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Direct Torque Control of Induction Motor Drive By using Fuzzy Logic Controller and Feedback Linearization Technique

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ABSTRACT

This paper presents a Direct Torque Controlled (DTC) Induction Motor (IM) drive that utilizes feedback linearization and Sliding-Mode Control (SMC). Another feedback linearization approach is proposed which is Fuzzy logic controller (FLC), which outputs a decoupled direct IM display with two state factors: torque and stator flux. This inherent linear model is utilized to actualize a DTC sort controller that preserves all DTC favorable circumstances and takes out its primary disadvantage, the flux and torque swell. Robust, quick, and swell free control is accomplished by utilizing FLC with corresponding control in the region of the sliding mode. Fuzzy logic controller guarantees robustness as in DTC, while the corresponding segment robustness out the torque and flux swell. The torque time reaction is like traditional DTC and the proposed arrangement is adaptable, profoundly tunable because of the P component. The controller design is displayed and its robustness solidness is analyzed in simulations. The FLC controller is contrasted and a direct DTC scheme with and without Feedback linearization. By using FLC controller extensive investigative comes about for dynamic response of a sensorless IM drive approve the proposed solution.

Keywords : Direct Torque Control, Adjustable Speed Drives, Feedback Linearization, Induction Motor Drives, Sliding Mode Control, Fuzzy Logic Controller

I. INTRODUCTION

Direct Torque Control (DTC) is a robust, quick responding control technique for Induction Machine (IM) drives [1]. Customary DTC utilizes shut circle hysteresis torque also, flux controllers and a changing table to choose the voltage vector connected to the motor. DTC accomplishes quick and robust torque and flux control without utilizing current controllers. DTC operation is related with expansive torque swell which causes ripple, vibrations, and expanded misfortunes, while the switching frequency of the Voltage Source Inverter (VSI) is variable and low. Enhanced DTC arrangements that keep running at consistent switching frequency and utilize present day control hypothesis have as of late been created to reduce the torque ripple. Novel DTC procedures in light of discrete Space Vector Modulation (SVM) techniques are portrayed in. DTC in light of linear torque and flux controllers (Linear DTC) and SVM was presented in [2]. A few designs utilizing the variable structure control standards have been proposed in [3].

Feedback Linearization (FBL) is a nonlinear control approach. The fundamental thought of FBL is to change a nonlinear system into a proportionate direct system, design a linear controller for the direct system, and afterward utilize the opposite change to acquire the desired controller for the first nonlinear system. Since the technique is touchy to displaying errors and disturbances, it has been once in a while connected to IM drives. FBL is utilized as a part of [4]-[5] to linearize the IM display with regard to speed, flux, and current. Two linearization designs in which just a single control amount is changed are discussed about in [5]. All arrangements in [4]-[5] depend on current linearization and control. Utilizations of FBL to control devices and PMSM drives are displayed. An error affectability investigation in demonstrates that the control performance may fall apart because of perturbations, parameter detuning, and estimation errors.

Sliding Mode Control (SMC) is a robust control strategy appropriate for control systems with uncertainties or demonstrating errors [6]. It has been effectively connected to IM drives and gives superb dynamic performance to a wide speed extend operation [3]. The switching conduct can be directed with the VSI operation as appeared in [6]. Fact be told, the customary DTC is a type of SMC which was composed to nearly direct the switching idea of the VSI.

This paper proposes another DTC controller that incorporates Feedback linearization together with fuzzy logic controller (FLC). The principle preferred advantage of FBL over traditional DTC is that the linear control hypothesis results can without much of a extend be connected to acquire a superior performance. We utilize this property to design and after that hypothetically explore the robustness and dependability of the proposed control strategy. Besides, the controller-spectator division rule enables the controller and the spectator to be autonomously designed, if the design display is around linear and estimation errors are small. The FBL load is the affectability of the linearized model to uncertainties and parameter detuning, which motivates the utilization of FLC.

The nonlinear IM demonstrate considered in this paper is fourth arrange with the state factors: torque, stator flux, rotor flux, described more, another flux subordinate state. The feedback linearized IM show is second order, with just the torque and stator flux magnitude as decoupled state factors. In this way, the new direct display is natural, extremely linear forward, and it generously rearranges the controller design. The flux and torque are controlled by the new DTC conspire and the proposed controllers utilize FLC to keep up robustness sensorless operation of the drive. This approach in light of torque-flux linearization and control is not the same as existing techniques in [4]-[5], which are based on current control. The mix of these systems preserves the quick and robustness reaction of traditional DTC while altogether taking out the torque and flux ripple.

II. FEEDBACK LINEARIZATION OF IMMODEL

Regular linearization of a nonlinear system depends on a first-arrange estimation of the system dynamic at a chosen working point while dismissing high-arrange flow. This linearization is satisfactory in numerous applications where typical system operation stays in the region of a settled or gradually differing balance, however it is generally wrong. In specific, linearization is proper for IM drives working at consistent rotor speed. Something else, the IM conduct is intrinsically nonlinear and different methodologies must be utilized.

Feedback linearization is a method that permits the designer to utilize linear control methodologies with naturally nonlinear systems, for example, the IM. The FBL logarithmically changes a nonlinear system display into a direct one, so that direct control systems can be utilized. Dissimilar to regular linearization, the linearization and the direct conduct are substantial comprehensively, as opposed to in the region of a harmony point. When all is said in done, the linearizing change is very hard to discover, however at times it is anything but difficult to acquire by a basic redefinition of factors [7]. Fortunately the FBL of an IM is achievable by a natural change of the state factors and an info redefinition.

The IM state space show in the stator reference design is

$$\frac{d\psi_s}{dt} = -\frac{1}{T_s\sigma}\psi_s + \frac{L_m}{L_r T_s \sigma}\psi_r + u_s \qquad (1)$$

$$\frac{d\psi_r}{dt} = \frac{L_m}{L_s T_r \sigma}\psi_s - (\frac{1}{T_s\sigma} - j\omega_r)\,\psi_r \qquad (2)$$

where □s, □r are stator and rotor flux space vectors, Rs and Rr are the stator and rotor resistances, Ls, Lr and Lm are the stator, rotor and magnetizing inductances, $Ts = Ls/Rs$, $Tr = Lr/Rr$, $\sigma = (LsLr - Lm2)/LsLr$, □r is the rotor speed, and $us = usd + jusq$ is the stator voltage vector which acts as Feedback.

The model can be linearized by selecting the new states:

$$M = \psi sq\psi rd - \psi sd\psi rq \qquad (3)$$

$$R = \psi sd\psi rd + \psi sq\psi rq \qquad (4)$$

$$Fs = \psi sd2 + \psi sq2 \qquad (5)$$

$$Fr = \psi rd2 + \psi rq \qquad (6)$$

Where M is the scaled torque, Fs and Fr are the squared extents of the stator and rotor flux, individually. The variable R relies upon the rotor and stator flux. For linear forwardness, we refer M as the torque and Fs as the flux size. We are essentially intense on controlling the torque M and the stator flux flux Fs. In any case, we should likewise safeguard that the remaining state factors, Fr and R, are limited.

The IM state conditions with the state factors (3) - (6) are

$$\frac{dM}{dt} = -(\frac{1}{Tr\sigma}+\frac{1}{T_s\sigma})\,M - \omega rR - \psi rqusd + \psi rdusq \quad (7)$$

$$\frac{dFs}{dt} = -\frac{2}{Ts\sigma}Fs + \frac{2Lm}{LrTs\sigma}R + 2\psi sdusd + 2\psi sqUsq \qquad (8)$$

$$\frac{dFr}{dt} = -\frac{2}{T_r\sigma}F_r + \frac{2L_m}{L_s T_r \sigma}R \qquad (9)$$

$$\frac{dR}{dt} = -(\frac{1}{Tr\sigma}+\frac{1}{T_s\sigma})\,R + \omega rM + \frac{Lm}{LrTs\sigma}Fs + \frac{Lm}{LrTs\sigma}Fr + \psi rdusd + \psi rqusq \qquad (10)$$

The first three state equations are feedback linearized if the Feedbacks redefined as

$$wq = -\omega rR - \psi rq\, usd + \psi rdusq \qquad (11)$$

$$wd = \frac{2Lm}{LrTs\sigma}R + 2(\psi sdusd + \psi sq\, u) \qquad (12)$$

Now the linearized system is

$$\frac{dM}{dt} = -(\frac{1}{Tr\sigma}+\frac{1}{T_s\sigma})\,M + wq \qquad (13)$$

$$\frac{dFs}{dt} = -\frac{2}{Ts\sigma}Fs + wd \qquad (14)$$

$$\frac{dFr}{dt} = -\frac{2}{T_r\sigma}F_r + \frac{2L_m}{L_s T_r \sigma}R \qquad (15)$$

$$\frac{dR}{dt} = -(\frac{1}{Tr\sigma}+\frac{1}{T_s\sigma})\,R + \frac{Lm}{LrTs\sigma}Fs + \frac{F_r}{2R}wd - \frac{M}{R}wq \qquad (16)$$

Solving (11) and (12) gives the control signals

$$u_{sd} = \frac{\psi_{rd}}{2R}\left(w_d - \frac{2Lm}{LrTs\sigma}\right) - \frac{\psi_{sq}}{R}\left(\omega_q + \omega_r R\right) \qquad (17)$$

$$u_{sq} = \frac{\psi_{rq}}{2R}\left(w_d - \frac{2Lm}{LrTs\sigma}\right) - \frac{\psi_{sd}}{R}\left(\omega_q + \omega_r R\right) \qquad (18)$$

FBL decouples the state factors of interest; specifically, the torque M and the stator flux Fs and in this manner fundamentally improves the controller design for the IM drive system. In expansion, since the subsequent system is linear, the traditional linear control approaches can be utilized. Since the M, Fs and Fr have elements with left design posts, the info output dependability of the rest of the state factors can be effortlessly ensured gave that R remains limited. The R state condition (16) demonstrates that its correct hand side is unbounded for zero R, which just happens in the unimportant condition when the stator or rotor flux is zero. With the exception of the startup, this condition never happens during normal operation. In the physical drive, the controller guarantees that the flux has been introduced before beginning the drive. Activity brings about Segment IV demonstrate that the torque control is begun with a 40 ms delay after the flux control, when flux are at apparent levels. It is in this manner accepted that the variable R has a lower bound, R_l. R is likewise upper limited practically speaking on the grounds that the flux extents are restricted because of attractive immersion.

III. DIRECT TORQUE CONTROL VIA SLIDING MODE

Sliding Mode Control (SMC) is utilized to accomplish a quick and robust operation of an IM drive. Fig. 1 demonstrates the block diagram of the proposed drive. The block Controllers and SVM contains the FBL and the torque and flux controllers portrayed next. The drive utilizes linear forward speed, torque, and flux observer, described more, a PI speed controller. Drive information and a concise representation of the spectators are given in the Appendix.

The control objective is to control the torque and stator flux level in the machine, i.e. to understand a DTC sort controller. To this end, we design controllers for the torque M described more, the stator flux Fs in the linearized demonstrate. Since the state conditions (13) and (14) representing M and Fs individually are decoupled, the design of their controllers to acquire the data sources wd and wq is very linear forward. These are then substituted in (17) described more, (18) to acquire the physical sources of info usd and usq separately. Be that as it may, errors in the computation of the physical sources of info are unavoidable and must be represented and corrected to give controlling performance.

The errors in the physical control sources of info can be represent to as proportionate errors in the direct state conditions (13) and (14).

Condition (13) can be reworked in the frame

$$\frac{dM}{dt} = G_m + w_q \qquad (19)$$

Where gM represents the uncertain dynamics of the FBL torque equation. The term gM is not exactly known; from (13) an estimate of the dynamics

Is $\hat{g}M = -\left(\frac{1}{Tr\sigma} + \frac{1}{T_s\sigma}\right) M$

We assume that the estimation error for gM is bounded as

$$|\hat{g}M - gM| \leq GM \qquad (20)$$

To design the SMC for the linear system of (19), we define the sliding mode as the torque error

$$SM = M - Md \qquad (21)$$

For this choice of sliding mode, we use the SMC

$$wq = -\hat{g}M - kM \text{ sgn } (SM),\ kM > 0 \qquad (22)$$

The term $-kM$sgn (SM) is known as the corrective control.

We choose the quadratic Lyapunov function candidate $V = SM\ 2\ /2$. The system converges to the sliding mode if the derivative of a Lyapunov function is negative along all the trajectories of the system. The derivative of V is

$$\frac{1}{2}\frac{d}{dt}SM^2 = (gM - \hat{g}M - kM\text{sgn } (SM)) = (gM - \hat{g}M) - kM|SM| \qquad (23)$$

For robust convergence to the sliding mode the derivative must remain negative in the presence of uncertainties. We choose the corrective control gain kM as in eq. (24).

$$kM = GM + \eta M \qquad (24)$$

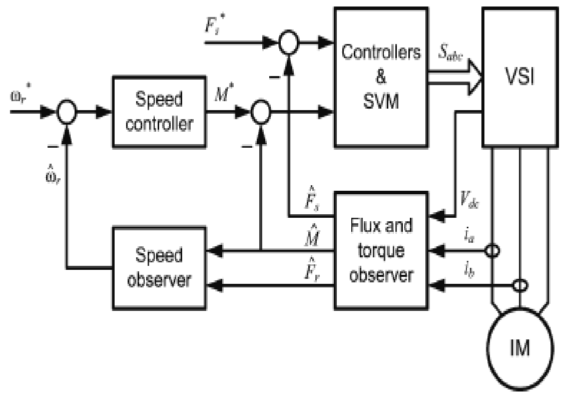


Fig. 1 Block diagram of the sensorless DTC IM drive with feedback linearization.

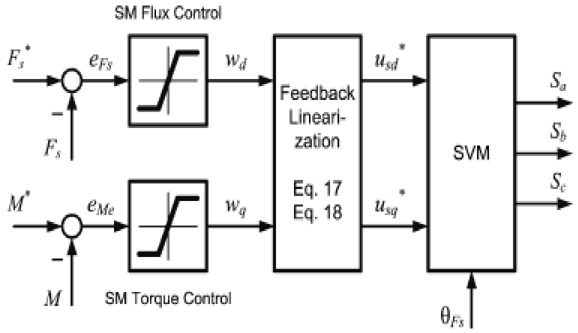


Fig. 2 Torque and flux SMC with feedback linearization for IM control.

This gives the sliding condition, eq. (25)

$$\frac{1}{2}\frac{d}{dt}SM^2 \leq -\eta M|SM| \quad (25)$$

Where ηM is a positive constant. The gain kM of (24) includes the term GM to ensure robust stability and the term ηM to control the speed of convergence to the sliding controller. A bigger ηM makes the system trajectory to reach the sliding mode in a shorter time but can result in higher chattering. Similar results can be obtained by using an integral sliding mode

$$SM = (\frac{d}{dt} + \lambda M) \int_0^t (M - Md)d\tau \quad (26)$$

Where λM is a positive constant design parameter. This parameter determines how fast the error goes to zero once the State is on the mode. The SMC effort can be chosen as

$$wq = -\hat{g}M - (M -) - kM\text{sgn}(SM), \; kM > 0 \; (\quad (27)$$

and the sliding condition holds for $kM = GM + \eta M$.

To avoid chattering we define a boundary layer around the sliding mode, $(t) = \{x, | (x)| \leq hM\}$, where $hM > 0$ is the boundary layer thickness. Inside the boundary layer, a proportional control term is added to the control of (22). Outside the boundary layer ($|(x)| > hM$), the corrective control drives the system to the sliding mode.

The stator flux dynamics in eq. (14) are almost identical to (13) and are similarly handled. Most of the analysis is omitted, for brevity. Similarly to torque, the sliding mode is

$$SFs = Fs - Fsd \quad (28)$$

and the linear system control Feedback is

$$wd = -\hat{g}Fs - kFs \text{ sgn } ( ), \; kFs > 0 \quad (29)$$

As for torque, we use a narrow boundary layer around the sliding mode, with proportional control to avoid chattering. Figure 2 shows the block diagram of the SMC with FBL torque and flux controller. To summarize, the controllers are given by (22) and (29) and the reference voltages are produced by (17) and (18) in the stator reference frame. A SVM unit produces the VSI switching signals Sa, Sb, Sc.

IV. ROBUSTNESS STUDY AND CONTROLLER DESIGN

This segment gives a design system to the sliding mode FBL controller that accomplishes powerful steadiness in face of the most imperative errors which influence the IM display: motor parameter detuning and speed perception errors. We consider these uncertainties limited, as in eq. (20) and explore how these uncertainties affect the decision of remedial additions for torque and flux control. For FBL performance we utilize consistent motor parameter esteems and design the controller to stay robustness as they change during operation. Rotor speed is gotten from observers with estimation errors, especially during homeless people and low speed operation. Then again, flux and torque observers give moderately great evaluations, and the effect of their errors on FBL is not discussed about here.

The errors in the control flux because of these uncertainties are indicated as Δu_{sd} and Δu_s. To assess these errors in terms of the rotor speed and parameter errors, and to dissect the impact of uncertainties on the SMC design we consolidate (17) described more, (18) in vector shape:

$$us = \left(\frac{wd}{2R} - \frac{LmR_s}{L_sL_r - L_m^2}\right)\psi_r + j(w_q/R + \omega_r) \quad (30)$$

Although wd and wq are produced by the SMC and have no uncertainty, we can replace the error in the control signal us with equivalent errors in wd and. The equivalent error is $\Delta w = \Delta wd + j\Delta$, and (30) can be rewritten as (31).

$$us = \left(\frac{wd+\Delta wd}{2R}\ \frac{\widehat{Lm}\widehat{R_s}}{(\widehat{Lm}+L_{s\sigma})(\widehat{Lm}+L_{r\sigma})-L_m^2}\right)\psi_r + j\left(\frac{wd+\Delta wd}{R} + \omega_r\right)\psi_s \quad (31)$$

Where $L\,\hat{}\,m$ is the measured magnetizing inductance, $R\,\hat{}\,s$ is measured stator resistance and $\omega\hat{r}$ is the rotor speed estimate.

Using (30) and (31), the equivalent error is (32).

$$\Delta w = \Delta wd + j\Delta wq = 2\left(\frac{\widehat{Lm}\widehat{R_s}}{(\widehat{Lm}+L_{s\sigma})(\widehat{Lm}+L_{r\sigma})-L_m^2} - \frac{LmR_s}{L_sL_r-L_m^2}\right)R + j(\omega_r\ \widehat{\omega}_r)R \quad (32)$$

The feedback linearized torque and stator flux dynamics in the presence of errors in wd and wq are

$$\frac{dM}{dt} = -\left(\frac{1}{Tr\sigma} + \frac{1}{T_s\sigma}\right)M + wd + \Delta wd \quad (33)$$

$$\frac{dFs}{dt} = -\frac{2}{Ts\sigma}Fs + wd - \Delta wd \quad (34)$$

It can be assumed that the maximum deviation of each uncertain parameter and the maximum measurement or estimation error for the rotor speed are known. For this analysis we use $\eta M = 10$, $\eta Fs = 10$, which give a realistic dynamic response for torque and flux. The main focus for this section is robust stability rather than dynamic response.

A. Speed (ωr)

Errors in speed estimation cause model perturbations that may influence the system response. Speed errors have no effect on stator flux dynamics but change the torque equation (13) to

$$\frac{dM}{dt} = -\left(\frac{1}{Tr\sigma} + \frac{1}{T_s\sigma}\right)M + (\widehat{\omega_r} - \omega r)R + w_q \quad (35)$$

Knowing the maximum speed estimation error, the corrective control gain can guarantee robust performance. The IM has a nominal value of R, R = 0.25 (parameters are listed in Appendix). Assuming a speed measurement with a maximum error of ±10 rad/s (±1.6 Hz), we have $|(\omega\hat{r} - \omega r)R| < 2.5$, which corresponds to $GM = 2.5$ and $kM = GM + \eta M = 12.5$. We use $kM = 20$, as in our experiments, which handles even larger errors. Since the speed error does not affect the stator flux dynamics, we use $kFs = \eta Fs + 0 = 10$. Simulation results in Fig. 3 show the torque and flux response for the drive starting from standstill with ±10 rad/s speed errors. The torque control is almost identical for any speed error and it remains stable and ripple-free. For bigger errors we simply choose a larger gain for robust stability, at the expense of increased chattering.

B. Stator resistance (Rs)

The stator resistance changes with temperature, and it impacts the stator flux dynamics. Introducing a perturbation due to stator resistance error, the stator flux dynamics (34) is

$$\frac{dFs}{dt} = -\frac{2}{Ts\sigma}Fs + \frac{2L_m}{L_sT_r\sigma}R(R_s - \widehat{R_s})wd \quad (36)$$

Where $R\hat{}s$ is the nominal stator resistance and Rs is its actual value. We consider a maximum error in the stator resistance of □50%, i.e. $|Rs - R\hat{}s| < 0.5 \times R\hat{}s = 1.15$. The corresponding model perturbation for the parameter values is $GFs = \frac{2L_m}{L_s T_r \sigma} R \times 0.69 = 28.16$. We choose the corrective control gain $kFs = \eta Fs + GFs = 40 > 38.16$. Since the torque dynamics is independent of the resistance error, we use the same value $kM = 20$, for similar dynamic performance. Simulation results in Fig. 4 show the stator flux and torque response for the drive starting from standstill with □50% stator resistance dynamic uncertainty. Note how the resistance error impact the flux response time, which is faster for lower resistances and due to larger gain. However, the steady state operation is ripple-free and robust with respect to Rs errors.

C. Rotor resistance (Rr)

Rotor resistance changes with temperature. The prominent advantage of the proposed FBL is that the changes in Rr do not change the dynamics of stator flux and torque and do not affect the control. However, they do change the dynamics of the other two state variables (R, Fr); this substantially impacts the speed estimate. Therefore, the rotor resistance errors are accounted for by speed errors discussed in section IV.A.

D. Magnetizing inductance (Lm)

The magnetizing inductance deviate from its measured value due to magnetic saturation. Changes in the magnetizing inductance create changes in together the stator and rotor inductances. This has no effect on torque dynamics, but changes the stator flux dynamics (34), as follows:

$$\frac{dFs}{dt} = -\frac{2}{Ts\sigma} Fs + \frac{2L_m}{L_s T_r \sigma} R \left(\frac{L_m}{L_s L_r - L_m^2} - \frac{\widehat{Lm}}{(\widehat{Lm} + L_{s\sigma})(\widehat{Lm} + L_{r\sigma}) - L_m^2} \right) + wd \quad (37)$$

We consider a maximum change in the magnetizing inductance of □30%, i.e. $0.7\hat{Lm}$□□Lm□$1.3\hat{Lm}$. We examine the term □L = $\frac{L_m}{L_s L_r - L_m^2} - \frac{\widehat{Lm}}{(\widehat{Lm} + L_{s\sigma})(\widehat{Lm} + L_{r\sigma}) - L_m^2}$ in (37) that depends on Lm. For $Lm = 0.7\hat{Lm}$ we have □L = –0.42467, and for $Lm = 1.3\hat{Lm}$ we have □L = 0.23176. For robust stability we use the maximum value of |□L |. The corresponding perturbation is $GFs = 2RRs \times 0.42467 = 0.49$. We use the gain $kFs = 12 > 10.49$. Since the torque dynamics is independent of the magnetizing inductance, we use $kM = 20$.

Simulation results in Fig. 5 show the stator flux and torque for the drive first from standstill with □30% magnetizing inductance errors. Again, it is proved that SMC provide robust and ripple-free steady state performance. Overall, the largest gains can be used for all situations. All simulations are for the sensorless drive shown in Fig. 1. The proposed SMC design is based on the required dynamic response (ηM, ηFs) and the maximum uncertainty (GM, GFs). The dynamic response is application-dependent and is chosen by the designer. Equation (34) gives the maximum uncertainty caused by FBL. Given η and G for flux and torque, the designer chooses a sliding gain larger than $GM + \eta M$ for the torque controller and better than $GFs + \eta Fs$ for the flux controller. This choice of the corrective control gains results in a robust and stable scheme that operates at the required speed while suppressing chattering. Comparing all simulation results, we terminate that larger gains result in a quicker and robust control, but can cause chatter if the increase in gain is excessive.

V. Extension Topic

FUZZY LOGIC CONTROLLER

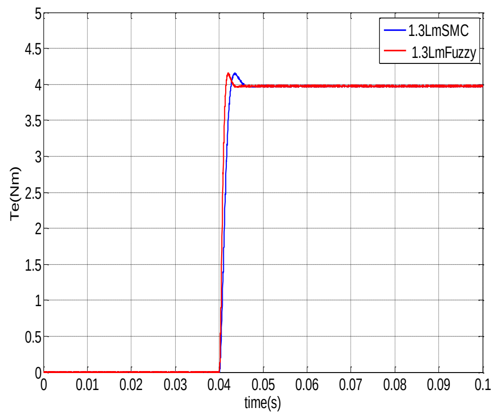
Fuzzy logic is a complex mathematical method that allows solving difficult simulated problems with many inputs and output variables. Fuzzy logic is able to give results in the form of recommendation for a specific interval of output state, so it is essential that this mathematical method is strictly distinguished from the more familiar logics, such as Boolean algebra.

Advantages of Fuzzy Controller over PI Controller

Usage of conventional control "PI", its reaction is not all that great for non-linear systems. The change is striking when controls with Fuzzy logic are utilized, acquiring a superior dynamic reaction from the system.

The PI controller requires exact direct numerical models, which are hard to get and may not give tasteful execution under parameter varieties, load unsettling powers, and so forth. As of late, Fuzzy Logic Controllers (FLCs) have been presented in different applications and have been utilized as a part of the power devices field. The benefits of fuzzy logic controllers over ordinary PI controllers are that they needn't bother with a precise scientific model, Can work with uncertain information sources and can deal with non-linearities and are more dynamic than traditional PI controllers.

VI. SIMULATION RESULTS



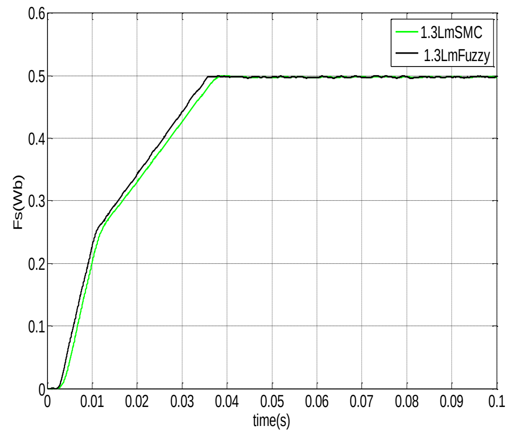
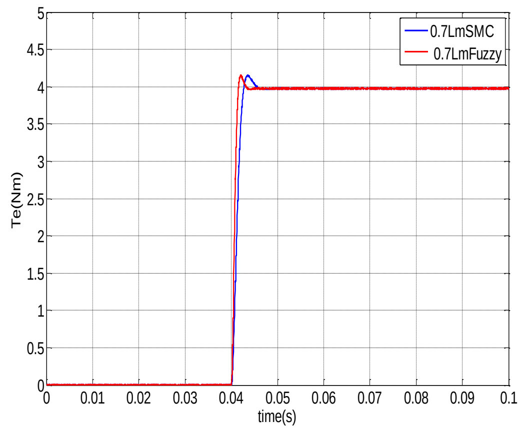


Fig 3: Simulation results for SMC and FBL for proposed and extinction with +30% L_m errors, at startup, torque Te, and stator flux □s.



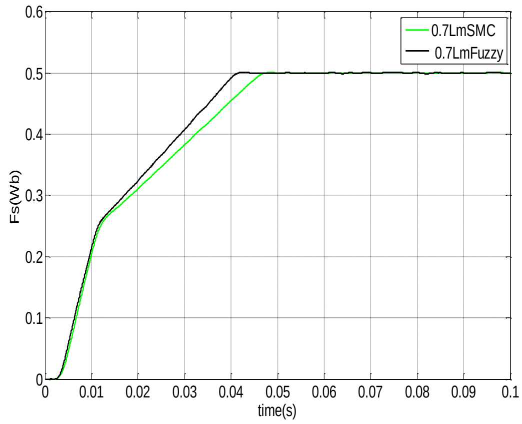
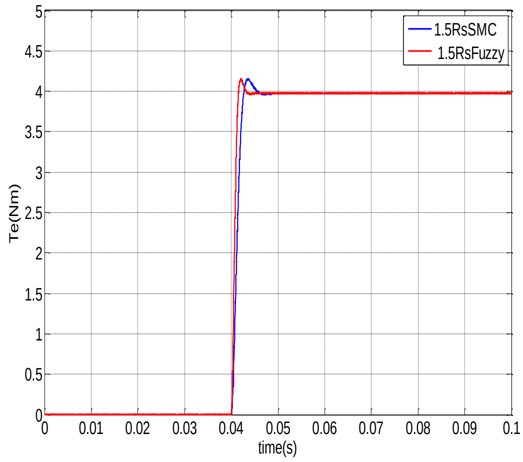


Fig4: Simulation results for SMC and FBL for proposed and extinction with -30% L_m errors, at startup, torque Te, and stator flux □s.



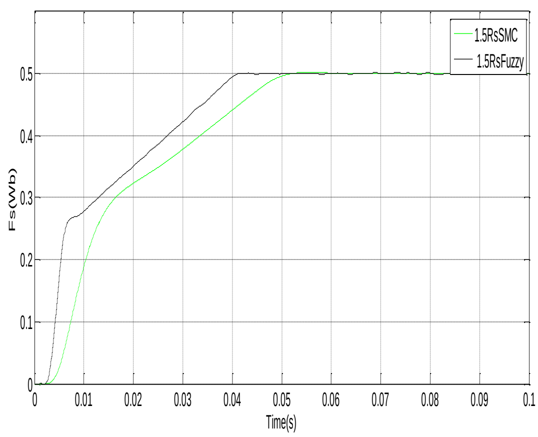
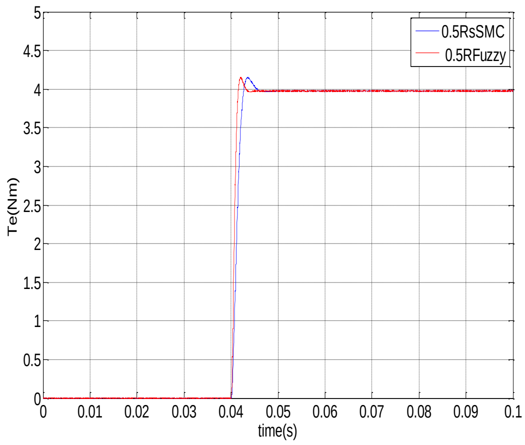


Fig 5: Simulation results for SMC and FBL for proposed and extinction with +50% Rs errors, at startup, torque Te, and stator flux ψs.



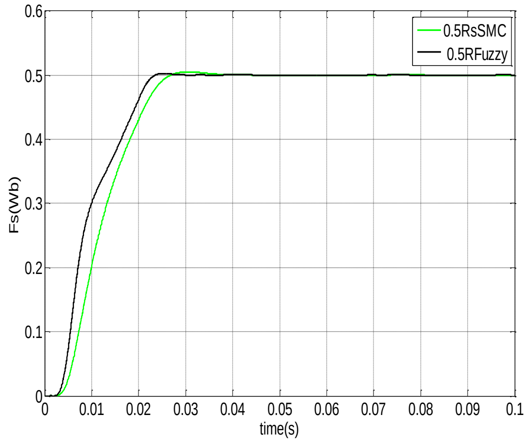
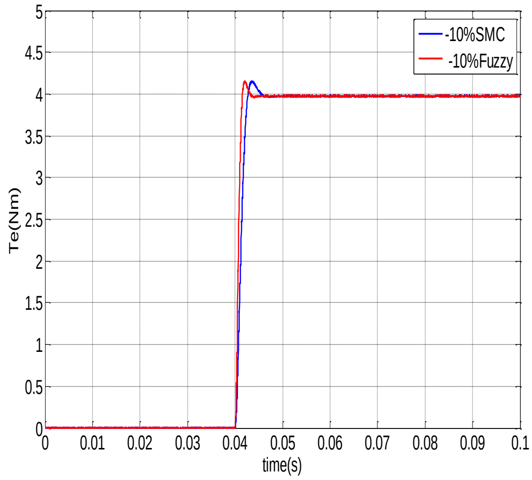


Fig 6: Simulation results for SMC and FBL for proposed and extinction with -50% Rs errors, at startup, torque Te, and stator flux ψs.



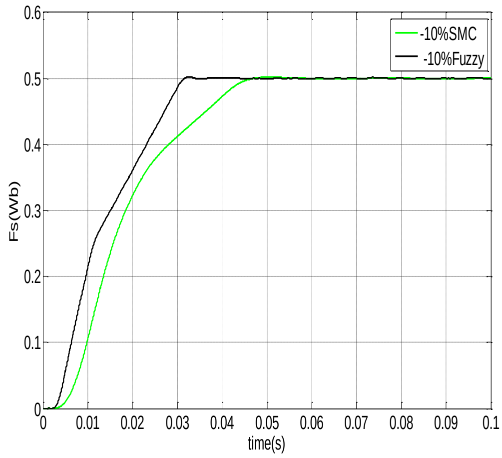


Fig 7: Simulation results for SMC and FBL for proposed and extinction with -10 rad/s speed errors, at startup, torque Te, and stator flux ψs.

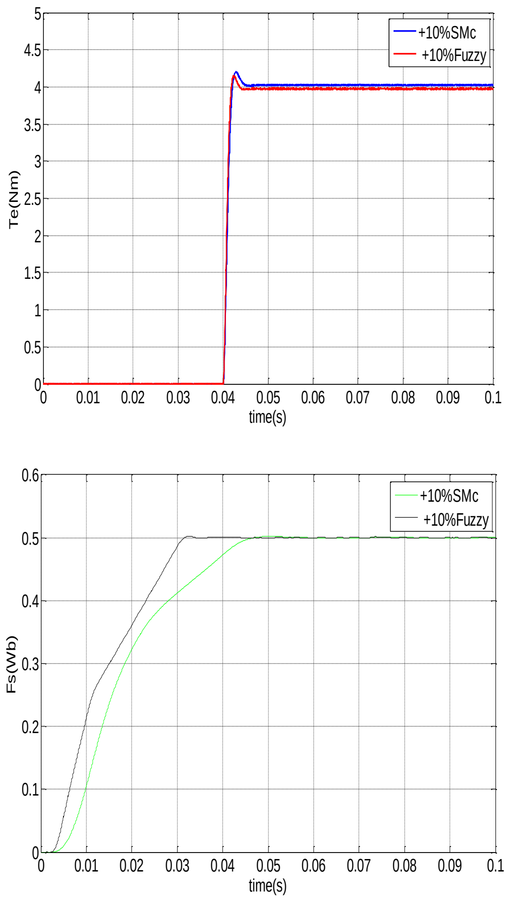
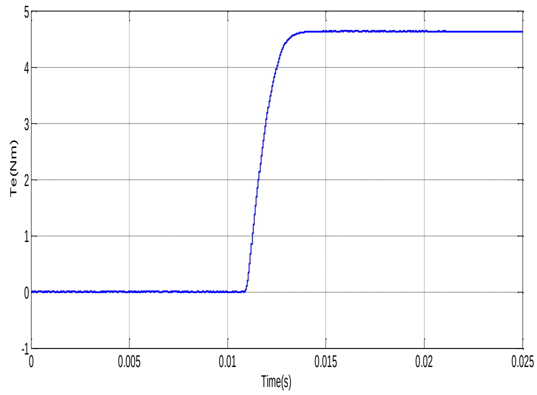
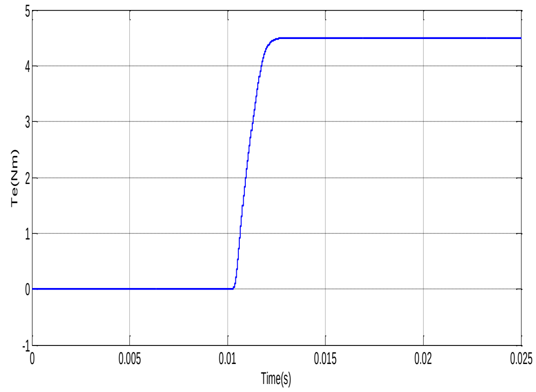


Fig 8: Simulation results for SMC and FBL for proposed and extinction with +10 rad/s speed errors, at startup, torque Te, and stator flux ψ_s.

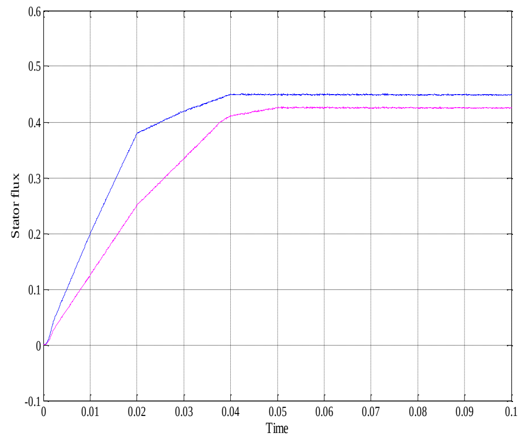


(a)

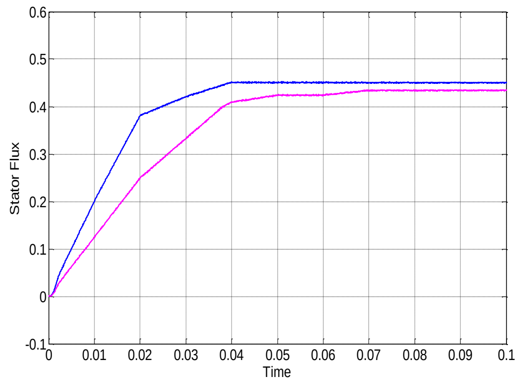


(b)

Fig 9: Torque response to 4.5 Nm step command for proposed and extinction with (a) PI controllers (Linear DTC) and (b) PI controllers and FBL. Startup from standstill.

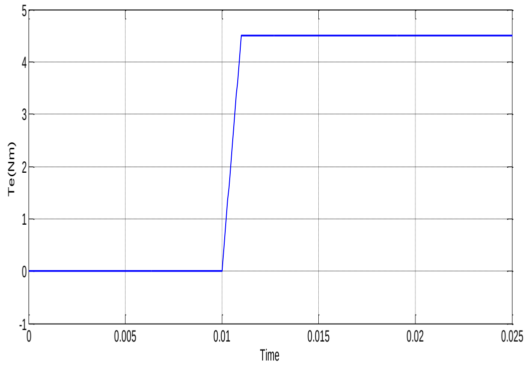


(a)

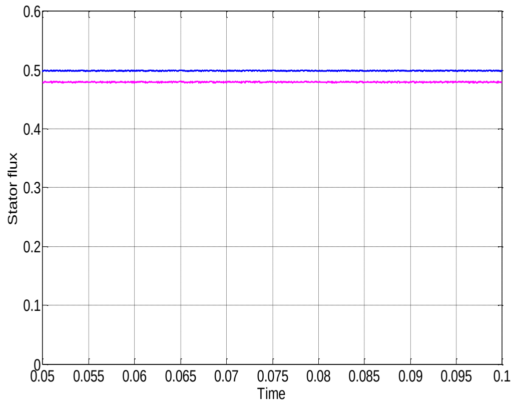


(b)

Fig 10: Stator (blue) and rotor (red) flux magnitude control at startup, for proposed and extinction with (a) PI controllers (Linear DTC) and (b) PI controllers and FBL.

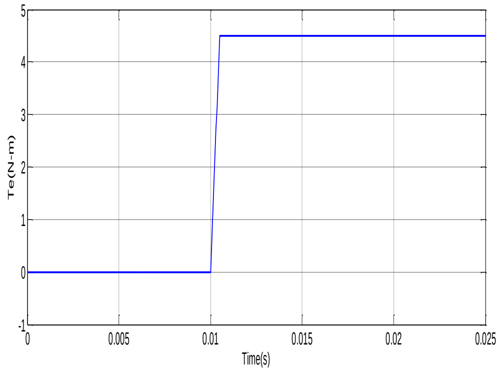


(a)



(b)

fig 11: Torque transients for startup from standstill with feedback linearization and SMC (a) torque, (b) stator and rotor flux magnitudes.



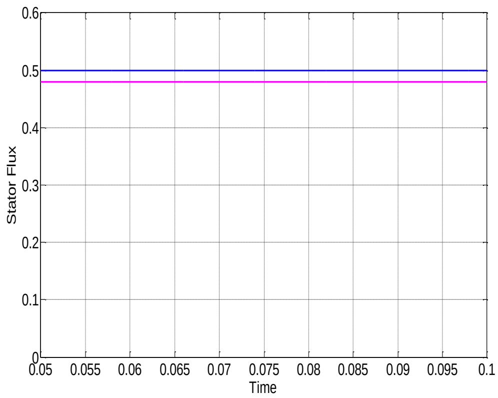


fig 12: Stator (blue) and rotor (red) flux magnitude response to 0.5 Wb step command for proposed and extinction with feedback linearization and SMC, at standstill.

VII. CONCLUSION

This paper proposes another design advance which incorporates Feedback linearization and Fuzzy logic controller with a DTC drive. This novel agreement in light of torque-flux linearization creates an automatic linear model of the IM, with torque and flux as decoupled state factors. For the linear IM show, the controller-observer partition guideline holds if estimation errors are little, which permit the controller and observer to be autonomously designed.

Fuzzy logic controller direct torque and flux control gives robustness against parameter uncertainties and their dynamics, as demonstrated by the correlation with a linear controller. The chattering related with sliding mode operation is disposed of by the corresponding controller utilized inside the limit layer. The drive has a similar quick and robust reaction, as a regular DTC drive and totally disposes of the torque and flux swell. Generally speaking, the arrangement consolidates the benefits of ordinary and linear DTC. These favorable circumstances are because of the sliding mode controller and the linearization which decouples the torque and stator flux extent.

VIII. REFERENCES

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