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High Performance Feedback Linearization Control of Induction Motor

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***Abstract* – In induction motor control, power efficiency is an important factor to be considered. We attempt to achieve both high dynamic performance and maximum power. In this paper a feedback linearization controller is presented for Induction Motor drive. The controller is designed based on the adaptive input-output feedback linearization control technique. The effects of variations in mechanical parameter values on the linearized system are derived and discussed. The subject of this paper is to ascertain various controller's algorithms' advantages and disadvantages and give recommendations for it's use under certain conditions and in distinct applications. The drive will be studied as fairly as possible by discussing it's parameter sensitivities, dynamic responses, and steady-state errors. Computer simulation results obtained, confirm the effectiveness and validity of the proposed control approach.**

***Keywords* – Induction Motor, Feedback Linearization, Steady State, Flux Observer, Sensitivity.**

I. Introduction

The application of induction motors in traction systems, including electric vehicles, requires comparison of available drives for traction.Feedback Linearization control will be analyzed dynamically and in steady state, simulated, and implemented. Recommendations will be given for it's use in different applications on the basis of it's controller's mathematical underpinnings and findings. The control algorithm to be analyzed is feedback linearization. The simulation for each type of control algorithm will be conducted using Matlab-Simulink. FB-Linearization scheme is given in figure 1.

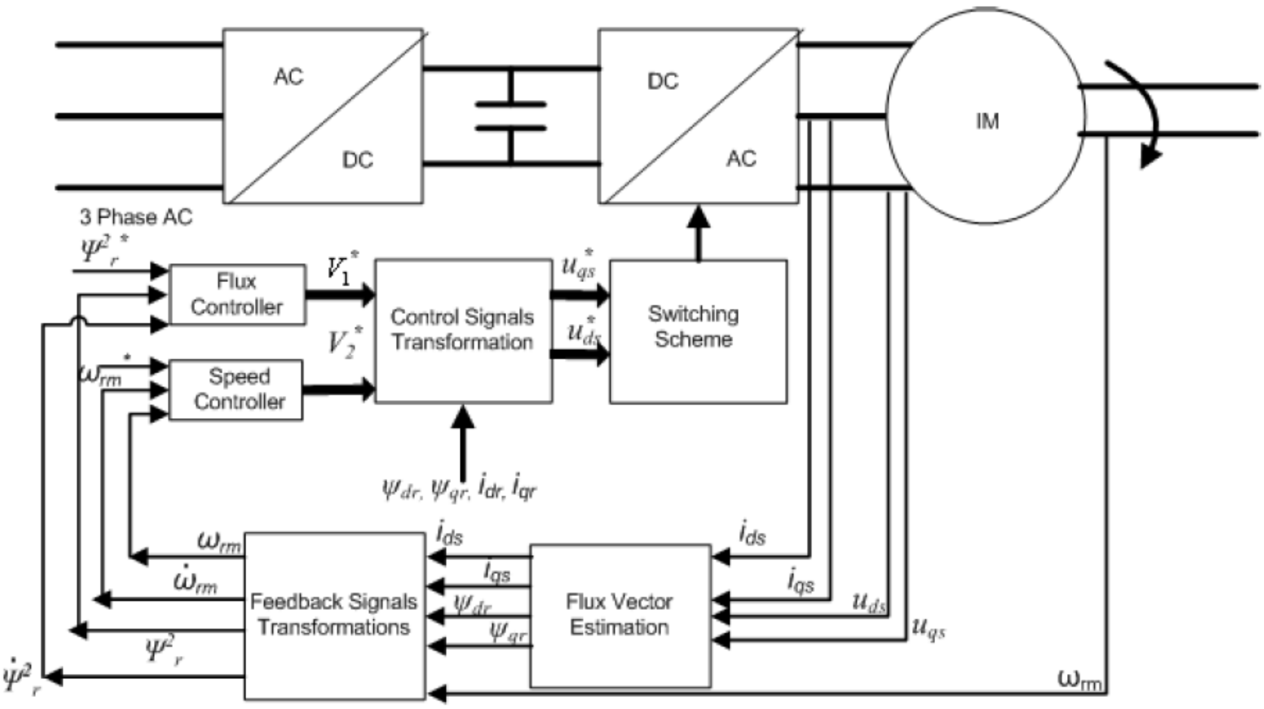

Fig.1. Block Diagram of Feedback Linearization

II. Literature Review

Direct comparisons of dynamic responses of IFOC and DTC do not appear to be available in the literature. Existing analyses of parameter sensitivities do not quantize the effect of parameter variations or errors on transient responses. Most of the literature deals with steady-state performance measures [3], [4], [5], [6], while [4], [5], [6], [7] provide some comparisons of dynamic responses. A detailed comparison of different induction motor drives is given in [3], including volts per hertz control (V/Hz), FOC, DTC, direct self control (DSC), and DTC with space vector modulation (DTC-SVM). This comparison mentions advantages and disadvantages relative to steady-state measures, such as phase current peaks, current and torque harmonics, and switching frequency variation. Structural measures, such as the need for flux observers, and decoupling of torque and flux commands are also presented. In [4], classical DTC and DTC-SVM, but not IFOC, are compared. The authors in [4] try to match the switching scheme with the drive in order to have similar switching frequencies, but DTC is used with a switching table tha results in variable switching frequencies, while DTC-SVM is used with fixed-frequency SVM. The speed and torque dynamic responses, including settling time and overshoot are compared. Tripathi et al. [8] propose a modified DTC which uses the stator flux to control the torque. No clear comparisons are made between DTC, DTC-SVM, and FOC, even though a vector diagram showing the dynamic operation of FOC and DTC is presented. Sikorski et al. [9] compare linear DTC-SVM to nonlinear DTC methods, DTC-δ, DTC-2x2, and DTFC-3A, using steady-state performance metrics while keeping the average switching frequency the same.

Cruz et al. [5] compare FOC, DTC and input-output linearization based on steady-state torque ripple, current peak, and switching frequency. They conclude that FOC and DTC are "good" in dynamic response, and that the parameter sensitivities are "low" and "medium" in DTC and FOC, respectively. Wolbank et al. [7] compare low and zero-speed applications of DTC and sensorless FOC. They study steadystate stability and speed overshoot, where FOC shows slower dynamics but better steady-state tracking compared to DTC. As both FOC and DTC have drawbacks, an interesting combination of DTC and FOC is presented in [10]. The resulting direct torque and stator flux control method (DTFC) uses no voltage modulation, current regulation loops, coordinate transformations, or voltage decoupling. Casadei et al. [11] evaluate standard DTC and DFOC and present a unique scheme, discrete space vector modulation (DSVM), which is a variation of the standard SVM. Performance criteria are steady-state current and torque ripples, and dynamic response due to a torque step.

Comparisons of other drives focus on steady-state response. Thomas et al. [12] propose and experimentally validate geometric sliding mode/limit cycle control. Three different inverter modes, asynchronous, synchronous, and square wave, are analyzed. Industrial control objectives such as stator and rotor flux regulation, torque, speed, and position control, minimal energy and harmonics criteria, and optimization of torque pulsations are evaluated. Refs. [13] and [14] discuss formal validation of DTC, from a singular perturbation and a geometric control perspective, respectively. In [13], the controls for DTC are established independent of the inverter, and it is shown that the DTC approach can be decoupled from the switching scheme.

The comparison of FOC and DTC (but not IFOC) provided in [6] is perhaps the most thorough in the literature. There, dynamic performance of both controls is compared and sensitivity analyses are done with respect to stator resistance for DTC and rotor-time constant for FOC. The main drawback is the lack of quantization of torque and flux dynamics, and parameter sensitivities. Vasudevan et al. [15] compare IFOC to DFOC, along with classical DTC-SVM and direct torque neuro-fuzzy control, using MATLAB/Simulink. Stator voltages and currents, angular velocity, torque, and flux responses to a change in torque or angular velocity, are compared. The effect of parameter variation such as stator resistance variation due to temperature increases is also discussed in relation to the DTC control method. Ref. [16] presents an interesting approach targeting the operation of DFOC, DTC with PWM and DTC-SVM under a driving cycle of an electric vehicle. However, the authors do not compare IFOC and DTC.

Vectorized volts-per-hertz is by far the simplest motor drive since it requires no parameter knowledge and is essentially an open-loop drive [22]. In [20], Krzeminski comes up with a type of nonlinear state feedback control which is completely input-output decoupled at all times even through transients. This differs from IFOC because IFOC is decoupled only when flux and speed are in steady state. FL-IODC achieves better performance than IFOC in theory due to accounting for the stator resistive drop and other terms that allow it to have complete decoupling. The drawback of this type of control is the additional parameter sensitivity that results from the addition of these extra terms.

III. Induction Motor Dynamic Model

The following set of seven time-varying equations in (1) from [1] are known collectively as the induction motor dynamic model in the arbitrary reference frame. The variables with the subscript s indicate that it is a stator variable, while the subscript r shows it is a rotor quantity. It should be noted that this model is exclusively for variables in the dq0 plane rather than the time-variant plane. The motor terminal voltages are indicated by v, the currents by i, the flux linkages by λ, and the rotor speed by w_r. The number of pole pairs is given by n_P, the load torque by T_{load}, and the shaft moment of inertia by J. These equations will be used in the simulations that follow and are also inherent in governing the induction motor dynamic behavior in the implementation.

$$\begin{aligned}
\frac{d\lambda_{qs}}{dt} &= -r_s i_{qs} - w\lambda_{ds} + V_{qs} \\
\frac{d\lambda_{ds}}{dt} &= -r_s i_{ds} + w\lambda_{qs} + V_{ds} \\
\frac{d\lambda_{0s}}{dt} &= -r_s i_{0s} + V_{0s} \\
\frac{d\lambda_{qr}}{dt} &= -r_r i_{qr} - (w - n_p w_r)\lambda_{dr} + V_{qr} \\
\frac{d\lambda_{dr}}{dt} &= -r_r i_{dr} + (w - n_p w_r)\lambda_{qr} + V_{dr} \\
\frac{d\lambda_{0r}}{dt} &= -r_r i_{0r} + V_{0r} \\
\frac{dw_r}{dt} &= \frac{3}{2}\frac{n_p}{J}\left(\lambda_{ds} i_{qs} - \lambda_{qs} i_{ds}\right) - \frac{T_{load}}{J}
\end{aligned} \tag{1}$$

The load is modeled as quadratic in speed,

$$T_L = 1.82 \times 10^{-4} w_{rm}^2 + 1.82 \times 10^{-2} w_{rm} \tag{2}$$

IV. Feedback Linearization: Input-Output Decoupling Control

In [20], Krzeminski comes up with a type of nonlinear state feedback control which is completely input-output decoupled at all times even through transients. This differs from IFOC because IFOC is decoupled only when flux

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and speed are in steady state. FL-IODC achieves better performance than IFOC in theory due to accounting for the stator resistive drop and other terms that allow it to have complete decoupling. The drawback of this type of control is the additional parameter sensitivity that results from the addition of these extra terms. The inputs in this control scheme are v_{qs} and v_{ds}, while the outputs are the rotor mechanical speed w_r, and the flux magnitude squared ψ^2, Krzeminski assumes in the paper that the load torque response is known. If this is the case, then the new state space is:

$$y_1 = w_r$$
$$y_2 = \frac{dw_r}{dt}$$
$$y_3 = \psi^2 = \lambda_{qr}^2 + \lambda_{dr}^2 \quad (3)$$
$$y_4 = \frac{dy_3}{dt} = \frac{d\psi^2}{dt} = \frac{d(\lambda_{qr}^2 + \lambda_{dr}^2)}{dt}$$
$$y_5 = tan^{-1}\left(\frac{\lambda_{qr}}{\lambda_{dr}}\right) = \rho$$

From (3) and the original dynamic motor model, we get the following dynamic system, as seen in [21]:

$$\begin{aligned}
\frac{dy_1}{\mathrm{dt}} &= \mathrm{y}_2 \\
\frac{dy_2}{dt} &= f_{21}(y_1 \ldots . y_5) + f_{22}(y_1 \ldots . y_5)v_{ds} + f_{23}(y_1 \ldots . y_5)v_{qs} \\
\frac{dy_3}{dt} &= y_4 \\
\frac{dy_4}{dt} &= f_{41}(y_1 \ldots . y_5) + f_{42}(y_1 \ldots . y_5)v_{ds} + f_{43}(y_1 \ldots . y_5)v_{qs} \\
\frac{dy_4}{dt} &= f_5
\end{aligned} \quad (4)$$

If one sets v_{qs} and v_{ds} to the vector

$$\begin{pmatrix} v_{ds} \\ v_{qs} \end{pmatrix} = \begin{pmatrix} f_{22}(y_1 \ldots y_5) & f_{23}(y_1 \ldots y_5) \\ f_{42}(y_1 \ldots y_5) & f_{43}(y_1 \ldots y_5) \end{pmatrix}^{-1} \times \begin{pmatrix} -f_{21}(y_1 \ldots y_5) + u_{speed} \\ f_{41}(y_1 \ldots y_5) + u_{flux} \end{pmatrix} \quad (5)$$

The system now looks like (6)

$$\begin{aligned}
\frac{dy_1}{\mathrm{dt}} &= \mathrm{y}_2 \\
\frac{dy_2}{dt} &= u_{speed} \\
\frac{dy_3}{\mathrm{dt}} &= \mathrm{y}_4 \\
\frac{dy_4}{dt} &= u_{flux} \\
\frac{dy_5}{dt} &= n_p y_1 + \frac{2r_r}{3n_p y_3}(Jy_2 + T_{load})
\end{aligned} \quad (6)$$

The following inputs in (7) can be set to completely decouple the inputs from the outputs when using constant design parameters k_1, k_2, k_3 and k_4. Thus, if there is a transient in the flux magnitude squared or the speed of the rotor, the transient will not affect the other variable [23].

$$u_{speed} = -k_1(w - w_{ref}) - k_2\left(\frac{T_e}{J} - \frac{T_{load}}{J} - \dot{w}_{ref}(t)\right) + \ddot{w}_{ref}(t) \quad (7)$$
$$u_{flux} = k_3(\psi^2 - \psi_{ref}^2) - k_4 \left(2\frac{R_r}{L_r}\left(M(\psi_d i_d + \psi_q i_q) - (\psi_d^2 + \psi_q^2)\right) - \dot{\psi}_{ref}^2\right) + \ddot{\psi}_{ref}^2$$

VII. Feedback Linearization: Input-Output Decoupling Control Stability Analysis

In this part, global stability will be proved for feedback linearization input-output decoupled control.As in the other motor control schemes, it is still required that the Lyapunov function (LF) be positive definite and its derivative negative definite. In this case, the candidate LF is given in (8) from [24]. This particular LF is positive definite by observation.

$$V(e, \hat{T}_L) = \frac{1}{2}e^2 + \frac{1}{2\gamma}\hat{T}_L^2$$

Where

$$e = (w - w_{ref}) - (w_M - w_{M,ref}) \quad (8)$$

Where $\gamma > 0 \in R$.

and $\hat{T}_L$ is the observed load torque. The derivative of the above candidate LF is given by (9).

$$\dot{V}(e, \hat{T}_L) = -ke^2 + \hat{T}_L\left(-\frac{e}{J} - \frac{1}{\gamma}\frac{d}{dt}\hat{T}_L\right) \quad (9)$$

Where $k > 0 \, \epsilon \, R$.

Then from the torque adaptation law given in (10) from [24], the last two terms cancel out and (11) is the resulting derivative. Since k is positive, the derivative is always negative and therefore negative definite. Since the LF is positive definite and its derivative is negative definite, this means that the candidate LF is indeed an LF. This shows that the error between the speed error and the model's speed error goes to zero as time goes to infinity. The same goes for the observed load torque. This shows that the controller is stable since the load torque is always a bounded real number.

$$\frac{d}{dt}\hat{T}_L = -\gamma\frac{e}{J} \quad (10)$$
$$\dot{V}(e, \hat{T}_L) = -ke^2 \quad (11)$$

VIII. Flux Observer Design

There are numerous examples of observers for flux estimation in the literature. One particular paper, [25], stands out when looking for a survey of observers for induction machine control. Verghese and Sanders discuss existing methods for flux observers and come up with some innovative insights including an observer that uses predictive flux error to improve the observer. The resulting observer looks like that in (12). Although this observer is shown in [25] to converge very quickly, it requires more computation than a simpler but less accurate observer (13). The observer in (13) was used in all of the simulations because of its simplicity and prevalence. It comes directly from the machine equations in (1).

$$\hat{\lambda}_r' = \left[\left(-\frac{1}{T_r}\right)I + wJ\right]\hat{\lambda}_r + \left(\frac{1}{T_r}\right)Mi_s + K(\hat{v}_s - v_s) \quad (12)$$

where K is a 2×2 matrix of observer gains.

$$\frac{d\hat{\lambda}_{qs}}{dt} = v_{qs} - R_s i_{qs} - w\hat{\lambda}_{ds} \quad (13)$$

$$\frac{d\hat{\lambda}_{ds}}{dt} = v_{ds} - R_s i_{ds} - w\hat{\lambda}_{qs}$$

IX. Simulation of Feedback Linearization

Feedback linearization uses the feedback of many variables for its control algorithm. The variables that are fed back are: stator currents, stator voltages, and the speed, as shown in Figure 1. From these, the rotor flux is estimated. With the rotor flux and the stator currents, the states $\dot{\Psi}_r^2$ and $\dot{w}_r^2$ along with their integrals Ψ_r^2 and w_r^2 are sent to flux and speed controllers. From the flux and speed controllers, the desired voltages are sent to the switching scheme of choice. As before, the switching scheme creates the gate pulses that will control the induction motor.

X. Dynamic Response

Now that all parts of the drives' scheme have been individually analyzed, this part will conclude with test. The test that was chosen is torque step of the motor drive, which would be useful for determining which drives have a high torque response. Sticking with the automotive theme, one application could be the sheer acceleration of a HEV or EV from standstill, or when coming onto the highway. The Simulation is run for 2s with torque commands of 4 and 3 N.m, changing at 1.5s. fixed stator and rotor flux command of 0.52 V.s is used for a 4% leakage inductance. By going through this test, one can arrive at a possible conclusion as to does drive is better in a range of applications. The results of this test can be seen in Figures 2 and 3.

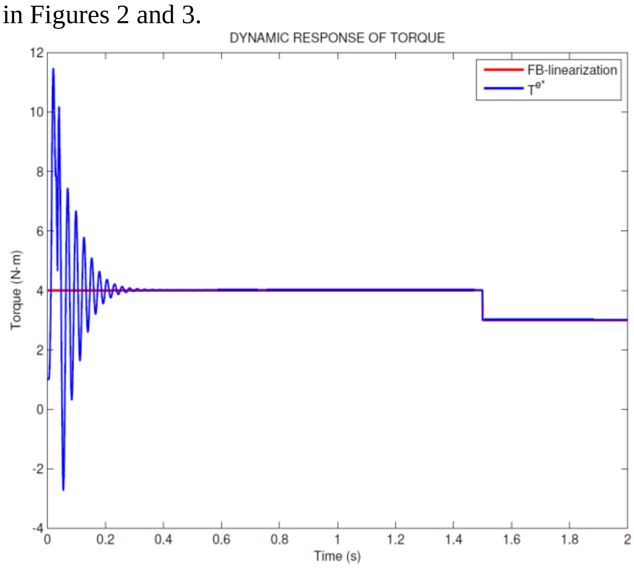

Fig.2. Torque Response of Feedback Linearization controller to a Step Torque Commandof 4 N-m to 3 N-m

The feedback linearization input-output decoupled control drive has an almost instantaneous response in the millisecond range. The plot in Figure 3 shows the start-up and step response for the corresponding speed response of the drive. Using standard motor drive and switching topology, FBlinearization clearly is the top performer.

XI. Sensitivity Analysis

A. Overview

Now that the block diagram for the simulation has been discussed, the parameter sensitivity results will be revealed. FL-IOL is the most sensitive to parameter change. FL-IOL is sensitive to errors in rotor and stator resistances, magnetizing inductance, and rotor and stator self inductances. This is demonstrated in the Jacobian shown in (14). The reason for this is that in order to totally decouple the input and output, all parameters must be known accurately; if not, the scheme does not work correctly.

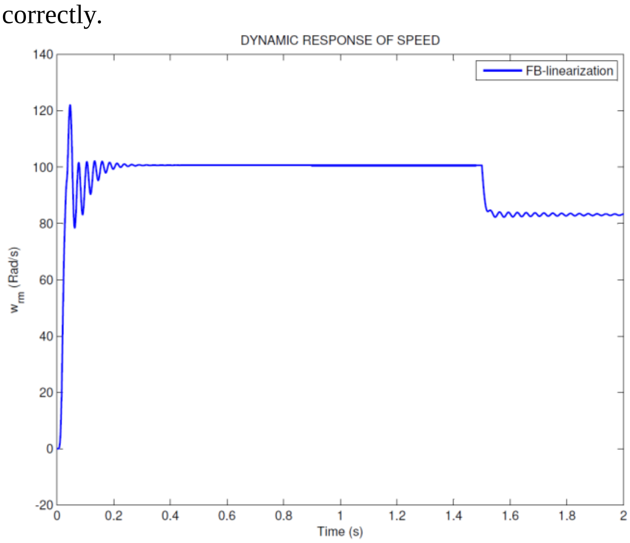

Fig.3. Speed Response of Feedback Linearization controller to a Step Torque Commandof 4 N-m to 3 N-m

$$\begin{pmatrix} \Delta T^e \\ \Delta w_{rm} \end{pmatrix} = J_{FL-IOL} \begin{pmatrix} \Delta r_r \\ \Delta r_s \\ \Delta L_m \\ \Delta L_r \\ \Delta L_s \end{pmatrix} \quad (14)$$

Where

$$J_{FL-IOL} = \begin{pmatrix} \frac{\partial T^e}{\partial r_r} & \frac{\partial T^e}{\partial r_s} & \frac{\partial T^e}{\partial L_m} & \frac{\partial T^e}{\partial L_r} & \frac{\partial T^e}{\partial L_s} \\ \frac{\partial w_{rm}}{\partial r_r} & \frac{\partial w_{rm}}{\partial r_s} & \frac{\partial w_{rm}}{\partial L_m} & \frac{\partial w_{rm}}{\partial L_r} & \frac{\partial w_{rm}}{\partial L_s} \end{pmatrix}$$

B. Feedback Linearization sensitivity results

The parameters that when change influence the feedback linearization motor drive are the rotor self inductance, magnetizing inductance, stator self inductance, the rotor resistance and stator resistance. The parameter sensitivity result for the rotor self inductance is shown in Figure 4. it shows the torque response for changes in the parameter and resulting speed plot. One can see that the results of deviations were ploted on each others, so the rotor self inductance has a very low sensitivity to deviations.

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Figure 5 shows the results for when the magnetizing inductance is altered by +/- 25%. Having a positive error of 25% causes a 12% increase of peak torque in addition to a slight increase in time required to get to steady-state. The opposite is the case for when the magnetizing inductance is lower thanpredicted by 25%; the peak torque is lowered by 11% and the steady-state time is reduced by 0.04 seconds, or 6.1%, from nominal conditions.

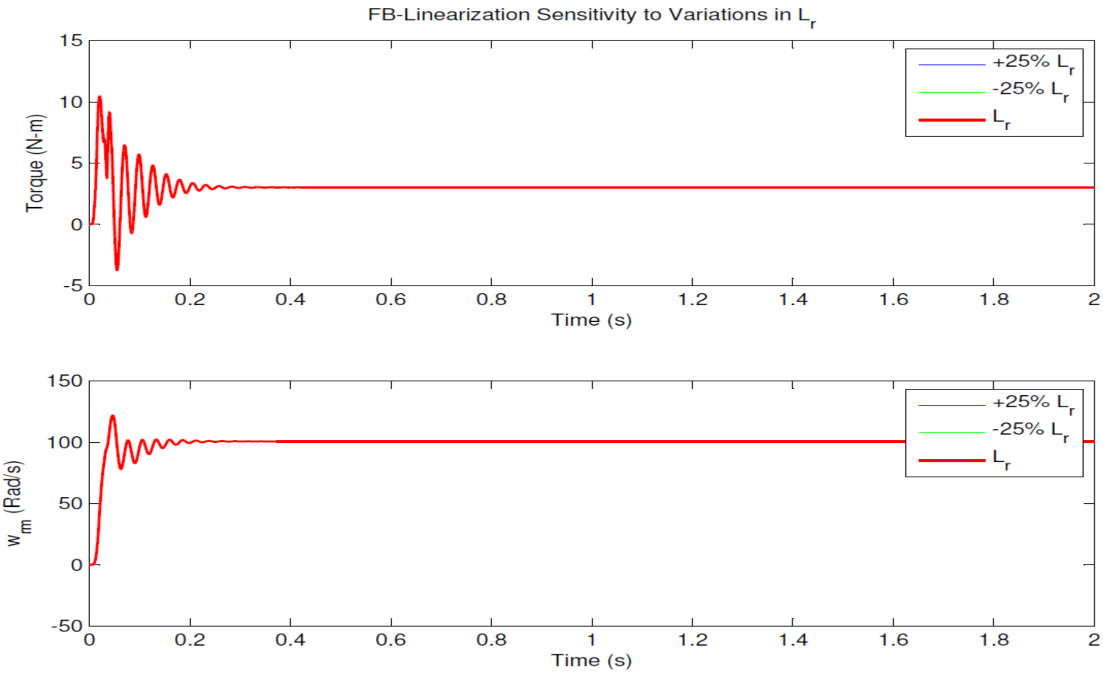

Fig.4. Feedback Linearization sensitivity to Variations in $\boldsymbol{L_r}$

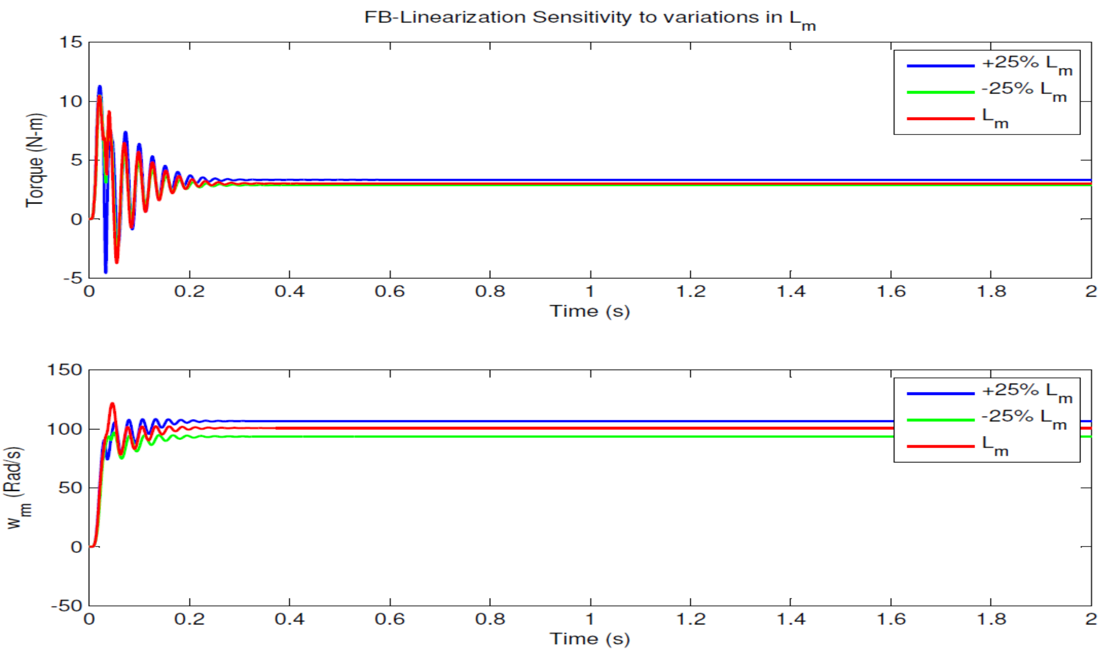

Fig.5. Feedback Linearization sensitivity to Variations in $\boldsymbol{L_m}$

Altering the stator resistance by 25% in a positive fashion causes the torque peak to be reduced by 9.1% and enter into steady-state 8.8% quicker than the nominal condition, as shown in Figure 6. Again, the opposite is true, a 25% decrease in expected rotor inductance results in a poorer performance from this motor controller. This causes the peak torque to increase by 14.5%, and the time to steady-state to increase by 21.4%. Therefore, if one was estimating the stator resistance.

The last parameter that needs to be examined for feedback linearization is the rotor resistance. This parameter is very sensitive to changes, as shown in Figure 7 for torque response and speed response. Altering the rotor resistance causes exactly the same response as changing the magnetizing inductance, because both parameters are contained in the feedback linearization flux observer and are in the same signal chain. This means that a 1% error in the same direction for either parameter's

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value causes the exact same erroneous response. Therefore, a positive error in the knowledge of the rotor resistance causes an undesired effect, while a negative error yields an increase in performance, as in the parameter sensitivity of the magnetizing inductance case above.

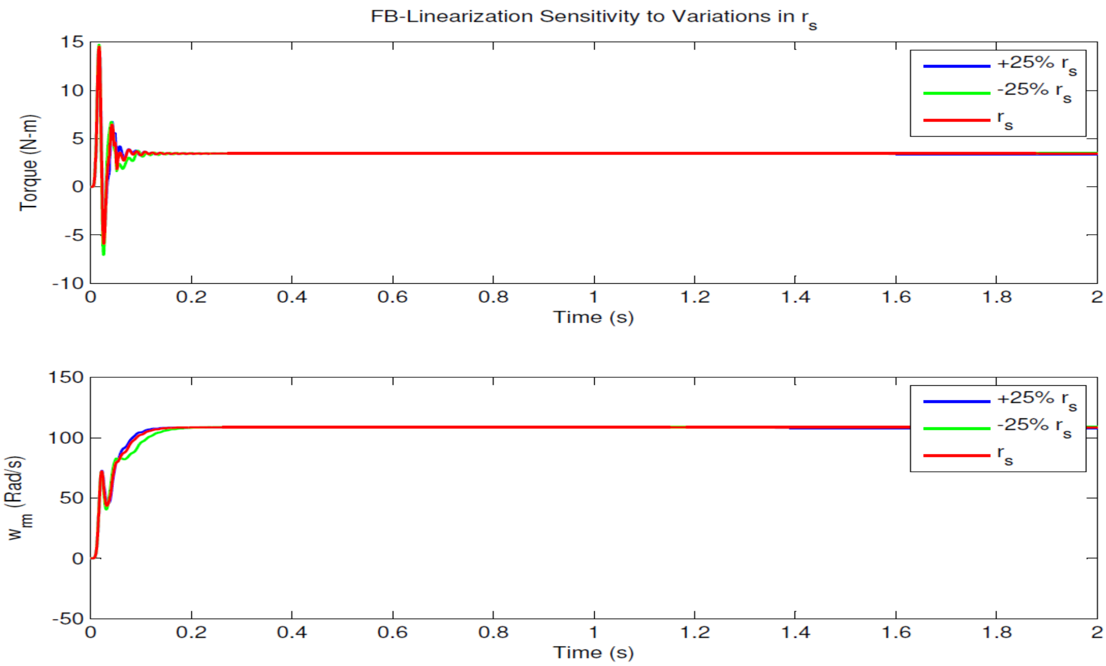

Fig.6. Feedback Linearization sensitivity to Variations in r_s

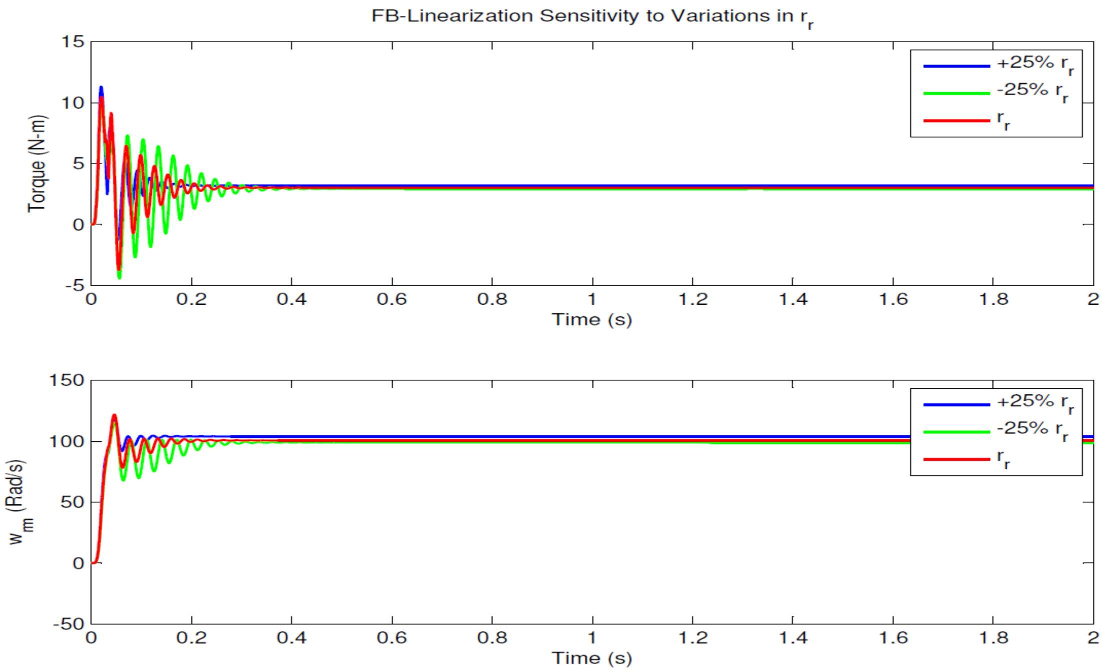

Fig.7. Feedback Linearization sensitivity to Variations in $\boldsymbol{r_r}$

XII. CONCLUSION

The paper models and analyzes motor controller, feedback linearization using a second order motor load, typical of many loads like a fan or industrial pump. All of the analysis has been done using this motor load and cannot be directly extrapolated to different applications, such as constant power loads, higher order loads, or loads that have inverse speed relationships. It was found and exhibited that the drive is advantageous in its own way. It was found that FB-linearization has superior torque step performance. The test, a torque step response, showed that FB linearization has desirable traction and performs well. Overall, it seems that FB linearization, given the right conditions and gains, performs the most admirably due to feedback signals. In the future, an investigation into which switching scheme is optimal with motor controller could be carried out.

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APPENDIX

Table 1 shows the induction motor parameters used in simulation.

Table 1: induction motor parameters

Induction motor Data		
Symbols	Description	Values
R_s	Stator Resistance	1.5293 Ω
R_r	Rotor Resistance	0.7309 Ω
L_s	Stator Inductance	0.20135 H
L_r	Rotor Inductance	0.20315 H
L_{ls}	Stator Leakage Inductance	0.00356 H
L_{lr}	Rotor Leakage Inductance	0.00535H
L_m	Magnetizing Inductance	0.19779 H
P	Number of poles	4
P_m	Power of Induction Motor	3 hp

NOMENCLATURE

P: Number of poles
r_s: Stator resistance (Ω)
r_r: Rotor resistance (Ω)
L_m: Mutual inductance (H)
L_s : Stator inductance (H)
L_r: Rotor inductance (H)
i_{qs}: Quadrature stator current (A)
i_{ds}: Direct stator current (A)
i_{abcs} : 3-phase stator current (A)
v_{qs}: Quadrature stator voltage (V)
v_{ds}: Direct stator voltage (V)
v_{abcs} : 3-phase stator voltage (V)
T_e: Electromechanical torque (N·m)
T_L: Load torque (N·m)
θ_e: Electrical angle (rad)
w_e: Electrical frequency (rad/s)
w_r: Rotor speed (rad/s)
w_{rm} : Mechanical speed (rad/s)
w_{sl}: Slip frequency (rad/s)
λ_{qs}: Quadrature stator flux linkage (V·s)
λ_{ds}: Direct stator flux linkage (V·s)
λ_{qr}: Quadrature rotor flux linkage (V·s)
λ_{dr}: Direct rotor flux linkage (V·s)
λ_s : Stator flux magnitude (V·s)
λr: Rotor flux magnitude (V·s)

The superscript "*" denotes a command input. The superscripts "*e*" and "*s*" denote variables in the synchronous and stationary reference frames, respectively.

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