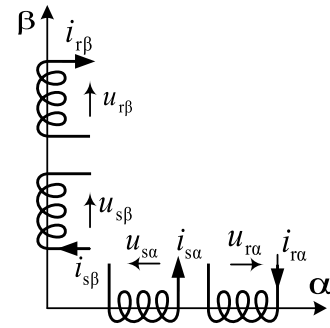

FIGURE 2. Static two-phase coordinate system α-β of the motor.

The flux equation of the motor can be written as

$$\begin{bmatrix} \psi_{s\alpha} \\ \psi_{s\beta} \\ \psi_{r\alpha} \\ \psi_{r\beta} \end{bmatrix} = \begin{bmatrix} L_s & 0 & L_m & 0 \\ 0 & L_s & 0 & L_m \\ L_m & 0 & L_r & 0 \\ 0 & L_m & 0 & L_r \end{bmatrix} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix}, \tag{8}$$

where L_s and L_r are the stator and rotor inductances, and L_m is the mutual inductance between the stator and rotor.

The torque equation of the motor is

$$T_e = n_p L_m (i_{s\beta} i_{r\alpha} - i_{s\alpha} i_{r\beta}). \tag{9}$$

The variables ω_r, $\psi_{r\alpha}$, $\psi_{r\beta}$, $i_{s\alpha}$, and $i_{s\beta}$ are selected as the state variables. According to Eqs. (6)–(9), the mathematical model of the motor can be written as

$$\begin{cases} \dfrac{d\omega_r}{dt} = k(\psi_{r\alpha}i_{s\beta} - \psi_{r\beta}i_{s\alpha}) - \dfrac{T_L}{J} \\ \dfrac{d\psi_{r\alpha}}{dt} = -\vartheta\psi_{r\alpha} - n_p\omega_r\psi_{r\beta} + \vartheta L_m i_{s\alpha} \\ \dfrac{d\psi_{r\beta}}{dt} = n_p\omega_r\psi_{r\alpha} - \vartheta\psi_{r\beta} + \vartheta L_m i_{s\beta} + \dfrac{u_{s\alpha}}{\delta} \\ \dfrac{di_{s\alpha}}{dt} = \vartheta\varepsilon\psi_{r\alpha} + n_p\varepsilon\omega_r\psi_{r\beta} - \xi i_{s\alpha} + \dfrac{u_{s\beta}}{\delta} \\ \dfrac{di_{s\beta}}{dt} = -n_p\varepsilon\omega_r\psi_{r\alpha} + \vartheta\varepsilon\psi_{r\beta} - \xi i_{s\beta}, \end{cases} \tag{10}$$

where $\delta = L_s - \frac{L_m^2}{L_r}$, $\vartheta = \frac{R_r}{L_r}$, $\varepsilon = \frac{L_m}{\delta L_r}$, $\xi = \frac{L_m^2 R_r}{\delta L_r^2} + \frac{R_s}{\delta}$, and $k = 1.5 \frac{n_p L_m}{J L_r}$.

V. PRECISE LINEARIZATION AND DECOUPLING OF DYNAMIC MODEL OF LIFTING MOTOR

The dynamic model (Eq.(10)) in the static two-phase coordinate system α-β can be described using Eq. (1), where the state variable vector is $x = [\omega_r\ \psi_{r\alpha}\ \psi_{r\beta}\ i_{s\alpha}\ i_{s\beta}]^T$, $h(x) = [\omega_r, \psi_{r\alpha}^2 + \psi_{r\beta}^2]^T$, the output $y_1 = \omega_r$ is the angular velocity of the rotor, and $y_2 = \psi_{r\alpha}^2 + \psi_{r\beta}^2$ is the sum of squares of the amplitude of the rotor flux, which is also referred to as the

rotor flux. The following functions are defined as

$$
f(x) = \begin{bmatrix} k(\psi_{r\alpha}i_{s\beta} - \psi_{r\beta}i_{s\alpha}) - \dfrac{T_L}{J} \\ -\vartheta\psi_{r\alpha} - n_p\omega_r\psi_{r\beta} + \vartheta L_m i_{s\alpha} \\ n_p\omega_r\psi_{r\alpha} - \vartheta\psi_{r\beta} + \vartheta L_m i_{s\beta} \\ \vartheta\varepsilon\psi_{r\alpha} + n_p\varepsilon\omega_r\psi_{r\beta} - \xi i_{s\alpha} \\ -n_p\varepsilon\omega_r\psi_{r\alpha} + \vartheta\varepsilon\psi_{r\beta} - \xi i_{s\beta} \end{bmatrix},
$$

$$
g(x) = [\, g_1 \quad g_2 \,] = \begin{bmatrix} 0 & 0 \\ 0 & 0 \\ 0 & 0 \\ \delta^{-1} & 0 \\ 0 & \delta^{-1} \end{bmatrix}, \quad u = \begin{bmatrix} u_{s\alpha} \\ u_{s\beta} \end{bmatrix}.
$$

According to Eq. (2), the relative order vector of the AC induction motor is $\gamma = (\gamma_1, \gamma_2)$, where $\gamma_1 = \gamma_2 = 2$, and

$$
D(x) = \begin{bmatrix} L_{g_1}L_f h_1(x) & L_{g_2}L_f h_1(x) \\ L_{g_1}L_f h_2(x) & L_{g_2}L_f h_2(x) \end{bmatrix}, \tag{11}
$$

where $L_{g_1}L_f h_1(x) = -\frac{k}{\delta}\psi_{r\beta}$, $L_{g_2}L_f h_1(x) = \frac{k}{\delta}\psi_{r\alpha}$, $L_{g_1}L_f h_2(x) = \frac{2\vartheta L_m}{\delta}\psi_{r\alpha}$, $L_{g_2}L_f h_2(x) = \frac{2\vartheta L_m}{\delta}\psi_{r\beta}$, $L_f h_1(x) = k(\psi_{r\alpha}i_{s\beta} - \psi_{r\beta}i_{s\alpha}) - \frac{T_L}{J}$, and $L_f h_2(x) = -2\vartheta(\psi_{r\alpha}^2 + \psi_{r\beta}^2) + 2\vartheta L_m(\psi_{r\alpha}i_{s\alpha} + \psi_{r\beta}i_{s\beta})$.

Because $\mathrm{Det}(D(x)) = -2\frac{\vartheta L_m k}{\delta^2}(\psi_{r\alpha}^2 + \psi_{r\beta}^2) \neq 0$, $D(x)$ is a nonsingular matrix, and $D(x)$'s inverse matrix is

$$
D^{-1}(x) = \frac{\delta}{\psi_{r\alpha}^2 + \psi_{r\beta}^2} \begin{bmatrix} -\dfrac{\psi_{r\beta}}{k} & \dfrac{\psi_{r\alpha}}{2\vartheta L_m} \\ \dfrac{\psi_{r\alpha}}{k} & \dfrac{\psi_{r\beta}}{2\vartheta L_m} \end{bmatrix}. \tag{12}
$$

The following coordinate transformation is defined as

$$
\begin{cases} z_1 = h_1(x) \\ z_2 = L_f h_1(x) \\ z_3 = h_2(x) \\ z_4 = L_f h_2(x) \\ z_5 = \arctan\left(\dfrac{\psi_{r\beta}}{\psi_{r\alpha}}\right), \end{cases} \tag{13}
$$

where $z_3 = y_2 = h(x) = \psi_{r\alpha}^2 + \psi_{r\beta}^2$ is the sum of squares of the amplitude of the rotor flux. In general, z_3 is not equal to zero if the motor is energized, and a magnetic field can be built into the motor. So, the inverse transformation is

$$
\begin{cases} \omega_r = z_1 \\ \psi_{r\alpha} = \sqrt{z_3}\cos z_5 \\ \psi_{r\beta} = \sqrt{z_3}\sin z_5 \\ \\ i_{s\alpha} = \dfrac{1}{\sqrt{z_3}}[\cos(z_5(\dfrac{z_4 + 2\vartheta z_3}{2\vartheta L_m})) - \dfrac{1}{k}\sin(z_5(z_2 + \dfrac{T_L}{J}))] \\ i_{s\beta} = \dfrac{1}{\sqrt{z_3}}[\sin(z_5(\dfrac{z_4 + 2\vartheta z_3}{2\vartheta L_m})) + \dfrac{1}{k}\cos(z_5(z_2 + \dfrac{T_L}{J}))] \end{cases} \tag{14}
$$

Therefore, in the new coordinate, the AC induction motor's dynamic model is

$$
\begin{cases} \dot{z}_1 = z_2 \\ \dot{z}_2 = L_f^2 h_1(x) + L_{g_1}L_f h_1(x)u_{s\alpha} + L_{g_2}L_f h_1(x)u_{s\beta} \\ \dot{z}_3 = z_4 \\ \dot{z}_4 = L_f^2 h_2(x) + L_{g_1}L_f h_2(x)u_{s\alpha} + L_{g_2}L_f h_2(x)u_{s\beta} \\ \dot{z}_5 = z_1 + \dfrac{\vartheta L_m}{kz_3}(z_2 + \dfrac{n_p}{J}T_L) \\ y_1 = z_1 \\ y_2 = z_3, \end{cases} \tag{15}
$$

where

$$
L_f^2 h_1(x) = -k\varepsilon n_p\omega_r(\psi_{r\alpha}^2 + \psi_{r\beta}^2) - k(\vartheta + \xi) \\
\times (\psi_{r\alpha}i_{s\beta} - \psi_{r\beta}i_{s\alpha}) - kn_p\omega_r(\psi_{r\alpha}i_{s\alpha} - \psi_{r\beta}i_{s\beta}),
$$

and

$$
L_f^2 h_2(x) = (4\vartheta^2 + 2\vartheta^2\varepsilon L_m)(\psi_{r\alpha}^2 + \psi_{r\beta}^2) \\
+ 2\vartheta L_m n_p\omega_r(\psi_{r\alpha}i_{s\beta} - \psi_{r\beta}i_{s\alpha}) \\
- (6\vartheta^2 L_m + 2\vartheta\xi L_m)(\psi_{r\alpha}i_{s\alpha} + \psi_{r\beta}i_{s\beta}) \\
+ 2\vartheta^2 L_m^2(i_{s\alpha}^2 + i_{s\beta}^2).
$$

The first four equations can be written based on Eq. (11) as follows:

$$
\begin{bmatrix} \ddot{y}_1 \\ \ddot{y}_2 \end{bmatrix} = \begin{bmatrix} L_f^2 h_1(x) \\ L_f^2 h_2(x) \end{bmatrix} + D(x)\begin{bmatrix} u_{s\alpha} \\ u_{s\beta} \end{bmatrix}. \tag{16}
$$

The virtual control inputs are defined as v_1 and v_2, and

$$
\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} L_f^2 h_1(x) \\ L_f^2 h_2(x) \end{bmatrix} + D(x)\begin{bmatrix} u_{s\alpha} \\ u_{s\beta} \end{bmatrix}. \tag{17}
$$

The actual control input of the motor is

$$
\begin{bmatrix} u_{s\alpha} \\ u_{s\beta} \end{bmatrix} = D^{-1}(x)(\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} - \begin{bmatrix} L_f^2 h_1(x) \\ L_f^2 h_2(x) \end{bmatrix}). \tag{18}
$$

Based on Eqs. (16)–(18), we obtain the AC induction motor's two completely decoupled second-order linear subsystems, rotor speed subsystem I and flux linkage subsystem II, as follows:

$$
\text{I:} \quad \begin{cases} \dot{z}_1 = z_2 \\ \dot{z}_2 = v_1 \\ y_1 = z_1, \end{cases} \tag{19-a}
$$

$$
\text{II:} \quad \begin{cases} \dot{z}_3 = z_4 \\ \dot{z}_4 = v_2 \\ y_2 = z_3. \end{cases} \tag{19-b}
$$

The virtual control inputs, v_1 and v_2, can independently control the motor rotor speed $y_1 = \omega_r$, and the sum of squares of the rotor flux amplitude, $y_2 = \psi_{r\alpha}^2 + \psi_{r\beta}^2$, which greatly reduces the design difficulty of the controller. Moreover, the matrix $\begin{bmatrix} L_f^2 h_1(x) \\ L_f^2 h_2(x) \end{bmatrix}$ is called the decoupling device, and the matrix $D^{-1}(x)$ serves as the inverter.
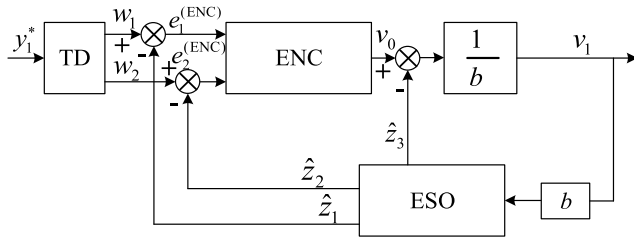
FIGURE 3. Block diagram of speed ADRC's structure.

VI. DESIGN OF THE ADRC FOR THE CRANE MOTOR DECOUPLING SYSTEM

The stator or rotor resistance varies in the vicinity of the rated value because of the motor temperature rise, and the uncertainty of the load torque magnitude. Thus, the goal of the AC induction motor control was to control the stator voltages $u_{s\alpha}$ and $u_{s\beta}$. When the rotor speed ω_r and the sum of squares of the amplitude of the rotor flux $y_2 = \psi_{r\alpha}^2 + \psi_{r\beta}^2$ are given by reference values, the motor's electromagnetic torque T_e can quickly track the unknown load torque T_L. Therefore, with regard to the second-order linear subsystems I and II, we designed two controllers with the same structure: motor speed ADRC and flux ADRC. The motor speed ADRC included a tracking differentiation (TD), error nonlinear combiner (ENC), and extended state observer (ESO), as shown in Fig. 3. The function of each part and the discrete form of the algorithm are illustrated through the speed ADRC as an example.

(1) Tracking differentiator: The transition process is arranged according to the set reference value, and the differential signal is extracted. The algorithm is as follows:

$$\begin{cases} e^{(TD)} = w_1 - y_1^* \\ w_1(k+1) = w_1(k) + Tw_2(k) \\ w_2(k+1) = w_2(k) + T\text{fhan}(e^{(TD)}, w_2, r, h), \end{cases} \quad (20)$$

where, w_1 and w_2 are the TD's outputs, $e^{(TD)}$ is the tracking deviation of a variable from a given reference value, r is the speed factor that determines how fast the variable can track a given reference value, h is the filtering factor that determines the filtering effect, and T is the integral step. The function fhan of variables m, n, r, and h is denoted as

$$\text{fhan}(m, n, r, h) = -r\left(\frac{a}{d}\right)\text{fsg}(a, d) - r\,\text{sign}(a) \\ \times (1 - \text{fsg}(a, d)), \quad (21)$$

where $d = rh^2$, $a_0 = hn$, $p = m + a_0$, $a_1 = \sqrt{d(d + 8\,|p|)}$, $a_2 = a_0 + 0.5\,\text{sign}(p)(a_1 - d)$, and $a = (a_0 + p)\text{fsg}(p, d) + a_2(1 - \text{fsg}(p, d))$. The function $\text{fsg}(p, d) = 0.5(\text{sign}(p+d) - \text{sign}(p - d))$ depends on variables p and d.

(2) Extended state observer: The ESO estimates the rotor speed subsystem state variables z_1 and z_2 and expands the unmodeled dynamics of the motor, motor parameter changes caused by heating and other sources, and unknown disturbances of the system into new state variable z_3 (hereinafter

referred to as the expanded state). The continuous-time model of ESO is as follows:

$$\begin{cases} \dot{\hat{z}}_1 = \hat{z}_2 - \beta_1^{(ESO)}e^{(ESO)} \\ \dot{\hat{z}}_2 = \hat{z}_3 - \beta_2^{(ESO)}fe + bv_1 \\ \dot{\hat{z}}_3 = -\beta_3^{(ESO)}fe_1, \end{cases} \quad (22)$$

where $\boldsymbol{\beta}^{(ESO)} = [\beta_1^{(ESO)}, \beta_2^{(ESO)}, \beta_3^{(ESO)}]$ is the parameter of ESO, and $e^{(ESO)} = \hat{z}_1 - y_1$.

Furthermore, Eq. (22) can estimate z_3, and the algorithm is as follows:

$$\begin{cases} e^{(ESO)} = \hat{z}_1 - y_1 \\ \hat{z}_1(k+1) = \hat{z}_1(k) + T(\hat{z}_2(k) - \beta_1^{(ESO)}e^{(ESO)}) \\ \hat{z}_2(k+1) = \hat{z}_2(k) + T(\hat{z}_3(k) - \beta_2^{(ESO)}fe + bv_1) \\ \hat{z}_3(k+1) = \hat{z}_3(k) + T(-\beta_3^{(ESO)}fe_1), \end{cases} \quad (23)$$

where $\hat{z}_1$, $\hat{z}_2$, and $\hat{z}_3$ are the estimated values of z_1, z_2, and the expansion state z_3, respectively. Parameter b is the compensation factor of the control strength. The following functions are defined as $fe = \text{fal}(e^{(ESO)}, 0.5, d)$ and $fe_1 = \text{fal}(e^{(ESO)}, 0.25, d)$, and fal is a function of variables e, a, and d, and defined as follows:

$$\text{fal}(e, a, d) = ed^{a-1}\text{fsg}(e, d) + |e|^a\,\text{sign}(e)(1 - \text{fsg}(e, d)). \quad (24)$$

(3) Error nonlinear combiner: The ENC realizes the nonlinear combination of the error feedback, and the algorithm is as follows:

$$v_0 = \beta_1^{(ENC)}\text{fal}(e_1^{(ENC)}, 0.75, d) \\ + \beta_2^{(ENC)}\text{fal}(e_2^{(ENC)}, 1.25, d), \quad (25)$$

where $e_1^{(ENC)} = w_1 - \hat{z}_1$ and $e_2^{(ENC)} = w_2 - \hat{z}_2$.

The virtual control input of the rotor speed subsystem is

$$v_1 = v_0 - \frac{\hat{z}_3}{b}. \quad (26)$$

Using the same algorithm and Eqs. (20)–(26), the AC induction motor flux linkage ADRC can also be designed, in the exact same manner as the motor speed ADRC.

The principle block diagram of the motor ADRC system based on accurate decoupling is shown in Fig. 4. As shown in Fig. 4, the decoupling device achieves precise decoupling of the motor dynamics model. The ADRC outputs are the virtual control v_1 of the speed subsystem and the virtual control v_2 of the flux linkage subsystem The inverter calculates the actual control voltages of motor stator $u_{s\alpha}$ and $u_{s\beta}$ using Eq. (18), and $u_{s\alpha}$ and $u_{s\beta}$ are transformed into the PWM's three-phase modulation signals u_A^*, u_B^*, and u_C^*, through the transformation from the two-phase coordinate $\alpha - \beta$ to the three-phase coordinate A-B-C.

Finally, we obtain the actual three-phase control voltages of the motor u_A, u_B, and u_C. The rotor flux $\psi_{r\alpha}$ and $\psi_{r\beta}$ can
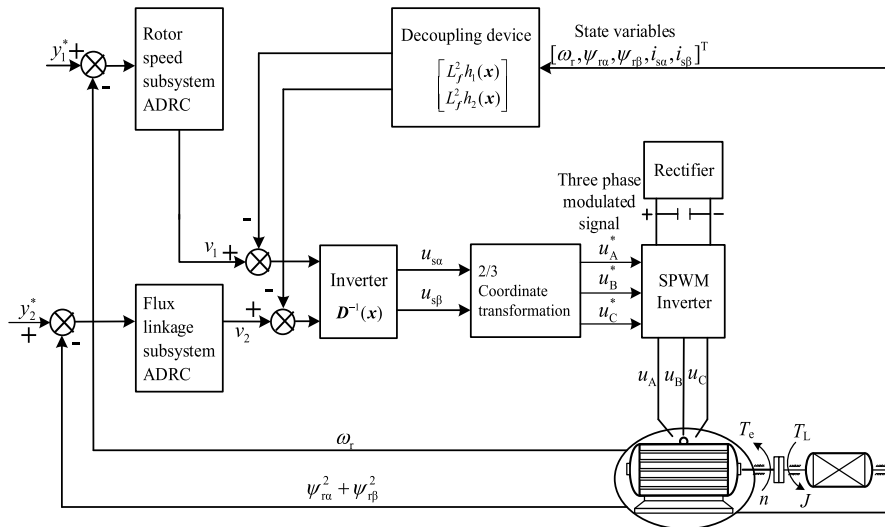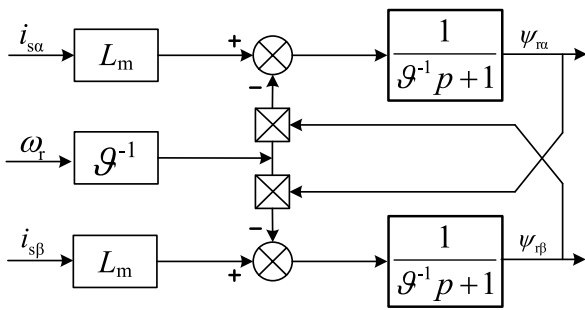
FIGURE 4. Schematic diagram of ADRC system.

FIGURE 5. Calculation block diagram of the rotor flux linkage.

be calculated by the stator current $i_{s\alpha}$ and $i_{s\beta}$, and

$$\begin{cases} \psi_{r\alpha} = \dfrac{1}{\vartheta^{-1}p + 1}(L_m i_{s\alpha} - \omega_r \vartheta^{-1} \psi_{r\beta}) \\ \psi_{r\beta} = \dfrac{1}{\vartheta^{-1}p + 1}(L_m i_{s\beta} - \omega_r \vartheta^{-1} \psi_{r\alpha}), \end{cases} \quad (27)$$

where $p = \frac{d}{dt}$ is the differential operator.

The calculation block diagrams of $\psi_{r\alpha}$ and $\psi_{r\beta}$ are shown in Fig. 5. The computer flow chart is shown in Fig. 6.

VII. EXPERIMENTAL STUDY

The experimental platform with container gantry crane is shown in Fig. 7. Different loads were provided with containers with different qualities for the hoisting motor drive system of the crane. The main controller was YASKAWA MP3300. The sampling time was 10 ms. The motor parameters were as follows: $R_s = 0.082 \ \Omega$, $R_r = 0.05 \ \Omega$, $L_s = 3.44$ mH, $L_r = 3.44$ mH, $L_m = 0.027$ mH, $n_p = 2$, and $J = 0.37$ kg·m^2. The nominal torque was $T_N = 80$ N·m. The nominal speed was $n_N = 1480$ r/min=155 rad/s.

The ADRC parameters were as follows: for the rotor speed subsystem, $r = 500$, $h = 0.01$, $T = 0.001$, $\boldsymbol{\beta}^{(ESO)} = [800, 562, 200]$, $\boldsymbol{\beta}^{(ENC)} = [3000, 26]$, $d = 0.01$, and $b = 1$.

For the flux linkage subsystem, they are $r = 600$, $\boldsymbol{\beta}^{(ESO)} = [2000, 2000, 2000]$, $\boldsymbol{\beta}^{(ENC)} = [1000, 100]$, $d = 0.01$, and $b = 1$.

A. VARIABLE DYNAMIC RESPONSE

To study the control effect of the closed-loop system with ADRC, we selected the motor load starting operation. Figs. 8 and 9 show the comparative results of the motor's rotational speed, flux linkage, and electromagnetic torque dynamic response of the open-loop system (no control system) and the closed-loop system when the motor was started with the rated load. The speed of the open-loop system was stable at the rated speed at about 1.3 s, and the flux was stable at 0.94 Wb^2 at about 1.3 s. The closed-loop system could reach the reference speed of 104.7 rad/s at about 1.3 s without overshoot, and the flux could reach the reference flux of 0.81 Wb^2 at about 0.3 s without overshoot.

The dynamic response of the electromagnetic torque is shown in Fig. 10. The electromagnetic torque of the open-loop system stabilized at the load torque when $t_s = 1.3$ s in the range of -250 to 450 N·m. The electromagnetic torque of the closed-loop system could be stabilized at the load torque when $t_s = 0.7$ s in the range of 0–250 N·m. The electromagnetic torque fluctuation range and adjustment time of the closed-loop system were significantly lower than those of the open-loop system, so the starting time was reduced by ~50%. The histograms of the starting times are shown in Fig. 11 where the motor was started at different load rates. From Fig. 11, the open-loop system's starting time increased with increasing load rate. For the closed-loop system, the starting time was within the range of 0.6–0.8 s within the rated load range, which was less than the starting time of the open-loop system.

Based on the different percentages of the motor rated torque (load rate) range, the motor's load was divided
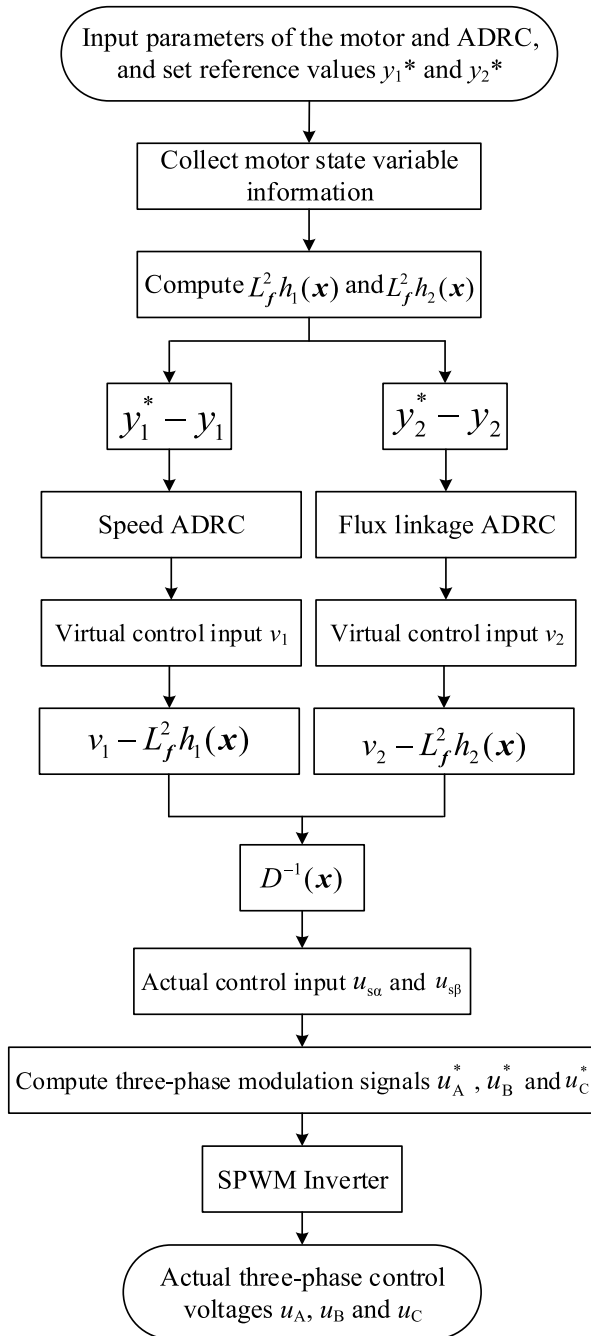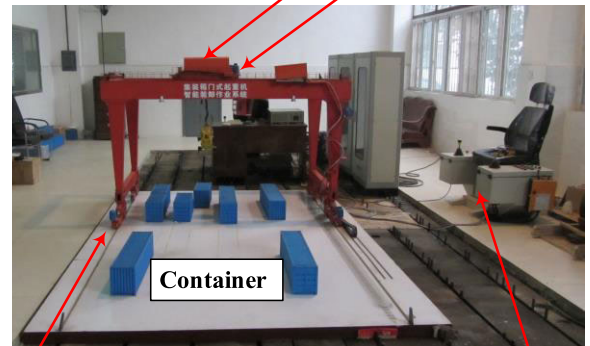
Input parameters of the motor and ADRC, and set reference values y_1* and y_2*

↓

Collect motor state variable information

↓

Compute $L_f^2 h_1(\boldsymbol{x})$ and $L_f^2 h_2(\boldsymbol{x})$

↓

$y_1^* - y_1$ $y_2^* - y_2$

↓   ↓

Speed ADRC   Flux linkage ADRC

↓   ↓

Virtual control input v_1 Virtual control input v_2

↓   ↓

$v_1 - L_f^2 h_1(\boldsymbol{x})$ $v_2 - L_f^2 h_2(\boldsymbol{x})$

↓

$D^{-1}(\boldsymbol{x})$

↓

Actual control input $u_{s\alpha}$ and $u_{s\beta}$

↓

Compute three-phase modulation signals u_A^*, u_B^* and u_C^*

↓

SPWM Inverter

↓

Actual three-phase control voltages u_A, u_B and u_C

FIGURE 6. Computer flow chart.

YASKAWA MP3300

FIGURE 7. Experimental platform.

into a light load ($T_L \leq 30\%T_N$), medium-light load ($30\%T_N < T_L < 50\%T_N$), medium-load ($T_L = 50\%T_N$), medium-heavy load ($50\%T_N < T_L < 80\%T_N$), heavy load ($80\%T_N \leq T_L < 100\%T_N$), and full load ($T_L = 100\%T_N$). To study the motor's transition process when the motor was suddenly loaded, the motor was initially operated under a light load ($T_L = 24$ N · m) with motor subjected to a sudden load at $t_1 = 2$s. The rotor flux of the closed-loop system was not affected by the sudden load and was always stable at the reference flux (as shown in Fig. 12).

The rotor speed decreased due to the sudden load, but the rate of decrease did not exceed 7 rad/s (as shown in Fig. 13). The electromagnetic torque fluctuated when the load was suddenly applied, as shown in Fig. 14. The experiment showed that the fluctuation range of the electromagnetic torque did not exceed 20 N·m, and the fluctuation range decreased as the load rate decreased. The transition time was approximately $t_s = 0.5$ s, as shown in Fig. 14, where $E_T = T_L - T_e$ is the electromagnetic torque tracking error. Figs. 12 and 13 show that the control system achieved independent and decoupled control of the motor speed and flux linkage.

B. ENERGY CONSUMPTION EXPERIMENT
To analyze the ADRC closed-loop control system's energy consumption, we set motor's starting stage with different load
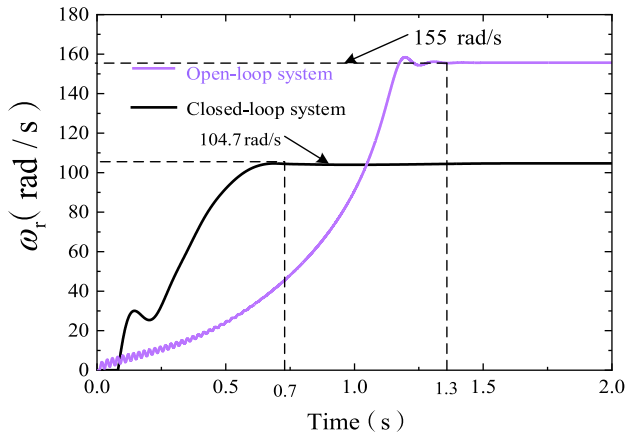
FIGURE 8. Dynamic response of the speed during the starting process at full load.

FIGURE 9. Dynamic response of the rotor flux during the starting process at full load.

FIGURE 10. Dynamic response of the electromagnetic torque during the starting process at full load.
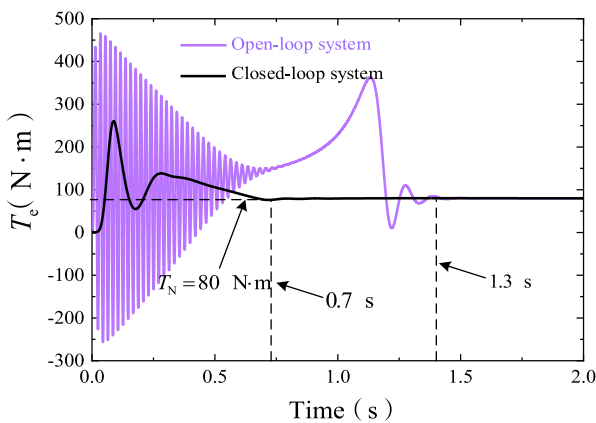

FIGURE 11. Comparison of the motor starting time.

FIGURE 12. Dynamic response of the flux under abrupt load changes.
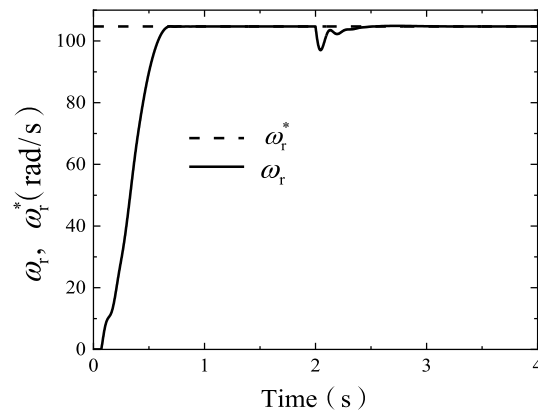
FIGURE 13. Dynamic response of the speed under abrupt load changes.

rates and the transition stage with different sudden loads. In the load's descending stage, the lifting mechanism mostly used the mechanical brake to decelerate the lifting mechanism's operation, and the motor was in a state of power generation. Thus, the energy consumption was studied only

during the lifting process. The motor's instantaneous power consumption is

$$P(t) = 3I(t)U(t), \tag{28}$$

where $I(t)$ and $U(t)$ are the motor's effective values of the stator current and voltage of a single phase, respectively.
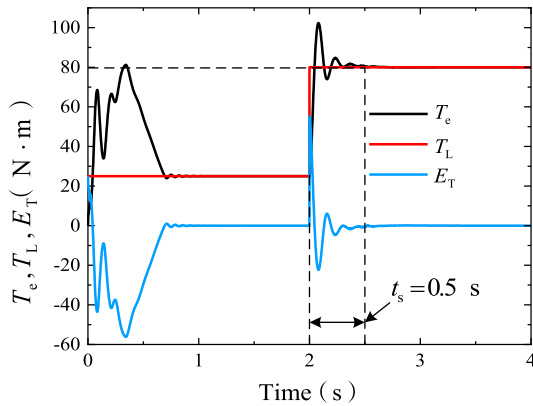
FIGURE 14. Dynamic response of the electromagnetic torque under abrupt load changes.
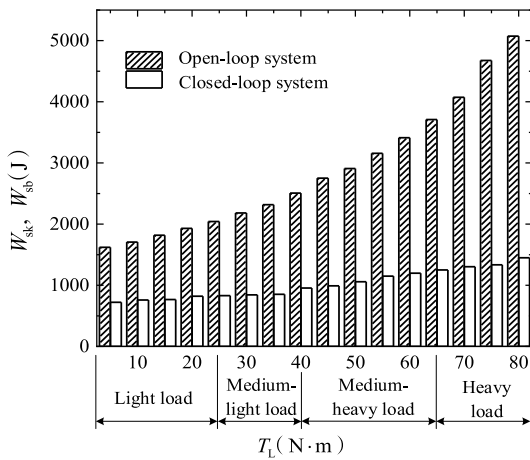
FIGURE 15. Comparison of motor starting energy consumption.

The energy consumption of the electromagnetic torque transition process $[t_1, t_1 + t_s]$ is defined as the energy consumption of the transition process,

$$W_s = \int_{t_1}^{t_1+t_s} P(t)\mathrm{d}t. \tag{29}$$

If $t_1 = 0$, the corresponding t_s is the starting time, and W_s is the starting energy consumption. According to Eq. (29), if $P(t)$ is constant, when t_s is smaller, the motor's starting energy consumption at the same load rate is smaller, and the motor saves more energy.

The power saving rate is defined as

$$\Delta E = \frac{W_{sk} - W_{sb}}{W_{sk}} \times 100\%, \tag{30}$$

where W_{sk} and W_{sb} are the starting energy consumption ($t_1 = 0$) or transition energy consumption ($t_1 \neq 0$) of the open-loop system and the closed-loop system, respectively. A bar chart of the motor starting energy consumption under different loads is shown in Fig. 15. The energy consumption of the closed-loop system during the starting process was slightly larger than 1000 J under heavy load and smaller
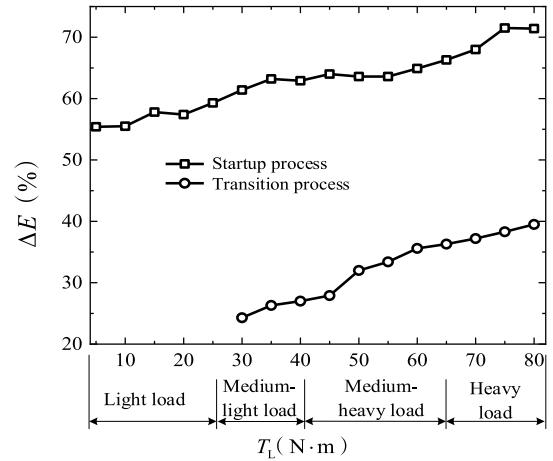
FIGURE 16. Power saving rate of the closed-loop system during the startup and transition.

than 1000 J under the other loads. For the open-loop system, the starting energy consumption was greater than that of the closed-loop system, and it increased with increasing the load rate, especially after a medium load.

The power saving rates of the closed-loop system during the starting and transition processes are shown in Fig. 16.

The closed-loop system had a significant energy saving effect during the starting process. Compared with the open-loop system, the power saving rate significantly increased as the load increased. The power saving rate was about 55%–59% under the light load and could reach 71% during the heavy load. The closed-loop system also had a significant energy saving effect in the transition process. The power saving rate was about 25% under the medium load and could reach 40% under the heavy load. This was because the closed-loop system reduced the starting time and increased the tracking speed of the electromagnetic torque to the load torque under the action of the ADRC.

During the starting process of the crane with a load, energy saving is an important factor. The ADRC closed-loop control system had a significant advantage in energy savings for the crane's starting process with a load.

C. ROBUSTNESS EXPERIMENT
Compared with the inductance of the motor's winding, motor parameters such as the stator and rotor resistances are more likely to change because of temperature changes during operation. Furthermore, unknown disturbances in the system, reduce the accuracy of the motor dynamic model and can diminish the control performance of the control system designed for the nominal dynamics.

The ADRC system can expand the unknown model, such as the motor's unmodeled dynamics, the motor's parameter changes, and the system's unknown disturbances into a new state variable. Furthermore, it can estimate the unknown model (denoted as $\hat{z}_3$) of the rotor speed subsystem and the unknown model (denoted as $\hat{z}_3'$) of the flux linkage subsystem,
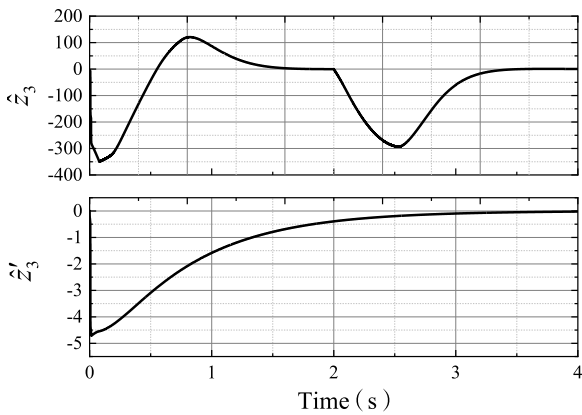
FIGURE 17. Estimation information of unknown model of motor speed and flux linkage subsystems.
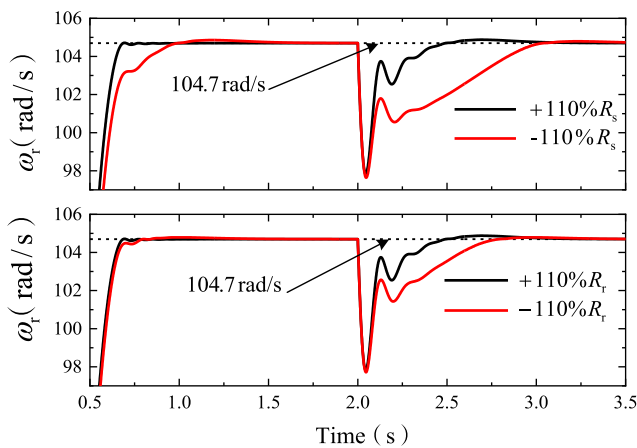

FIGURE 18. Rotor speed variations as the resistance of the stator or rotor varied.
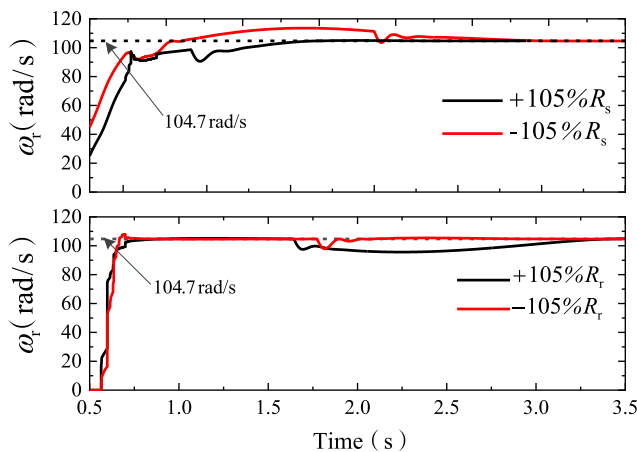

FIGURE 19. Rotor speed variations as the resistance of the stator or rotor varied.

as shown in Fig. 17. During the experiment, the load torque changed as shown in Fig. 14, and the motor stator and rotor resistances changed within ±10%. The motor speed changes are shown in Fig. 18, which indicate the ADRC control

system could adapt to the motor parameter changes within a certain range. This also highlights the robustness of the control system.

ADRC is developed on the basis of a PID controller. If we used the PID control method, detailed by the paper [53] and the load torque change as shown in Fig. 14, the motor stator and rotor resistances changed less than ±5%. The motor speed changes are shown in Fig. 19, which further highlight the robustness of the ADRC control system.

VIII. CONCLUSION AND FUTRUE WORK

A. CONCLUSION

The nonlinear feedback linearization method was used to decouple the motor's nominal dynamics model into two simple and linear subsystems: a rotor speed subsystem and a flux linkage subsystem. Two ADRC controllers with identical structures were designed for the two decoupled linear subsystems. The unmodeled dynamics of the motor, the variation of the motor parameters due to heating, and the unknown disturbances of the motor system were estimated by the ESO in the ADRC. Through a comparative experimental study of the closed-loop and the open-loop systems, the following important conclusions were drawn.

1. Compared with the open-loop system, the motor's full-load starting time was reduced by about 50%.
2. When the motor ran smoothly at different load rates and the rated load was suddenly applied, the electromagnetic torque fluctuation range did not exceed 20 N·m. The rotor flux was always stable at the reference value. The motor speed decreased, but the amount of decrease did not exceed 7 rad/s. The nominal dynamics model of the motor was completely decoupled into two independent subsystems of the rotor speed and flux linkage.
3. The starting time of the closed-loop system within the rated load range was about 0.6–0.8 s, and the starting energy consumption was less than 1000 J. The closed-loop system had a significant energy-saving effect during the motor's starting process. The power saving rate was about 55%–59% if the motor started with a light load. The power saving rate could reach 71% if the motor started with a heavy load.
4. The ADRC system could estimate the unknown model of the rotor speed and flux linkage subsystems. Furthermore, the ADRC system could adapt to the parameter variations of the motor stator and rotor resistances in the range of ±10% and exhibited high robustness.

B. FUTURE WORK

In the future, we plan to apply the theoretical research results of this paper to engineering practice. To achieve this purpose, at least three key pieces of technology need to be developed. The first is real-time monitoring or estimation of crane load torque and motor running parameters. The second is the design of hardware and programming of software for the control system. The third is the on-line setting and adjustment method of ADRC's parameters.

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