

# Backstepping control of two-mass system using induction motor drive fed by voltage source inverter with ideal control performance of stator current

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## ABSTRACT

This paper describes the design and the simulation of a non-linear controller for two-mass system using induction motor basing on the backstepping method. The aim is to control the speed actual value of load motor matching with the speed reference load motor, moreover, electrical drive's response ensuring the "fast, accurate and small overshoot" and reducing the resonance oscillations for two-mass system using induction motor fed by voltage source inverter with ideally control performance of stator current. Backstepping controller uses the non-linear equations of an induction motor and the linear dynamical equations of two-mass system, the Lyapunov analysis and the errors between the real and the desired values. The controller has been implemented in both simulation and hardware-in-the-loop (HIL) real-time experiments using Typhoon HIL 402 system, when the drive system operates at a stable speed (rotor flux is constant) and greater than rated speed (field weakening area). The simulation and HIL results presented the correctness and effectiveness of the controller is proposed; furthermore, compared to PI method to see the response of the system clearly.

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## 1. INTRODUCTION

The two-mass system with flexible shaft can be found in many industrial applications such as machine tools, rolling-mills, robot arms, conveyer belts, etc. Due to the finite stiffness and damping coefficients of the shaft, the speed oscillation of the drive system is inevitable and may cause detrimental influences on not only the quality of the products but also the mechanical and electrical components of the drive system. Hence, speed control design for these drive systems plays a key role in improving the system characteristic and products quality. Suppressing the shaft torsional vibration, rejecting the load disturbance torque and tracking the reference speed are main goals of the controller [1-4]. However, these tasks are not easy to be achieved due to the fact that not all system state variables are available. Furthermore, unknown nonlinearities such as friction and backlash may also cause undesirable inaccuracy for the control system.

Various control schemes have been developed in the literature to suppress the torsional vibration of the aforementioned two-mass system. Obviously, the classical cascade structure using proportional-integral (PI) speed controller is the most popular control scheme. However, this method cannot effectively suppress the vibration since the PI controller does not have enough degree-of-freedom to handle such high order dynamic system like this [5]. Thus, advanced control techniques such as sliding mode control, flatness-based control, adaptive control, and fuzzy adaptive control are also employed to get a better performance [6-11].

In this research, the field-oriented control (FOC) which has been widely used in industrial inverters is used to control the induction motor of the drive system [12]. Based on an observation that if the response of the stator current controller is fast enough, the order of the two-mass dynamic system can be reduced from 7th to 3rd in nominal speed range and from 7th to 4th in field weakening range. Consequently, the control design for the reduced order system can be simplified. Since the dead-beat controller, which forces the stator current to its desired value in finite step, has already successfully been developed for the stator current loop in our previous study [13], the remaining objective of this research is to achieve a high dynamic speed control with torsional oscillation reduction for the drive system.

In this paper, a backstepping control algorithm is proposed to control the load speed of two-mass systems. For achieving this goal, a two-block control strategy is proposed. It is backstepping control algorithm for load speed and motor speed. On the other hand, rotor flux control is designed according to backstepping control algorithm. Generating the desired motor torque by dead-beat current control, not only guarantees accurate load speed but also proves the stability of the system [14, 15].

## 2. MODELING OF DRIVE SYSTEM

### 2.1. Mathematical model of three phase induction motor

In field synchronous coordinate, the three-phase induction motor can be described by the following dynamic equations [12].

$$\begin{cases} \frac{di_{sd}}{dt} = -\left(\frac{1}{\sigma T_s} + \frac{1-\sigma}{\sigma T_r}\right)i_{sd} + \omega_s i_{sq} + \frac{1-\sigma}{\sigma T_r} + \frac{1}{\sigma L_s} u_{sd} \\ \frac{di_{sq}}{dt} = -\omega_s i_{sd} - \left(\frac{1}{\sigma T_s} + \frac{1-\sigma}{\sigma T_r}\right)i_{sq} - \frac{1-\sigma}{\sigma} \omega_m + \frac{1}{\sigma L_s} u_{sq} \\ \frac{di_m}{dt} = -\frac{1}{T_r} i_m + \frac{1}{T_r} i_{sd} \\ \frac{d\omega}{dt} = k i_m i_{sq} - \frac{z_p}{J} m_L \end{cases} \quad (1)$$

In which,  $i_{sd}, i_{sq}$  are  $dq$  components of the stator current;  $\omega, \omega_s$  are mechanical rotor velocity and synchronous speed, respectively;  $\psi_{rd}, \psi_{rq}$  are  $dq$  components of the rotor flux;  $\sigma$  is total leakage factor;  $T_r$  is rotor time constant;  $u_{sd}, u_{sq}$  are  $dq$  components of the stator voltage;  $L_s$  is stator inductance.

It can be seen that the original state (1) is bilinear and is of 4<sup>th</sup> order. Assume that the current controller is perfect with ideal response, the induction motor model can be reduced as:

$$\begin{cases} \frac{di_m}{dt} = -\frac{1}{T_r} i_m + \frac{1}{T_r} i_{sd} \\ \frac{d\omega}{dt} = k i_m i_{sq} - \frac{z_p}{J} m_L \end{cases} \quad (2)$$

Where:

$$i_m = \frac{\psi_{rd}}{L_m}; k = \frac{3 z_p^2 L_m^2}{2 L_r J}$$

In which,  $L_m, L_r$  is mutual, rotor inductance;  $i_m$  is vector of magnetizing current;  $m_L$  is torque load;  $z_p$  is number of pole pairs;  $J$  is torque of inertia;

The state (2) is of 2<sup>nd</sup> order, stator current  $i_{sd}$  is used to control the motor flux and  $i_{sq}$  is dedicated to speed control.

### 2.2. Model of the two-mass system

Typical configuration of a two-mass system show as Figure 1:

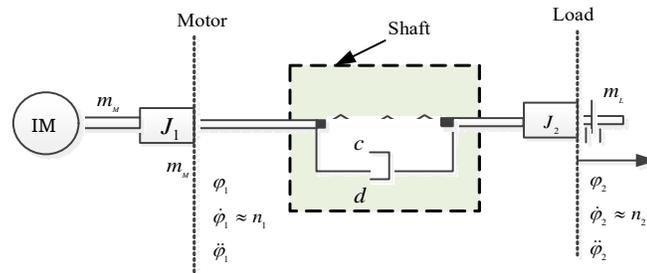


Figure 1. Typical configuration of a two-mass system

The typical configuration of a two-mass system is illustrated in Figure 1. The system can be described by the following linear dynamical equations [2].

$$\begin{cases} \ddot{\varphi}_1 = -\frac{d}{J_1}\dot{\varphi}_1 - \frac{c}{J_1}\Delta\varphi + \frac{d}{J_1}\dot{\varphi}_2 + \frac{1}{J_1}m_M \\ \Delta\dot{\varphi} = \dot{\varphi}_1 - \dot{\varphi}_2 \\ \ddot{\varphi}_2 = \frac{d}{J_2}\dot{\varphi}_1 + \frac{c}{J_2}\Delta\varphi - \frac{d}{J_2}\dot{\varphi}_2 - \frac{1}{J_2}m_L \end{cases} \quad (3)$$

Where:

$\dot{\varphi}_1, \dot{\varphi}_2$  Are motor speed, load speed;  $\ddot{\varphi}_1, \ddot{\varphi}_2$  are motor and load angle acceleration;  $\varphi$  is angle;  $d$  is shaft damping coefficient;  $c$  is shaft stiffness.

When the drive system operates at field weakening range, the mathematical model of the system is [2].

$$\begin{cases} \frac{di_m}{dt} = \frac{1}{T_r}i_{sd} - \frac{1}{T_r}i_m \\ \ddot{\varphi}_1 = -\frac{d}{J_1}\dot{\varphi}_1 - \frac{c}{J_1}\Delta\varphi + \frac{d}{J_1}\dot{\varphi}_2 + \frac{1}{J_1}m_M \\ \Delta\dot{\varphi} = \dot{\varphi}_1 - \dot{\varphi}_2 \\ \ddot{\varphi}_2 = \frac{d}{J_2}\dot{\varphi}_1 + \frac{c}{J_2}\Delta\varphi - \frac{d}{J_2}\dot{\varphi}_2 - \frac{1}{J_2}m_L \end{cases} \quad (4)$$

For control design purpose, term (4) is rewritten in the following state-space form:

$$\begin{bmatrix} \dot{i}_m \\ \dot{\varphi}_1 \\ \Delta\dot{\varphi} \\ \dot{\varphi}_2 \end{bmatrix} = \begin{bmatrix} -\frac{1}{T_r} & 0 & 0 & 0 \\ 0 & -\frac{d}{J_1} & -\frac{c}{J_1} & \frac{d}{J_1} \\ 0 & 1 & 0 & -1 \\ 0 & \frac{d}{J_2} & \frac{c}{J_2} & -\frac{d}{J_2} \end{bmatrix} \begin{bmatrix} i_m \\ \varphi_1 \\ \Delta\varphi \\ \varphi_2 \end{bmatrix} + \begin{bmatrix} \frac{1}{T_r} & 0 \\ 0 & \frac{k_w i_m}{J_1} \\ 0 & 0 \\ 0 & 0 \end{bmatrix} \begin{bmatrix} i_{sd} \\ i_{sq} \end{bmatrix} + \begin{bmatrix} 0 \\ 0 \\ 0 \\ -\frac{1}{J_2} \end{bmatrix} m_L \quad (5)$$

With:

$$y^T = i_m, \varphi_2 : \begin{bmatrix} i_m \\ \varphi_2 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} i_m \\ \varphi_1 \\ \Delta\varphi \\ \varphi_2 \end{bmatrix} \quad (6)$$



The Lyapunov stability condition can be verified if the stabilization function  $\dot{V}(z_1)$  is chosen as follows:

$$\dot{V}(z_1) = -c_1 z_1^2 \leq 0 \forall z_1 \quad (11)$$

where  $c_1$  is the positive constant design that is determine the closed-loop dynamics  
Substitute (11) into (10) Result in:

$$i_{sd} = -c_1 T_r z_1 + i_m + T_r \frac{di_m^*}{dt} \quad (12)$$

with: 
$$\begin{cases} i_{sd}^* \approx i_{sd} \\ i_{sq}^* \approx i_{sq} \end{cases} \quad (13)$$

The (12) as shown:

$$i_{sd}^* = -c_1 T_r z_1 + i_m + T_r \frac{di_m^*}{dt} \quad (14)$$

so, the  $i_{sd}^*$  is the actual control signal, the control rule for the flux controller.

### 3.2. Speed controller design

For designing this nonlinear control law, it is assumed that all feedback signals from the motor, the shaft and the load sides are available in this section. As Figure 3 shows, the proposed backstepping control algorithm is designed in two steps. In each step a subsystem will be studied. Each step includes the dynamics of the previous subsystems for backstepping technique. Selecting a proper Lyapunov function for each step, guarantees the asymptotic stability of the related subsystem. Completing all two steps, results in not only the asymptotic stability of the whole system, but also the desired input (motor torque) law for accurate load speed control.

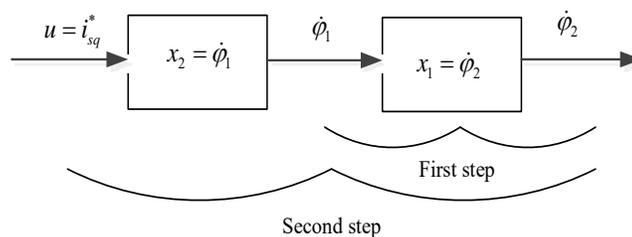


Figure 3. Block diagram of two-step backstepping control algorithm of a two-mass system

The speed control equation is given:

$$\begin{cases} \ddot{\phi}_1 = \frac{1}{J_1} k_m i_{sq} - \frac{c}{J_1} \Delta\varphi - \frac{d}{J_1} \dot{\phi}_1 + \frac{d}{J_1} \dot{\phi}_2 \\ \ddot{\phi}_2 = \frac{d}{J_2} \dot{\phi}_1 + \frac{c}{J_2} \Delta\varphi - \frac{d}{J_2} \dot{\phi}_2 - \frac{1}{J_2} m_L \end{cases} \quad (15)$$

The (15) has triangular form as follows:

$$\begin{cases} \frac{d(\dot{\varphi}_2)}{dt} = f(\dot{\varphi}_2, \dot{\varphi}_1) \\ \frac{d(\dot{\varphi}_1)}{dt} = f([\dot{\varphi}_2, \dot{\varphi}_1], i_{sq}) \end{cases} \quad (16)$$

these two step for designing the backstepping control law are mentioned in the following:

*Step 1: apply the backstepping for load speed  $\dot{\varphi}_2$ :*

Since load speed magnitude are our control variables, the tracking errors are defined by:

$$z_2 = \dot{\varphi}_{2d} - \dot{\varphi}_2 \quad (\dot{\varphi}_2 \text{ Is the reference load speed}) \quad (17)$$

taking the derivative of both sides of (17) Gives:

$$\dot{z}_2 = \ddot{\varphi}_{2d} - \ddot{\varphi}_2 \quad (18)$$

to stabilize the flux signal given by (18), the candidate Lyapunov fuction can be defined as following:

$$V(z_2) = \frac{1}{2} z_2^2 \geq 0 \quad (19)$$

which the derivative of (19) Is:

$$\dot{V}(z_2) = z_2 \dot{z}_2 \quad (20)$$

then:

$$-z_2 = -\dot{\varphi}_1 - \frac{c}{d} \Delta\varphi + \dot{\varphi}_2 + \frac{1}{d} m_L \quad (21)$$

the (21) Is result in:

$$\dot{\varphi}_1 = \dot{\varphi}_{2d} + \frac{1}{d} m_L - \frac{c}{d} \Delta\varphi \quad (22)$$

The Lyapunov stability condition can be verified if the stabilization function  $\dot{V}(z_2)$  is chosen as follows:

$$\dot{V}(z_2) = -\frac{d}{J_2} z_2^2 \leq 0 \forall z_2 \quad (23)$$

where  $d$  is the positive constant design that is determine the closed-loop dynamics.

From (18) and (15) We obtain:

$$-\frac{d}{J_2} z_2^2 = z_2 (\ddot{\varphi}_{2d} - \frac{d}{J_2} \dot{\varphi}_1 - \frac{c}{J_2} \Delta\varphi + \frac{d}{J_2} \dot{\varphi}_2 + \frac{1}{J_2} m_L) \quad (24)$$

then

$$-z_2 = -\dot{\varphi}_1 - \frac{c}{d} \Delta\varphi + \dot{\varphi}_2 + \frac{1}{d} m_L \quad (25)$$

so load speed shown as:

$$\dot{\varphi}_1 = \dot{\varphi}_{2d} + \frac{1}{d} m_L - \frac{c}{d} \Delta\varphi \quad (26)$$

Step 2: apply the backstepping for motor speed ( $\dot{\varphi}_1$ ):

Since load speed magnitude are our control variables, the tracking errors are defined by:

$$z_1 = \dot{\varphi}_{1d} - \dot{\varphi}_1 \quad (\dot{\varphi}_{1d} \text{ Is the reference motor speed}) \quad (27)$$

taking the derivative of both sides of (27) Gives:

$$\dot{z}_1 = \ddot{\varphi}_{1d} - \ddot{\varphi}_1 \quad (28)$$

To stabilize the flux signal given by (28), the candidate Lyapunov fuction can be defined as following:

$$V_1 = \frac{1}{2} z_2^2 + \frac{1}{2} z_1^2 \quad (29)$$

which the derivative of (29) is:

$$\dot{V}_1 = z_2 \dot{z}_2 + z_1 \dot{z}_1 \quad (30)$$

then substitute (15) and (28) into (30), we get

$$\begin{cases} \ddot{\varphi}_{1d} - \frac{1}{J_1} k_{\omega} i_m i_{sq} + \frac{c}{J_1} \Delta\varphi + \frac{d}{J_1} \dot{\varphi}_{1d} - \frac{d}{J_1} \dot{\varphi}_2 + \frac{d}{J_2} z_2 = 0 \\ \ddot{\varphi}_{2d} + \frac{d}{J_2} \dot{\varphi}_{1d} - \frac{c}{J_2} \Delta\varphi + \frac{d}{J_2} \dot{\varphi}_{2d} + \frac{1}{J_2} m_L = 0 \end{cases} \quad (31)$$

the (31) is result in:

$$i_{sq}^* = \frac{1}{k_{\omega} i_m} (c\Delta\varphi + d\dot{\varphi}_{1d} - d\dot{\varphi}_2 + \frac{dJ_1}{J_2} z_2) \quad (32)$$

so, the  $i_{sq}^*$  is the actual control signal, the control rule for the speed controller.

### 3.3. Design the reference trajectories

When designing the controller, we usually set the desired value to a constant (step-by-step trajectory). At the first interval of time, the large deviation will cause a large input requirement. However, the response of the actuator is finite, it is unlikely that the controller will meet the requirements. Thus the backstepping control method will "soften" the desired signal by changing the desired signal slowly so that the response of the system is soft and better. For this reason, the second order is chosen for designing the reference trajectories. Because the control rules are represented in (14) And (32) Use the first order derivative, the form of trajectories  $i_m^*$  and  $\omega^*$  will have to be first order differentiable.

Flux reference trajectory designing  $i_m^*$

$$\frac{i_m^*(s)}{i_m^d(s)} = \frac{1}{(1+T_1s)^2} = \frac{1}{1+2T_1s+T_1^2s^2} \quad (33)$$

where:  $T_1$  is time constant;  $i_m^*$  is output of flux reference trajectory;  $i_m^d$  is input of flux reference trajectory.

Therefore, on the time domain:

$$T_1^2 \frac{d^2 i_m^*}{dt^2} + 2T_1 \frac{di_m^*}{dt} + i_m^* = i_m^d \quad (34)$$

trajectories of flux reference:

$$\frac{d^2 i_m^*}{dt^2} = \frac{1}{T_1^2} \left( i_m^d - i_m^* - 2T_1 \frac{di_m^*}{dt} \right) \quad (35)$$

Speed reference trajectory designing  $\omega^*$

$$\frac{\omega^*(s)}{\omega^d(s)} = \frac{1}{(1+T_2s)^2} = \frac{1}{1+2T_2s+T_2^2s^2} \quad (36)$$

where:  $T_2$  is time constant;  $\omega^*$  is output of speed reference trajectory;  $\omega^d$  is input of speed reference trajectory. Therefore, on the time domain:

$$T_2^2 \frac{d^2 \omega^*}{dt^2} + 2T_2 \frac{d\omega^*}{dt} + \omega^* = \omega^d \quad (37)$$

trajectories of flux reference:

$$\frac{d^2 \omega^*}{dt^2} = \frac{1}{T_2^2} \left( \omega^d - \omega^* - 2T_2 \frac{d\omega^*}{dt} \right) \quad (38)$$

#### 4. RESULT AND ANALYSIS

In this section, the proposed control strategy is verified by simulation using Hardware-in-the-Loop (HIL). The proposed control is also verified in Hardware-in-the-Loop (HIL) environment using a Typhoon device in Figure 4. This device consists of an HIL402 card that simulates grid source, load, and three full-bridge with a common DC-link capacitor using IGBTs. The system hardware is simulated in real time on the HIL platform with a time step of 1  $\mu$ s that means very nearly physical model, while the pulse width modulation (PWM) carrier frequency is 5 kHz. Voltage and current controllers as well as PLL are implemented in DSP TMS320F2808 card.

The simulation parameters for rated power: 0.5kW; rated phase voltage: 220 VRMS; rated frequency: 50Hz;  $d=0.313$ Nm/rad/s;  $c=300000$ Nm/rad. Simulation procedure is as follows:

At  $t = 0$ s, the magnetic current is created. Then at time instance  $t = 0.4$ s, the motor starts to speed up to 2500rpm. At  $t=1$ s, the full load is inserted. At  $t = 1.4$ s, the motor is continued to speed up to 3500rpm with full load. Finally, the motor starts to change the rotating direction with a reference speed, i.e., 2500rpm, at time instance  $t = 2.0$ s.

The simulation results of the proposed method are presented and compared to those getting from the conventional PI controller to evaluate the advantage of the backstepping method.

The responses of magnetizing current are shown in Figure 5. Both conventional PI and backstepping methods show stable operation not only in nominal but also field weakening range. However, it can be observed that the backstepping gives better transient response, i.e., without overshoot as seen in Figure 5 and shorter settling time (0.2s in comparison with 0.35s for PI control).

The response of the torque, speed as well as torque ripple of the motor are shown in Figure 6. As can be observed, the torque and speed quickly match their commands for both methods. Nevertheless, the backstepping always shows better performance, i.e., lower torque ripple and shorter accelerating time, as can be seen in Figure 7.

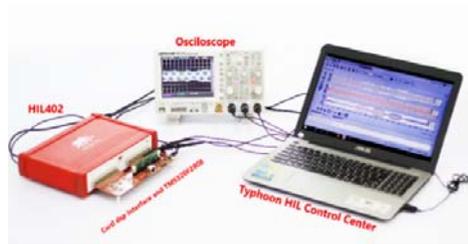


Figure 4. Picture of hardware HIL platform.



Figure 5. Magnetizing current response

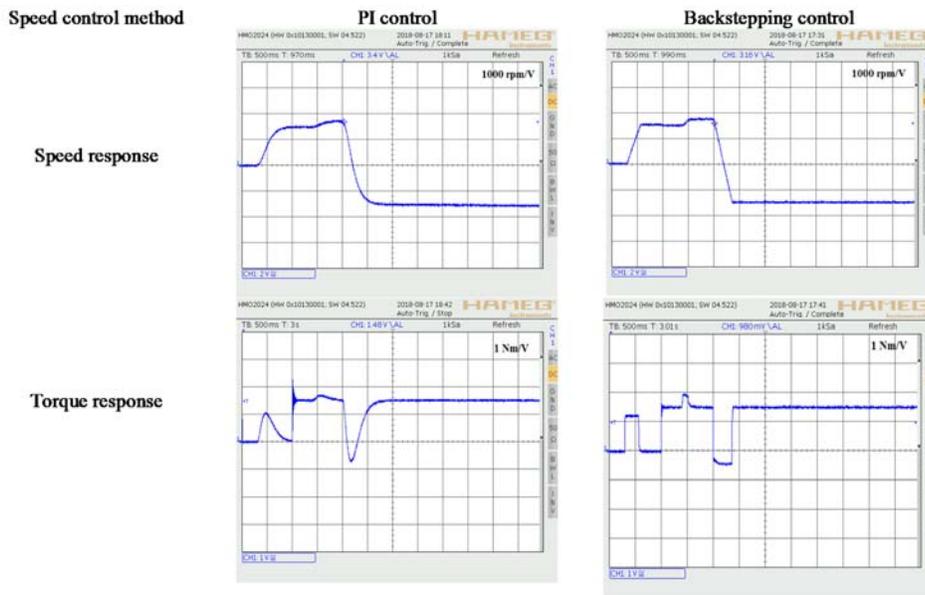


Figure 6. Speed and torque responses

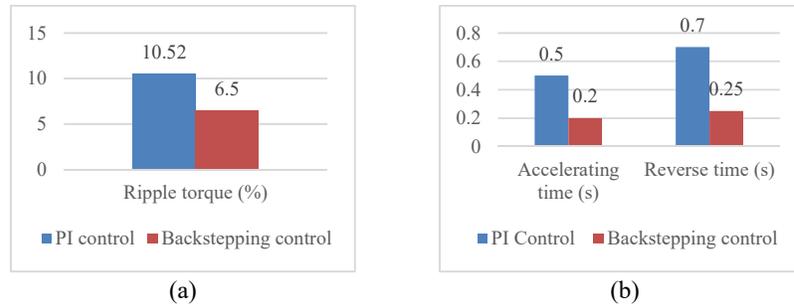


Figure 7. Chart of torque and speed responses, (a) torque responses, (b) speed responses

## 5. CONCLUSION

In this paper, the two-mass system with flexible couplings comprises of an induction motor and a load is considered. At the motor side, the current response is assumed to be ideal leading to a simplified model. Backstepping control is employed to solve flux and speed control problem of the system. The system can operate at field weakening region. The simulation results show that high dynamic and suppression of the mechanical oscillation of the drive system can be achieved.

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