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J. Electrical Systems 13-4 (2017): 723-741

Regular paper

Robustness Enhancement of MRAC Using Modification Techniques for Speed Control of Three Phase Induction Motor

JES
Journal of Electrical Systems

This paper addresses the lack of robustness in three phase Induction Motors (I.M) speed control in presence of uncertainty and external unknown disturbance of Model Reference Adaptive Control (MRAC), and present an enhancement of robustness by employing the adaptive law modifications for robustness. Considered modifications were Dead zone modification, Sigma modification, e-modification and projection operator algorithm modification. Speed control of three phase induction motor in this paper was depend on decoupling benefits of rotor flux Indirect Field Orientation (IFO). Simulation results shows improved robustness characteristics and performance. In order to make the case study under consideration more realistic, the work considered the limitation imposed on supplied voltage in real application, where this was the other challenge which has been avoided by many researchers.

Keywords: MRAC; Dead zone modification; Sigma modification; e-modification; Projection operator modification; three phase I.M; IFO.

Article history: Received 12 June 2017, Accepted 10 November 2017

1. Introduction

Due to its simple structure, high torque capability and no need for maintenance the three phase induction motors became one of the most used motors in wide range of applications [1]. The robust driving of such machines is a challenging problem which requires a concrete solution [2].

In any application of the three phase induction motor, it is recommended to get robust speed characteristics. Productivity cycle includes a varying loads imposed on motor shaft which adds an uncertainty on the driving system. Control design of motor speed with that uncertainty is difficult without using a control methodology possess adaptation behavior like adaptive control. Recently, Model Reference Adaptive Control (MRAC) has acquainted significant attention due to its ability to adjust controller elements in real-time situation. MRAC in presence of disturbance suffer from Drift phenomenon; which appears in adaptive law gains [3]. Parameters drift phenomenon has an adverse effect on stability; where increasing in adaptive gains leads to high gain problem which as an effect leads to instability problems or to controller wind up where the controller get saturated [4]; these effects on stability caused by parameters drift addresses the lack of robustness.

In theory, a Model Reference Adaptive Controller can drive the plant to asymptotically track a specified reference model if certain matching conditions on the uncertainties are met. However, practically, such matching conditions usually do not fit due to the presence of matched and unmatched unmodeled dynamics, disturbances and parameterization errors. This may lead to weakening of stability property of Uniform Ultimate Boundedness (UUB) and also a parameter drift would arise which in turn may lead to instability problems and to lack of robustness. In order to cope with such types of uncertainties and to have a robust MRAC, this requires a modification in the adaptive law of MRAC strategy. As such, several structures of modification have appeared like dead-zone modification, σ-

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modification e-modification, Projection operator modification and others [5]. Boundedness of the adaptive gains got from applying the modification addresses the relevance of the modification to the robustness of the modified MRAC.

Different related studies are briefly reviewed here; Firstly implantation of the scalar control for controlling the open-end winding induction machine on FPGA has been proposed by (Abdelmonoem.Nayli, etal) [6]. Also, Direct Torque control of induction motor using Fuzzy logic switching controller are presented by (H. Sudheer and etal) [7]. Fuzzy PID Controller for Fast Direct Torque Control of Induction Motor Drives method for control has been proposed by (Mohamed E. El-Shimy and etal) as indicated in [8]. Two techniques for the induction machine "IM" a chattering-free Sliding Mode Controller based on Integrated Control (SMIC) and a PI based control technique are performed by (Arafet Ltifi and etal) in [9]. New techniques of fuzzy gain scheduling of PID (FGS-PID) for speed control three phase induction motor based on indirect field oriented control (IFOC) are performed by (Indra Ferdiansyah and etal) in [10]. Speed observer and reduced nonlinear model for sensorless control of induction motors has been proposed by (Hassan K. Khalil and etal) [11].

It is clear from the above literature survey that most researchers are interested in controller design and no mention of how well the robustness will be in presence of uncertainty effect on the rotor angular speed due to unknown rotor inertia and loads torque.

Robust adaptive control theory are still seeking for new aspects for improving and enhancing the performance in electrical machines and that was the motivation of the present work [12],[13].

The main key of work contribution is to introduce robust adaptive speed controller for the three phase induction motor in the presence of unknown and bounded load torque, unknown rotor inertia and unknown viscous friction coefficient. Therefore, the following objectives are the milestone of the work:

- Estabilshment of indirect field orented control (IFOC) induction motor (IM)
- Stability analysis of modified model reference adaptive controllers for IFOC I.M..
- Investigating, via simulations, the performances of both classic MRAC and robust MRACs in terms of robust characteristics under considered uncertainties.

2. Robust MRAC based on Modification Techniques

2.1. MRAC

In Lyapunov direct MRAC technique, the interested system class is in the form of:

$$\dot{\boldsymbol{x}} = \boldsymbol{A}\,\boldsymbol{x} + \boldsymbol{B}\,\boldsymbol{\Lambda}\left(\boldsymbol{u} + \boldsymbol{\theta}^T \boldsymbol{\phi}\right) + \boldsymbol{\zeta} \quad , \quad \boldsymbol{y} = \boldsymbol{C}\,\boldsymbol{x} \tag{1}$$

where the unknown matrices were; $\boldsymbol{A} \in \mathcal{R}^{n\times n}$, $\boldsymbol{\Lambda} \in \mathcal{R}^{m\times m}$, $\boldsymbol{\theta} \in \mathcal{R}^{N\times m}$ and the known matrices were; $\boldsymbol{B} \in \mathcal{R}^{n\times m}$, $\boldsymbol{\phi} \in \mathcal{R}^{N\times m} \to \mathcal{R}^{N\times m}$ (function matrix), $\boldsymbol{C} \in \mathcal{R}^{q\times n}$; and $\boldsymbol{y} \in \mathcal{R}^{q}$ is output vector , $\boldsymbol{x} \in \mathcal{R}^{n}$ is state vector. The vector $\boldsymbol{\zeta}$ represents the mismatched part of the unwanted functions terms of the state variables with upper bound availability and also represents the exogenous disturbance terms. A prerequisite assumption requires that ϕ must be *Lipchitz* continuous [14],[15].

According to the given reference model, a suitable control algorithm must be chosen to make the system of Eq.(1) tracks the reference model with minimum error; such that the system must behave as follows:

$$\dot{\boldsymbol{x}}_m = \boldsymbol{A}_m \boldsymbol{x}_m + \boldsymbol{B}_m \boldsymbol{r} \tag{2}$$

where:

$A_m \in \mathcal{R}^{n\times n}$ is reference model matrix, $B_m \in \mathcal{R}^{n\times m}$ is reference input matrix, $x_m \in \mathcal{R}^n$ is reference state vector and $r \in \mathcal{R}^m$ is the desired input trajectory. Other assumptions are required which include that the two matrices (A, B) have to be controllable and A_m must be *Hurwitz* [14],[15].

In order to design a controller u able to make the system of Eq.(1) tracks the desired input r based on the specifications of the reference model system of Eq.(2) the controller must have the following responsibilities:

- ❑ Canceling the nonlinearity of $\theta^T \phi$.
- ❑ Modifying the matrix A to be A_m (feedback gains).
- ❑ Modifying the matrix B to be B_m (feed forward gains).
- ❑ Dealing with the uncertainty of Λ (controller ineffectiveness).
- ❑ Keeping some level of performance robustly in presence of ζ (Modification for Robustness).

Choosing the control input as, $u = \hat{K}_x^T x + \hat{K}_r^T r - \hat{\theta}\phi$ and for space saving in the equations it will be written in vector form as;[2]:

$$u = \hat{K}^T \Psi \tag{3}$$

where, $\hat{K}^T = \begin{bmatrix} \hat{K}_x^T & \hat{K}_r^T - \hat{\theta}^T \end{bmatrix}$, $\Psi = \begin{bmatrix} x & r & \phi \end{bmatrix}^T$, $\hat{\theta} \in \mathcal{R}^{N\times m}$ is a matrix of an estimated values of θ, $\hat{K}_x \in \mathcal{R}^{m\times n}$ is a matrix of an estimated values of the ideal feedback gains, $\hat{K}_r \in \mathcal{R}^{m\times m}$ is a matrix of an estimated values of the ideal feed forward gains.

Matching condition must be satisfied to mimicking the reference mod or tracking the desired input with desired specifications. Mathematically, existence of matching condition means that the controller is capable to satisfy the following:

$$\mathbf{A} + \mathbf{B}\,\Lambda\, K_x^T = A_m, \quad \mathbf{B}\,\Lambda\, K_r^T = B_m, \qquad \Delta\theta^T = \hat{\theta}^T - \theta^T = 0 \tag{4}$$

where K_x^T, K_r^T and θ^T represent the ideal feedback and feed forward and unwanted functions parameters coefficients, respectively. Closed loop system will be:

$$\dot{x} = Ax + B\,\Lambda\left(\hat{K}^T\ \Psi + \theta^T \phi\right) + \zeta \tag{5}$$

and the error between the closed loop system and the reference model will be:

$$e = x - x_m \tag{6}$$

The time derivative of the error will be:

$$\dot{e} = A_m e + B\ \Lambda\ \Delta K^T\ \Psi + \zeta \tag{7}$$

According to ;[2],[14],[15] choosing the Lyapunov candidate for the error system of Eq.(7) as:

$$V\left(e, \Delta K^T\right) = e^T P e + tr\left(\Delta K^T \Gamma_K^{-1} \Delta K\ \Lambda\right) \tag{8}$$

where $V\left(e, \Delta K^T\right) : \mathcal{R}^{n\times m} \to \mathcal{R}$; and $P \in \mathcal{R}^{n\times n}\ |P = P^T > 0$ is a positive definite symmetric matrix. The matrix P is calculated by solving the equation $PA_m + A_m^T P = -Q$ where $Q \in \mathbb{R}^{n\times n}$ is a positive definite symmetric matrix; and $\Gamma_K \in \mathcal{R}^{z\times z}\ |\Gamma_K = \Gamma_K{}^T > 0$, $z = n + m + N$ is a positive definite symmetric diagonal matrix (adaptation constants rate). The derivative of $V(e, \Delta K^T)$ will be:

$$\dot{V} = -\boldsymbol{e}^T \boldsymbol{Q} \boldsymbol{e} + 2tr\{\Delta \boldsymbol{K}^T (\boldsymbol{\Gamma}_K^{-1} \dot{\hat{\boldsymbol{K}}} - \boldsymbol{\Psi} \boldsymbol{e}^T \boldsymbol{P} \boldsymbol{B}\ \boldsymbol{\Lambda})\} + 2\boldsymbol{e}^T \boldsymbol{P} \boldsymbol{\zeta} \tag{9}$$

Negative definiteness of Lyapunov time derivative depends on both the adaptive law [2];

$$\dot{\hat{\boldsymbol{K}}} = \boldsymbol{\Gamma}_K \boldsymbol{\Psi}\ \boldsymbol{e}^T \boldsymbol{P}\ \boldsymbol{B} \tag{10}$$

and the disturbance $\boldsymbol{\zeta}$ is equal to zero when the effect of disturbance is not taken into account, then by considering the adaptive law of Eq.(10), Lyapunov time derivative becomes;

$$\dot{V} = -\boldsymbol{e}^T \boldsymbol{Q}\ \boldsymbol{e} \tag{11}$$

which it is a negative semi-definite and the error system can be driven to the origin uniformly. According to *Barbalat's* lemma, it can be shown that the error is asymptotically stable.

With presence of bounded disturbance case, $\|\zeta\| \le \zeta_{max}$, the analysis of Uniform Ultimate Boundedness (UUB) is applied to show bounded convergence of the error. Starting with Eq.(9) together with adaptive law described by Eq.(10), we have:

$$\dot{V} = -\boldsymbol{e}^T \boldsymbol{Q} \boldsymbol{e} + 2\boldsymbol{e}^T \boldsymbol{P}\ \boldsymbol{\zeta} \tag{12}$$

Using *Rayleigh-Ritz* quotient and, from elementary linear algebra, the following inequality must hold:

$$\dot{V} \le -\lambda_{min}(\boldsymbol{Q})\|e\|^2 + 2\|e\|\lambda_{\max}(\boldsymbol{P})\zeta_{max} \tag{13}$$

Therefore, based on above equation, the negative definiteness of $\dot{V}$ is assured if either $\|e\| > 0$ or $\|e\| > e_o$; where $e_o = \frac{2\lambda_{\max}(P)\zeta_{max}}{\lambda_{min}(Q)}$ is the ultimate bound and $\dot{V}$ will be a negative definite outside the set:

$$\beta = \{(\boldsymbol{e}, \Delta\boldsymbol{K}^T) : \boldsymbol{e} \in \mathcal{R}^n, \Delta\boldsymbol{K} \in \mathcal{R}^{N\times m} \mid \|e\| \le e_0\} \tag{14}$$

One can notes that there is no mention of $\Delta\boldsymbol{K}$ inside or outside the set β; $\Delta\boldsymbol{K}$ represent the convergence of the estimated parameters to unknown parameters of the adaptive law or to any combination that drive the error to the equilibrium point asymptotically. One may argue that uniform ultimate boundedness, which lead the error trajectory to zero equilibrium point, may be conflicted with set β; as the asymptotic stability is violated inside the set and the equilibrium error will finally reside on the bound confined by the set and there error may never reaches zero unless the boundedness of set β is set to zero.

As such, to confine boundedness of adaptive parameters modification techniques are introduced on the adaptive law such that enforcing all signals to be bounded. In other words, a robust adaptive controller has to be developed in the framework of direct Lyapunov method to overcome the "parameters drift" phenomenon and keeping the adaptive gains bounded without effecting on some level of the overall system performance.

2.2. Modification for Robustness

The parameter drift of adaptive gains caused by the presence of disturbance or uncertainty may lead to instability problems. The modification to adaptive law is introduced to keep adaptive gains confined and bounded. Four modification techniques are considered in the present work. They are namely Dead Zone modification, Leakage or Sigma "σ-modification", e-modification and Projection Operator algorithm modification. The stability analysis of each technique based on *Lyapunov* direct method is presented here in details.

2.2.1. Dead Zone Modification

Dead Zone modification was firstly developed based on adaptation hibernation principle [16]. The basic idea of this modification is to stop the adaptive law as soon as the error trajectory touches the disturbance-dependent boundary around the equilibrium point. Such boundary can be described mathematically by the set β_d ,

$$\beta_d = \{(\boldsymbol{e}, \Delta\boldsymbol{K}^T) : \boldsymbol{e} \in \mathcal{R}^n, \Delta\boldsymbol{K} \in \mathcal{R}^{N\times m} \mid \|e\| \le e_d\} \tag{15}$$

The adaptive law will be disabled when the error reaches the value of e_d in the neighborhood of the error equilibrium point, that is ;[2];

$$\dot{\hat{\boldsymbol{K}}} = \begin{cases} -\boldsymbol{\Gamma}_K \Psi \boldsymbol{e}^T PB & \text{if } \|e\| > e_d \\ 0_{N\times m} & \text{if } \|e\| \le e_d \end{cases} \tag{16}$$

As long as the error $\|\boldsymbol{e}\|$ is larger than ($e_d = e_o$), the stability is guaranteed and the adaptive controller works normally. If the error touches the boundary described by the set β_d , then the adaptive controller stops adaptation; as there is no meaning of adaptation inside the disturbance ball. Stopping the adaptation produces bounded adaptive gains which satisfy the requirement of the robustness.

2.2.2. Sigma modification (σ-modification):

In spite of its simplicity, dead zone modification acquired some problems; especially where a prior knowledge about the upper bound of the disturbance is required. The basic idea of the sigma modification (σ-modification) was to add damping term to the adaptive law as follows ;[2];

$$\dot{\hat{\boldsymbol{K}}} = \boldsymbol{\Gamma}_K (\boldsymbol{\Psi} \boldsymbol{e}^T \boldsymbol{PB} - \sigma \hat{\boldsymbol{K}}) \tag{17}$$

where $\sigma > 0$. This modification takes different forms depending on the choice of σ , which may be scalar or a function [4],[17]. Again, the derivative of $V\left(\boldsymbol{e}, \Delta\boldsymbol{K}^T\right)$ will be

$$\dot{V} = -\boldsymbol{e}^T \boldsymbol{Q} \boldsymbol{e} + 2 \operatorname{tr}\left\{\left(\Delta\boldsymbol{K}^T \left(\boldsymbol{\Gamma}_K^{-1} \boldsymbol{\Gamma}_K (\boldsymbol{\Psi} \boldsymbol{e}^T \boldsymbol{PB} - \sigma \hat{\boldsymbol{K}}\right) - \boldsymbol{\Psi} \boldsymbol{e}^T \boldsymbol{PB}\right) \boldsymbol{\Lambda}\right\} + 2\boldsymbol{e}^T \boldsymbol{P} \boldsymbol{\zeta} \tag{18}$$

If the last norm is incorporated together with *Rayleigh-Ritz* inequality and *Schwarz* inequality we get, $\dot{V} < 0$ if the following inequalities are both satisfied $\|\Delta\boldsymbol{K}\| > \Delta K_\sigma$, $\|e\|(-\lambda_{min}(\boldsymbol{Q})\|e\| + 2\lambda_{max}(\boldsymbol{P})\zeta_{max}) < 0$ where, $\Delta K_\sigma = \sqrt{\dfrac{\|\boldsymbol{K}\|_F^2 \|\Lambda\|_F}{(2\Lambda_{min} + \|\Lambda\|_F)}}$ The above second condition is satisfied if both $\boldsymbol{e} < 0$ and $\boldsymbol{e} < e_\sigma$ are satisfied, where $e_\sigma = e_0$. Based on the above analysis, the region of stability can be guaranteed outside the following compact set

$$\beta_\sigma = \{(\boldsymbol{e}, \Delta\boldsymbol{K}^T) : \boldsymbol{e} \in \mathcal{R}^n, \Delta\boldsymbol{K} \in \mathcal{R}^{N\times m} \|e\| < e_\sigma \wedge \|\Delta\boldsymbol{K}\|_F < \Delta K_\sigma\} \tag{19}$$

From the set described by β_σ , it is clear that this modification guarantees the uniform ultimate boundedness of error, $\boldsymbol{e} = \boldsymbol{x} - \boldsymbol{x}_m$, and it also guarantees the boundedness of all adaptive gains and control action signal. Boundedness of the adaptive gains and control action gotten from applying the modification addresses the relevance of the modification to the robustness of the modified MRAC. It is worthy to notice that the size of the set β_σ is dedicated by the upper bound of the disturbance ζ_{max} and when the disturbance effect is removed, then the size of the convex set β_σ would vanish and the error trajectory would finally settle at zero equilibrium points.

2.2.3. e- modification:

Adding term in the sigma modification may not be convenient in some situations. In absence of disturbance and in nearby of equilibrium point with small norm disturbance, the damping term of the sigma modification dominates over the original adaptive law. This in turn will weaken the learning mechanism and unlearn phenomenon will appears which drive the adaptive gains to the zero in some occasions.

A magic trick achieved by *e*-modification is to reduce the unbounded behavior of the adaptive law due to drift by adding an error dependent leakage term to adaptive law [18],[19]. In accordance to the e-modification, the adaptive law will be ;[2] :

$$\dot{\hat{\boldsymbol{K}}} = -\boldsymbol{\Gamma}_K (\boldsymbol{\Psi}\, \boldsymbol{e}^T \boldsymbol{PB} - \varepsilon \boldsymbol{e}^T \boldsymbol{PB}\, \hat{\boldsymbol{K}}) \tag{20}$$

where $\varepsilon > 0$ is a constant chosen by the designer.

The same argument as in σ -modification is followed here so,

$$\dot{V} = -\boldsymbol{e}^T \boldsymbol{Qe} + 2\, tr\{ \Delta\boldsymbol{K}^T (\boldsymbol{\Gamma}_K^{-1} \boldsymbol{\Gamma}_K (\boldsymbol{\Psi}\boldsymbol{e}^T \boldsymbol{P}\,\boldsymbol{B} - \varepsilon \boldsymbol{e}^T \boldsymbol{PB}\,\hat{\boldsymbol{K}}) - \boldsymbol{\Psi}\boldsymbol{e}^T \boldsymbol{PB})\boldsymbol{\Lambda}\} + 2\boldsymbol{e}^T \boldsymbol{P}\,\boldsymbol{\zeta} \tag{21}$$

The condition $\dot{V} < 0$ is strictly satisfied if the following inequalities hold:

$$\|e\| > 0 \vee \|e\| > \mathrm{e_e} \wedge \|\Delta\boldsymbol{K}\|_F > \Delta K_e \tag{22}$$

where $\mathrm{e_e} = e_0$, $\Delta K_e = \Delta K_\sigma$ .Inequalities of Eq.(22) tells that the stability of e-modification, $V > 0$ and $\dot{V} < 0$, is guaranteed outside the compact set:

$$\beta_e = \{(\boldsymbol{e}, \Delta\boldsymbol{K}^T):\ \boldsymbol{e} \in \mathcal{R}^n, \Delta\boldsymbol{K} \in \mathcal{R}^{N\times m} \mid \|e\| > \boldsymbol{e}_\mathrm{e} \wedge \|\Delta\boldsymbol{K}\|_F > \Delta K_e\} \tag{23}$$

The set β_e shows that e-modification guarantees the uniform ultimate boundedness of error, $e = \boldsymbol{x} - \boldsymbol{x}_m$, and also it guarantee boundedness of all adaptive gains and the actuating control signal as well. Boundedness of the adaptive gains and control action gotten from applying the modification addresses the relevance of the modification to the robustness of the modified MRAC. The upper bound of the set β_e is determined by the upper bound of the disturbance ζ_{max}; if the disturbance is removed, then the set β_e will shrink to zero at equilibrium point and the error trajectory could finally reaches to equilibrium points.

2.2.4 Projection Operator modification:

Projection operator algorithm in adaptive control theory was firstly adopted by *Pomet* and *Parly* [20]. Later projection operator become a widely used in adaptive control theory in various methodologies. The projection operator is used if it is required to confine the adaptive law solution or trajectory within defined region such that it never leaves it at all. Suppose that a parameter vector η belongs to a convex set $\beta_0 = \{\eta \in \mathcal{R}^n \mid f(\eta) \le 0\}$ and there is another convex set $\beta_1 = \{\eta \in \mathcal{R}^n \mid f(\eta) \le 1\}$, it becomes obvious that $\beta_0 \subseteq \beta_1$ as indicated in Figure(1). As seen from Figure(1), *Proj* (η, y) does not change the vector y if η belongs to convex set β_0. In the annulus set $\{0 \le f(\eta) \le 1\}$, the Projection Operator subtracts a vector normal to the boundary $\{f(\eta) = \lambda\}$ from y . As a result, we get a smooth transformation from the original vector field y for $\lambda = 0$ to the tangent to the boundary vector for $\lambda = 1$. If the above analysis is applied to adaptive law dynamics, i.e., $\dot{\hat{\boldsymbol{K}}} = \boldsymbol{\Gamma}\ \boldsymbol{\Psi}\boldsymbol{e}^T \boldsymbol{PB}$ and the requirement is to limit the adaptive gains such that they are not to exceed specified limits, then the following Projection Operator holds where $\hat{\boldsymbol{K}}$ represents η and $\boldsymbol{\Gamma}\ \boldsymbol{\Psi}\boldsymbol{e}^T \boldsymbol{PB}$ represents [2] ;

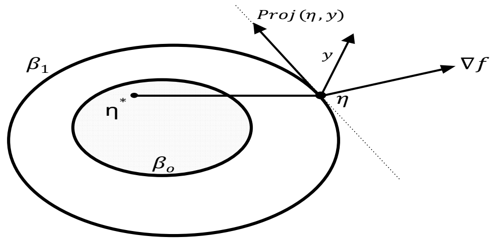


Figure (1) *The Projection Operator*

$$\dot{\hat{\boldsymbol{K}}} = Proj(\boldsymbol{K}, \boldsymbol{\Gamma}\ \boldsymbol{\Psi}\boldsymbol{e}^T \boldsymbol{PB}, f(\hat{\boldsymbol{K}})) \tag{24}$$

where,

$$Proj(\hat{\boldsymbol{K}}, \boldsymbol{\Gamma}\boldsymbol{\Psi}\boldsymbol{e}^T\boldsymbol{PB}, f(\hat{\boldsymbol{K}})) = \boldsymbol{\Gamma}\begin{cases} \boldsymbol{\Psi}\boldsymbol{e}^T PB - \Gamma\dfrac{\nabla f \nabla f^T \boldsymbol{\Gamma}\boldsymbol{\Psi}\boldsymbol{e}^T \boldsymbol{P}\boldsymbol{B}}{\|\nabla \boldsymbol{f}\|_\Gamma^2} f(\hat{K}),\ if \quad f(\hat{K}) > 0 \wedge (\boldsymbol{\Psi}\boldsymbol{e}^T \boldsymbol{PB})^T \nabla f > 0 \\ \boldsymbol{\Psi}\boldsymbol{e}^T PB \qquad\qquad otherwise \end{cases} \tag{25}$$

A convex function $f(\hat{\boldsymbol{K}})$ is a prerequisite for *Projection Operator* algorithm where $\boldsymbol{\Gamma} \in \mathcal{R}^{n\times n}$ is a constant symmetric positive-definite matrix, ∇f represent the Gradient vector of convex function $f(\hat{\boldsymbol{K}})$ and $\|\nabla f\|_\Gamma^2 = \nabla f^T \boldsymbol{\Gamma} \nabla f$ is the weighted Euclidean squared norm of.

In what follows, we shall investigate how to force the trace term to be semi negative via the matrix version of the Projection Operator algorithm while enforcing uniform boundedness of the corresponding solutions $\dot{\hat{\boldsymbol{K}}}$. Starting with time derivatives of the selected Lyapunov function

$$\dot{V} = -\boldsymbol{e}^T\boldsymbol{Q}\boldsymbol{e} + 2\ \text{trace}(\Delta\boldsymbol{K}^T\{\boldsymbol{\Gamma}_K^{-1}\dot{\hat{\boldsymbol{K}}} - \boldsymbol{\Psi}\boldsymbol{e}^T\boldsymbol{PB}\ \}\boldsymbol{\Lambda}) + 2\boldsymbol{e}^T\boldsymbol{P}\boldsymbol{\zeta} \tag{26}$$

and the design task was to choose $\dot{\hat{\boldsymbol{K}}}$ such that the trace term in Eq.(42) became non-positive, while the adaptive parameters $\hat{\boldsymbol{K}}$ remained uniformly bounded functions of time.

$$\text{trace}(\underbrace{\Delta\boldsymbol{K}^T}_{(\eta-\eta^*)^T}\{\boldsymbol{\Gamma}_K^{-1}\underbrace{\dot{\hat{\boldsymbol{K}}}}_{Proj(\eta,\ \boldsymbol{\Gamma}\mathrm{y})}\underbrace{\boldsymbol{\Psi}\boldsymbol{e}^T\boldsymbol{PB}}_{y}\}\boldsymbol{\Lambda}) = \sum_{j=1}^{M}\left(\hat{\boldsymbol{K}} - \boldsymbol{K}\right)_j^T \underbrace{\{\boldsymbol{\Gamma}_K^{-1}Proj\left(\hat{K}, \boldsymbol{\Gamma}y_j\right) - y_j\}}_{\le 0}\underbrace{\lambda_j}_{\ge 0} \le 0 \tag{27}$$

where the diagonal elements of $\boldsymbol{\Lambda}$ are greater than zero.

Thus, the term of the inequality (27) can be considered as a bargain because it contributes to make the *Lyapunov* candidate time derivative negative definite. So, $\dot{V}\left(\mathbf{e}, \Delta\mathbf{K}\right) < 0$ outside the set:

$$\beta_p = \{(\boldsymbol{e}, \Delta\boldsymbol{K}^T):\ \boldsymbol{e} \in \mathcal{R}^n, \Delta\boldsymbol{K} \in \mathcal{R}^{N\times m} \mid \|e\| < \boldsymbol{e}_e \wedge \|\Delta\boldsymbol{K}\|_F < \Delta\boldsymbol{K}_{max}\} \tag{28}$$

where $\Delta\boldsymbol{K}_{max} = 2(\underbrace{\boldsymbol{K}_1^{max} \cdots \boldsymbol{K}_m^{max}}_{\boldsymbol{K}_{max}}) = 2\boldsymbol{K}_{max}$ and $\boldsymbol{K}_j^{max}$ is the maximum allowable bound for the j^{th} column $\hat{\boldsymbol{K}}_j(t)$. This formal argument proves UUB property of all signals in the corresponding closed-loop system. Boundedness of the adaptive gains and control action gotten from applying the modification addresses the relevance of the modification to the robustness of the modified MRAC.

3. Field Oriented Control of Squirrel cage three phase Induction Motor

In the analysis of induction machine, there are three commonly used frames; the arbitrary rotating frame, stationary reference frame and synchronously rotating frame. The induction machine dynamics in arbitrary rotating frame can be described by the following system of differential equations [1],[21],[22]

$$\begin{aligned} d\lambda_{qs} / dt &= v_{qs} - r_s\, i_{qs} \\ d\lambda_{ds} / dt &= v_{ds} - r_s\, i_{ds} \\ d\lambda'_{qr} / dt &= v'_{qr} - (\omega - \omega_r)\,\lambda'_{dr} - r'_r\, i'_{qr} \\ d\lambda'_{dr} / dt &= v'_{qr} + (\omega - \omega_r)\,\lambda'_{dr} - r'_r\, i'_{dr} \end{aligned} \tag{29}$$

where ω, ω_r are the angular speed of frame and machine rotor, respectively. The primed rotor quantities denote referred values to the stator side. Before we proceed, it is worthy to mention that

- ❒ Particularly in adjustable speed drives, it is suitable to simulate I.M and its converter on a stationary reference frame.
- ❒ The establishment of field orientation technique is based on dynamic model on synchronous frame.

The stationary and synchronously rotating reference frame are indicated in machine dynamic by setting the speed of arbitrary reference frame, ω, to zero and ω_e , respectively where ω_e represent the synchronous speed . In addition, one can distinguish whether the machine dynamic is in stationary frame or in synchronous rotating frame by attaching the superscript s (for stationary frame) and e (for synchronous rotating frame). To be specific, the machine dynamic in synchronous rotating frame can be depicted as follows:

$$\begin{aligned} d\,\lambda^e_{qs} / dt &= v_{qs} - r_s\, i^e_{qs} \\ d\,\lambda^e_{ds} / dt &= v_{ds} - r_s\, i^e_{ds} \\ d\,\lambda'^e_{qr} / dt &= v'^e_{qr} - (\omega_e - \omega_r)\,\lambda'^e_{dr} - r'_r\, i'^e_{qr} \\ d\,\lambda'^e_{dr} / dt &= v'^e_{dr} + (\omega_e - \omega_r)\,\lambda'_{dr} - r'_r\, i'^e_{dr} \end{aligned} \tag{30}$$

The mechanical part of the induction motor obeys *Newton*'s second law of force and torque balance [2]:

$$J \frac{d}{dt} \omega_{rm} = T_e - T_{load} \tag{31}$$

where J represents the rotor inertia, ω_{rm} represents the rotor mechanical speed, T_e represents the electromechanical torque produced by the motor and T_{load} represents the load torque on the rotor shaft. This load torque constitutes from internal viscous friction hindering F_{vs} and the applied load to be rotated and thus described by $T_{load} = T_l + F_{vs}\,\omega_{rm}$. Hence Eq.(31) can be rewritten as [2]:

$$J\, d\omega_{rm} / dt = T_e - T_l - F_{vs}\,\omega_{rm}$$

One of the expression used to express the developed electromechanical torque is given by [1];

$$T_e = \frac{3}{2}\frac{P}{2}\left(\lambda'^e_{qr}\, i'^e_{dr} - \lambda'^e_{dr}\, i'^e_{qr}\right) \tag{32}$$

Field Orientation (FO) had more than one scheme depending on stator, rotor, air gap flux orientation; flux orientation is the main core of the FO. In flux orientation, the flux is projected into two orthogonal axes; the quadrature axis (q-axis) flux will govern the torque control loop and direct axis (d-axis) flux will govern the speed control loop. A filed controlled of three phase induction motor would gain the induction motor the characteristics of DC motor.

The stator and rotor flux linkage relationships with stator and Rotor currents can be expressed compactly as,

$$\begin{bmatrix} \lambda_{qs} \\ \lambda_{ds} \\ \lambda'_{qr} \\ \lambda'_{dr} \end{bmatrix} = \begin{bmatrix} L_{ls}+L_m & 0 & L_m & 0 \\ 0 & L_{ls}+L_m & 0 & L_m \\ L_m & 0 & L'_{lr}+L_m & 0 \\ 0 & L_m & 0 & L'_{lr}+L_m \end{bmatrix} \begin{bmatrix} i_{qs} \\ i_{ds} \\ i'_{qr} \\ i'_{dr} \end{bmatrix} \tag{33}$$

where L_m is the magnetizing inductance, L_{ls} is the leakage inductance of stator and L'_{lr} leakage inductance of rotor referred to the stator side; Eq.(33) is very important for simulating the I.M model and implementing Simulink Function block of the Indirect Field Orientation.

Indirect Field Orientation (IFO) is the most commonly used field orientation technique which is based no rotor orientation. The basic concept of rotor flux *Indirect Field Orientation* is to orientate all the rotor flux in the d-axis of the synchronously rotating frame such that no flux linkage in direction of q-axis; i.e., $\lambda'^{e}_{qr} = 0$. If this is satisfied, then from Eq.(33), we have [1];

$$i'^{e}_{qr} = -\frac{L_m}{L_r} i^{e}_{qs} \tag{34}$$

Taking into account no flux linkage in direction of q-axis and substitute Eq.(34) into Eq.(32), then

$$T^{*}_{e} = \frac{3PL_m}{4L'_r} \lambda'^{e}_{dr} i^{e}_{qs} \tag{35}$$

It is clear from above equation that the electromechanical torque can be controlled by the q-axis stator current only if the term of λ'^{e}_{dr} can be hold fixed. Since no flux is present in q-axis, then $d\left(\lambda'^{e}_{qr}\right)/dt = 0$. Also, if squirrel cage rotor is considered, the voltage across rotor winging is equal to zero; i.e., $v'^{e}_{qr} = 0$. Using Eq.(35), we get

$$\omega_{sl} = \frac{r'_r\ L_m}{L'_r} \frac{i^{e}_{qs}}{\lambda'^{e}_{dr}} \tag{36}$$

where $\omega_{sl} = (\omega_e - \omega_r)$ represents the slip speed. From Eq.(36), the slip speed also can be controlled by the stator q-axis current if the term of λ'^{e}_{dr} can be hold fixed. Holding the term of λ'^{e}_{dr} fixed means that $d\left(\lambda'^{e}_{dr}\right)/dt = 0$ and this fact together with Eq.(33) leads to $\lambda'^{e}_{dr} = L_m i^{e}_{ds}$ and hence Eq.(36) becomes

$$\omega_{sl} = \frac{r'_r\, i^{e}_{qs}}{L'_r\, i^{e}_{ds}} \tag{37}$$

Eq.(37) tells that the slip speed can be controlled by stator current in direct and quadrature axes. If i_{ds}^{e} is kept constant, then slip speed is only controlled by i_{qs}^{e} [21],[22]. Figure (2) describes the mechanism of implementing speed control of IFOC three phase induction machine based on MRAC.

4. Robust MRAC for Speed Control of Field Oriented Three Phase Induction Machine

MRAC based filed orientation controlled induction machine is shown in Figure (2). The mechanical system of the induction motor can be rewritten as [2]:

$$\frac{d}{dt}\omega_{rm} = -\frac{F_{vs}}{J}\omega_{rm} + \frac{1}{J}u - \frac{1}{J}T_l \tag{38}$$

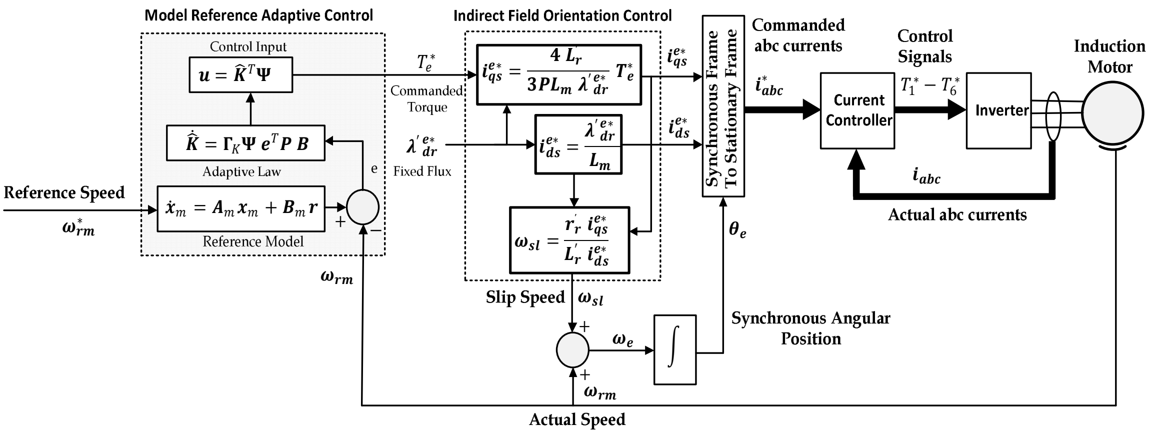


Figure (2) Speed Control of IFOC three phase induction machine based on MRAC

where u represents the control input (T^*_e the commanded torque) and F_{vs} is viscous friction coefficient. Model Reference adaptive controller would inject the IFO induction motor with required quadrature stator current i_{qs}^* such that actual speed ω_{rm} would follow reference speed subjected to the following uncertainties;

- ❒ Uncertain inertia J (due to mechanical load system inertia uncertainty or modeling uncertainty).
- ❒ Uncertain load torque T_l (upper bound known).
- ❒ Uncertain viscous friction coefficient F_{vs}.

Due to the above uncertainties, it is expected that in steady state the actual speed would approaches model reference response within a bound which is dictated by the uncertainty level and may not coincide with it at all. To put Eq.(38) into the form described in Eq.(1), we have

$$\dot{\omega}_{rm} = a\,\omega_{rm} + b\,\Lambda\,u + \zeta_1 \tag{39}$$

where $a = -F_{vs}/J$ which is unknown, $b = 1$ which is known, $\Lambda = 1/J$ which is also unknown and $\zeta_1 = -(P/2J)T_l$ stands for upper bound uncertainty and it is known. The reference model of the system is given by [2];

$$\dot{x}_m = A_m x_m + B_m\,\omega_r^* \tag{40}$$

where $x_m \in \mathbb{R}^2$ represent the reference model state vector and A_m is given by

$$A_m = A - BK_m \tag{41}$$

where K_m is the ideal feedback gain which guarantees the motor speed tracks the desired speed profile ω_r^* with bounded error and satisfy the desired specification [2].

Let us define $\boldsymbol{x} = \begin{bmatrix} e_{yi} & \omega_{rm} \end{bmatrix}^T \in \mathbb{R}^2$ such that $\dot{e}_{yi} = e_y$ and $e_y = \omega_{rm} - \omega_r^*$ represent the integral errors. The Eq.(39) can be augmented in the integral state feedback control form as [2]:

$$\dot{\boldsymbol{x}} = \boldsymbol{A}\,\boldsymbol{x} + \boldsymbol{B}\,\boldsymbol{\Lambda}\, u + \boldsymbol{B}_m\,\omega_{des} + \boldsymbol{\zeta} \tag{42}$$

Where $\boldsymbol{A} = \begin{bmatrix} 0 & 1 \\ 0 & \mathrm{a} \end{bmatrix}$, $\boldsymbol{B} = \begin{bmatrix} 0 \\ b \end{bmatrix}$. , $\Lambda = 1/J$, $\boldsymbol{B}_m = \begin{bmatrix} -1 \\ 0 \end{bmatrix}$, $\boldsymbol{\zeta} = \begin{bmatrix} 0 \\ \zeta_1 \end{bmatrix}$.The control action will be chosen as [2]:

$$u = \hat{\boldsymbol{K}}\,\boldsymbol{x} \tag{43}$$

where $\hat{\boldsymbol{K}}$ represents the adaptive gain vector which guarantees the system output ω_{rm} to track the desired speed ω_r^*; adaptive gain vector are gotten by solving the adaptive law nonlinear equation on real time once for the non-modified MRAC case Eq.(10) and for four modifications of MRAC cases Eq's.(16,17,20 and 24) respectively and make the conclusion due to connection between mathematical analysis and simulation observations.

5. Simulated results

In this work the reference model is chosen with respecting of high level of integral feed forward controller for tracking and disturbance rejection and with excellent feedback to feed forward ratio controller; hence reference model can constructed with [2]. The control action u described by Eq.(43) is the basis of the machine speed control. The outcome of the adaptive law is fed directly to control action which in turn results in commanded torque required to compensate any change in speed. Repeatedly retuning the diagonal elements of matrix $\boldsymbol{\Gamma}$ would lead to a suitable performance with the following setting [2]; A positive definite matrix P_m has been proved by solving *Lyapunov* equation where the $\boldsymbol{Q}$ matrix setting is given by [2]: $A_m = \begin{bmatrix} 0 & 1 \\ -300 & -40 \end{bmatrix}$, $\boldsymbol{\Gamma} = \begin{bmatrix} 0.05 & 0 \\ 0 & 0.05 \end{bmatrix}$, $\boldsymbol{Q} = \begin{bmatrix} 0.01 & 0 \\ 0 & 500 \end{bmatrix}$.

5.1 Speed Control of IFOC Induction Motor based on conventional MRAC

In what follows, the speed behavior of IFOC I.M is instigated based on classic MRAC. The motor is assumed to be load free. Figure (3) shows the responses of model reference speed and motor actual speed when subjected to step commanded speed of height (104.72 rad/sec) which equal to (1000 r.p.m).

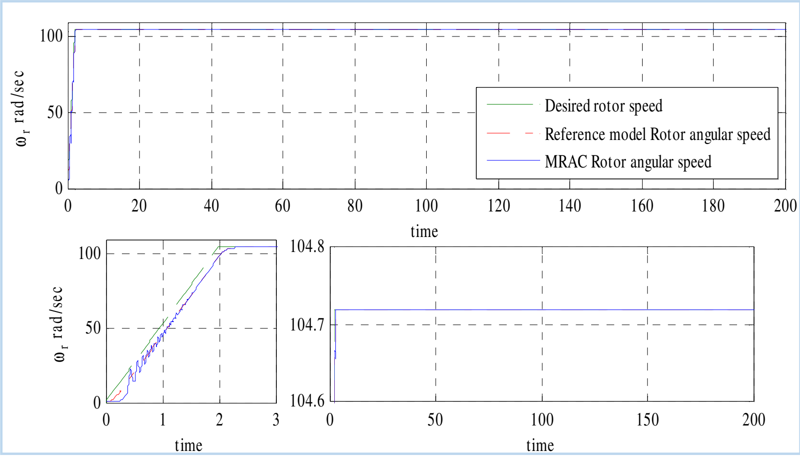


Figure (3) *Rotor angular speed in (Rad/sec)*

Figure (4) and (5) show the behavior of feed forward and feedback gains, respectively. It is evident from Fig.(4) that K_{ff}. tends to increase without bound in spite of its considerable small change. However, for long operation of system this gain would reach to infeasible high values that might lead to instability problems; this is the key behind suggesting modification techniques which work to limit the adaptive gains from drifting and to confine them inside prescribed region.

In the next scenarios, a load torque is applied to induction motor and the responses of rotor speed and adaptive gains are again monitored under classical MRAC. The torque profile is depicted in Figure (6) and the rotor angular speed behavior in (*Rad/Sec*) is shown in Figure (7).

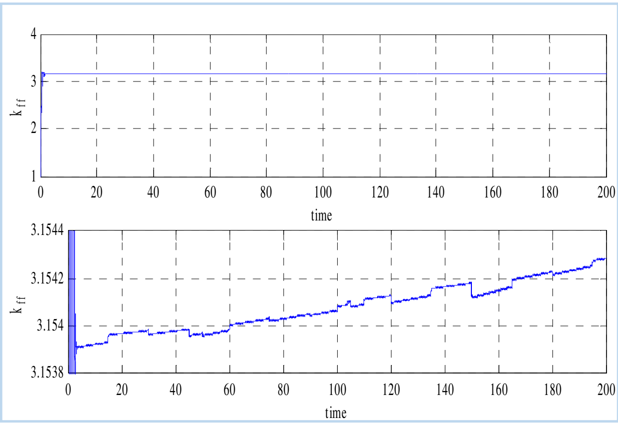


Figure (4) feed forward adaptive gain

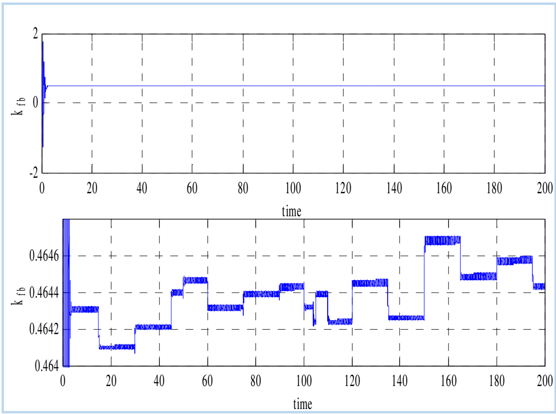


Figure (5) feedback adaptive gain

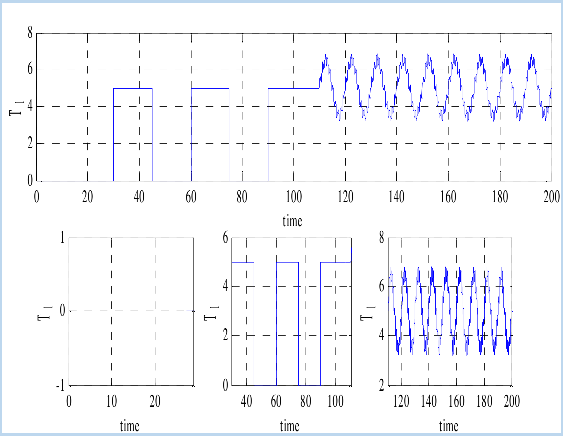


Figure (6) Load torque in (Rad/Sec)

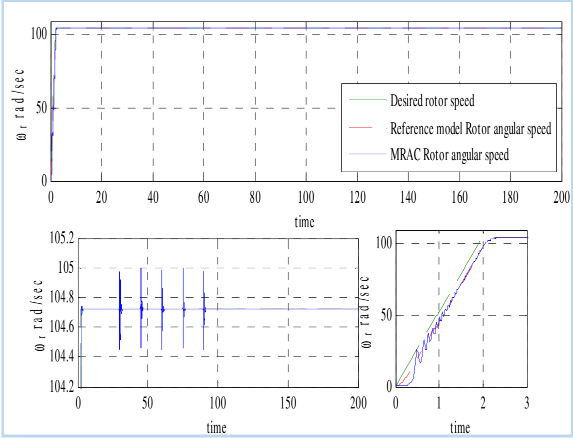


Figure (7) Rotor angular speed in (Rad/Sec)

The adaptive feed forward and feedback gains are displayed in Fig. (8) and Fig. (9), respectively. It is clear from the figures that a drift would appear in both feed forward and feedback gains which indicate that classical MRAC has no robust characteristics.

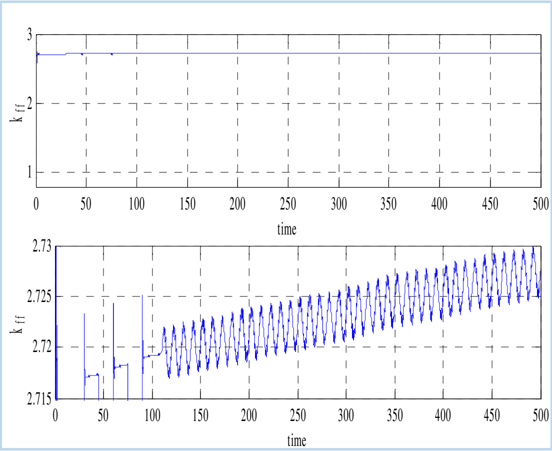


***Figure (8)** feed forward adaptive gain cycle*

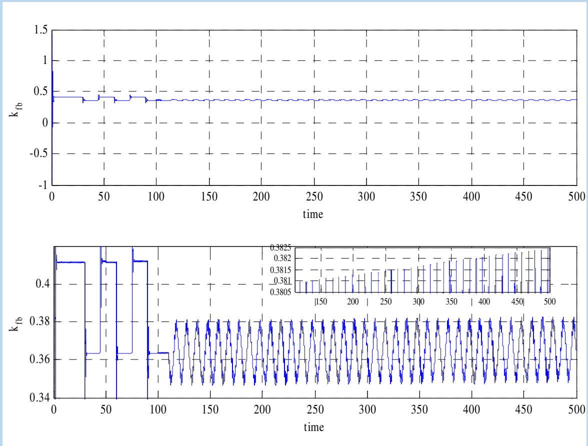


***Figure (9)** feedback adaptive gain cycle*

5.2 Speed Control of IFOC Induction Motor based on Modified MRAC

In what follows, the modified model reference adaptive controllers are included to control the speed of a squirrel cage three phase induction motor. Although the effect of modifications is considerably small on transient and steady state characteristics, some differences have been seen in system performance and will be highlighted.

In case of no-load torque, the speed responses of rotor due to different robust adaptive controllers are illustrated in Fig.(10). It is evident that the e-modification results in least variance of error at the steady state as compared to that given by others.

Figure (11) and (12) shows the speed responses of rotor due to modified MRAC controllers over all period of exerted load torque. It is clear from the figures that e-modification give less bounded error at steady state than others in both step and sinusoidal variations of load.

Fig.(13) and (14) show the behaviors of feed forward and feedback adaptive gains, respectively, for all adaptive techniques subjected to different load changes over the total simulation time. Firstly, all techniques could successfully confine the adaptive gains bounded. Secondly, the behavior of gains differs from technique to another. It is evident that projection operator modification could regain adaptive gain faster to low bound. On the other hand, due its damping characteristics, σ -modification could bring the adaptive gain to lower bound level than projection operator could give.

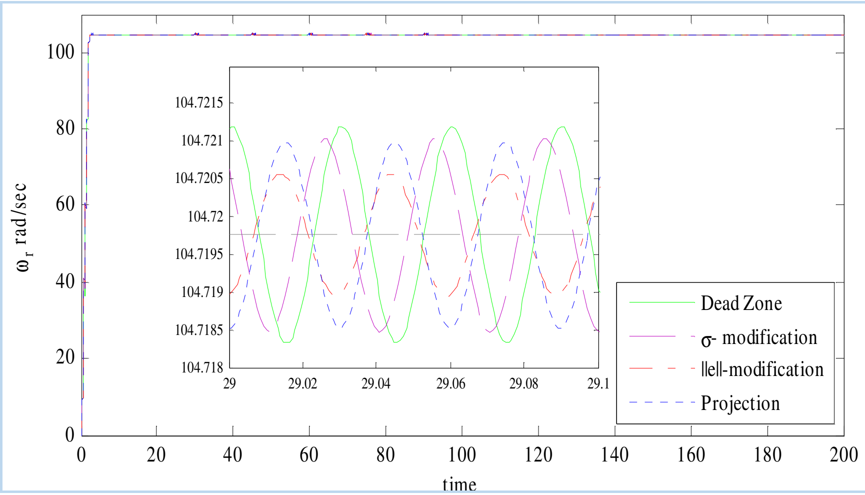
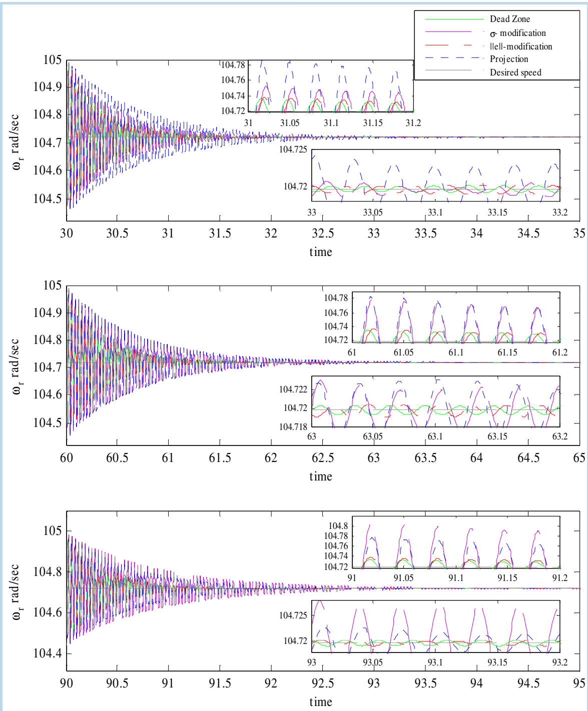


Figure (10) *Rotor angular speed comparison in (Rad/Sec)*



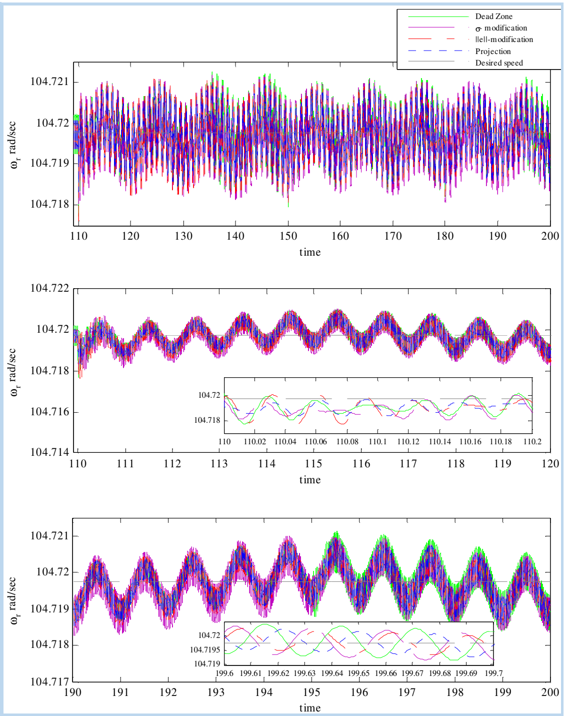
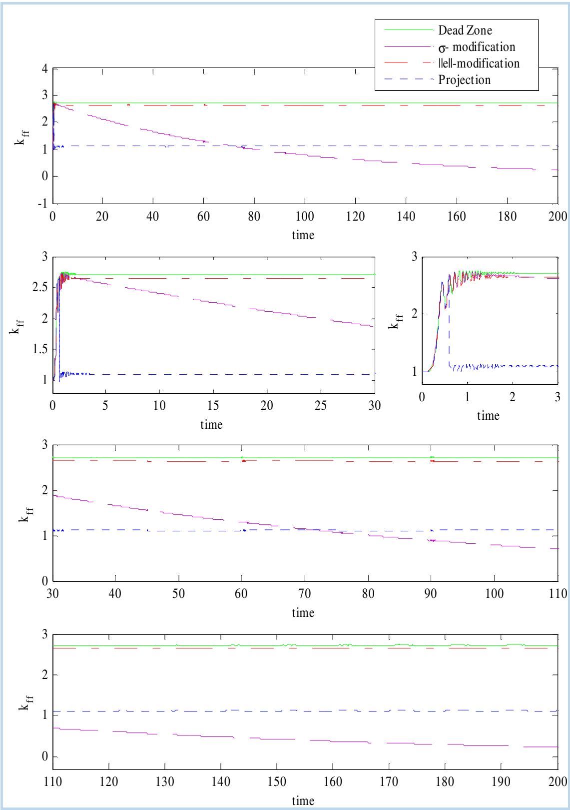


Figure (11) *Rotor angular speed due to step load change (rad/sec)*

Figure (12) *Rotor angular speeds due to sinusoidal load change (rad/sec)*

Figure (15) shows the behaviors of commanded torque, commanded q-axis stator current and developed torque for various modified adaptive controllers. Figure (16) are the zoomed-out versions of previous ones. It is clear from the figure that dead zone modification shows relatively better transients of commanded toque and current over other schemes. The transients of commanded signals due to dead zone show lowest amplitude than that resulting from other strategies.



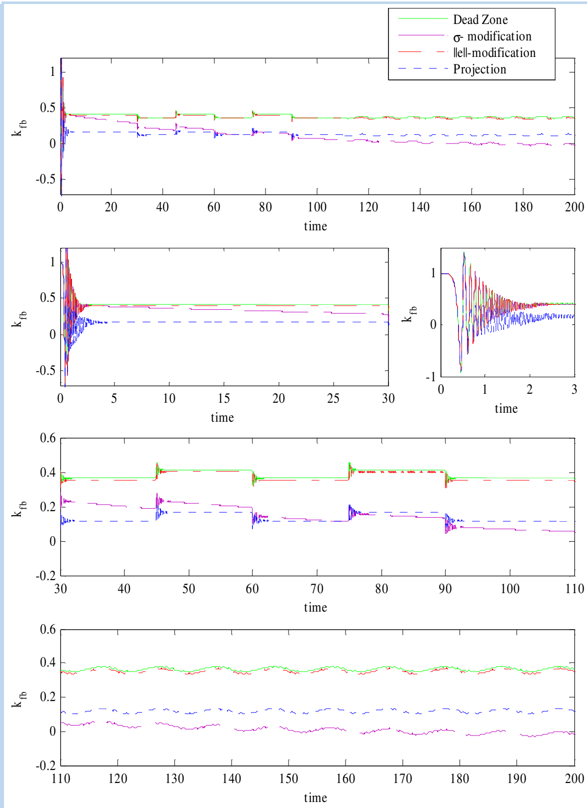
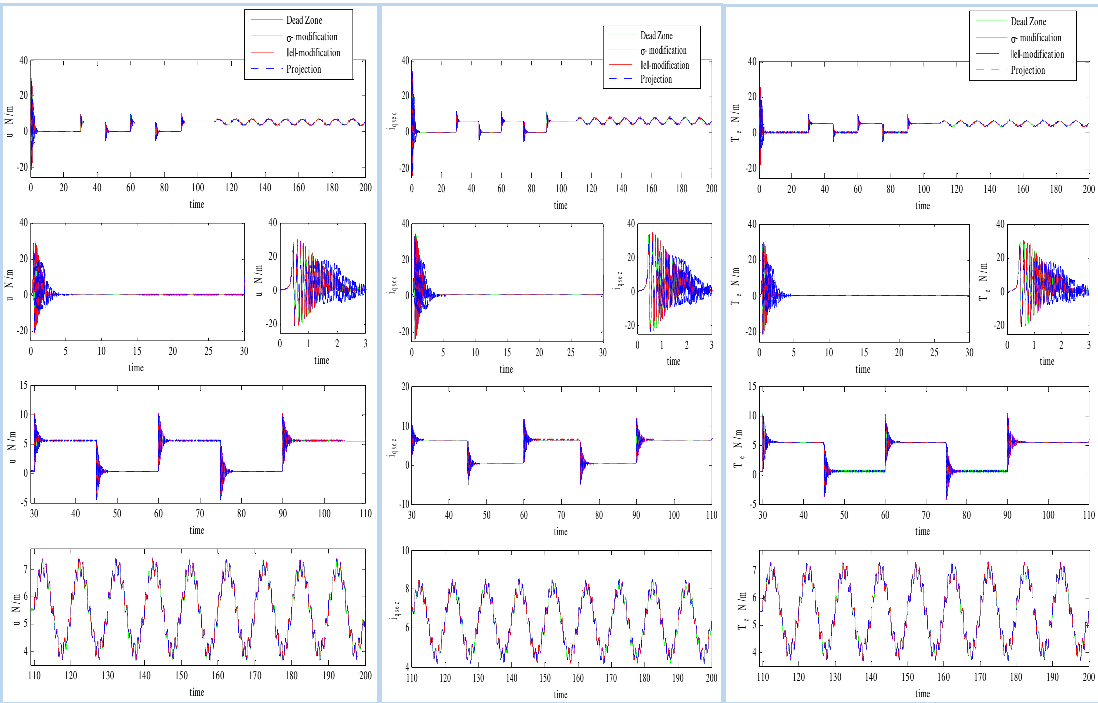


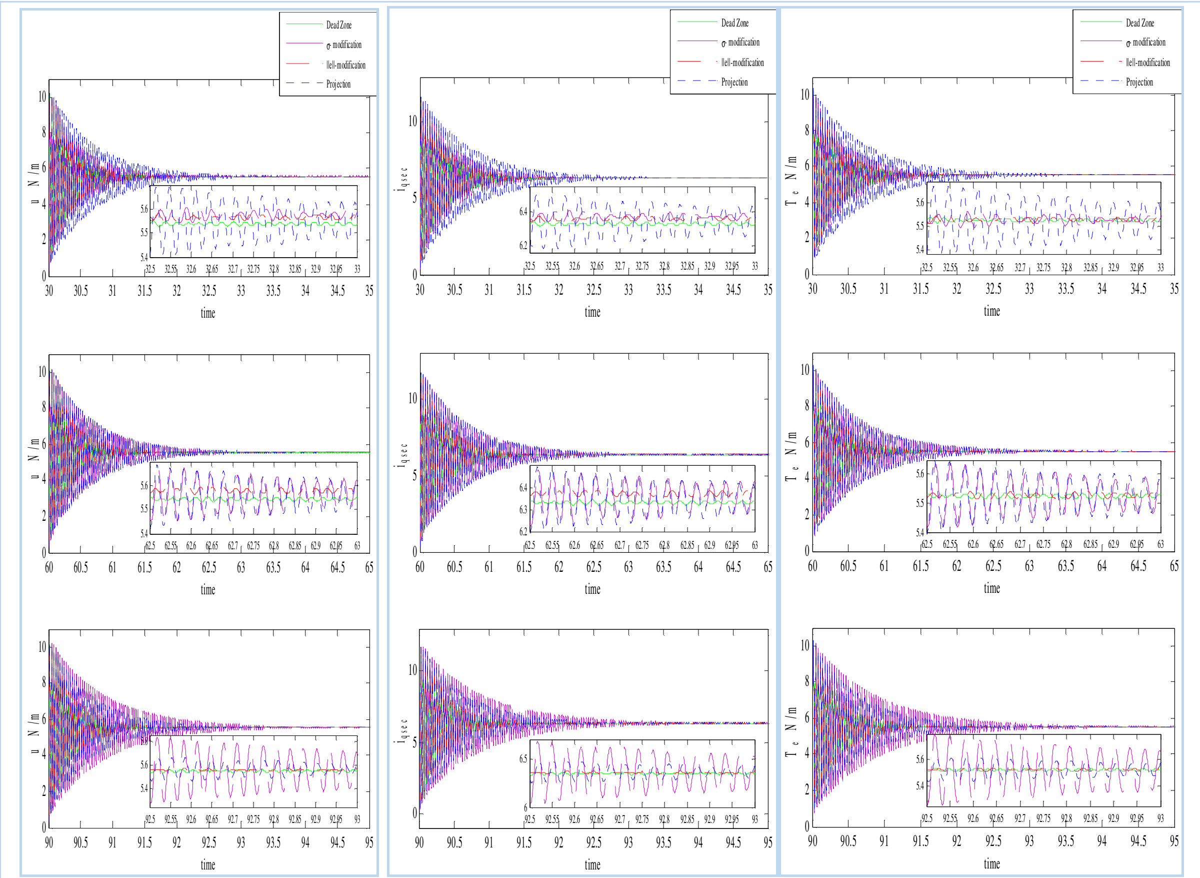
Figure (13) *Behavior of adaptive feed forward gains due to step load change*

Figure (14) *Behavior of adaptive feedback gains due to step load change*



(a) *Commanded torque* ***(b)*** *Commanded q-axis current* ***(c)*** *Developed torque*

Figure (15) *Commanded torque, commanded q-axis stator current and developed torque*



(a) Commanded torque *(b) Commanded q-axis current* *(c) Developed torque*

Figure (16) *Zoomed versions of commanded torque, commanded q-axis stator current and developed torque*

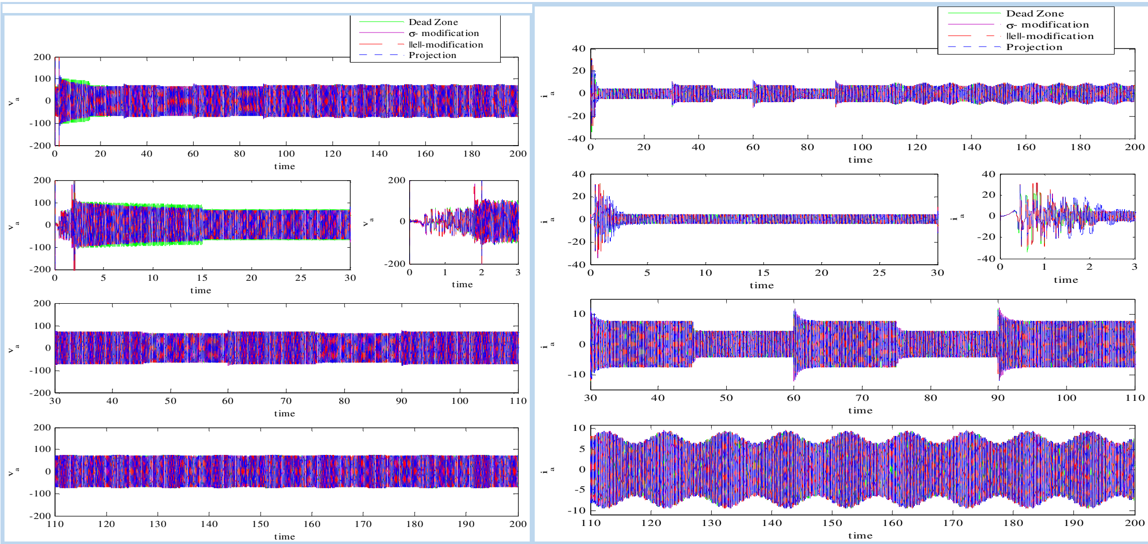
Figure (17) shows the supplied voltage and current of phase *a* inverter for all modifications and for all periods of load exertion. Figure (18) shows the zoomed versions current and voltage for all periods of load change (no load torque, step load change and sinusoidally load change). It is seen that all modifications have little effect on inverter current. However, it has been shown from the zoomed figures that *Projection Operator* modification gives the least current than the other techniques in case of no load period and in case of sinusoidal load period. On other hand, Sigma modification gives least current than other schemes in case of step change period.

5. Conclusion

This paper has studied the problem of robustness degradation due to presence of disturbance and uncertainty in speed control of IFOC three phase I.M. based on Lyapunov direct MRAC. Four modification schemes are suggested for enhancing the robustness; *Dead Zone* modification, σ-modification, e-modification and Projection Operator modification. Based on observations of simulated results, the following conclusions can be drawn;

- ❒ Due to the presence of uncertainty, uniform ultimate boundedness may not be guaranteed and may lead to drift of adaptive gains which in turn lead to instability problems and lack of robustness.
- ❒ Modification techniques work together with adaptive law to confine the adaptive gains within prescribed or undefined region and hence they strengthen the robustness of adaptive controller.

- ❒ All modification techniques could successfully enhanced the robustness due to uncertainty and disturbance change such that they all could confine the adaptive gains within specified bound.
- ❒ The results have showed that the effect of all modification techniques on motor state transients is considerably small; and in steady state the modification success to enhance the robustness of the controller.



(a) Phase a inverter Supplied voltage in (Volts) (b) Phase a inverter Supplied current in (Amperes)

Figure (17) *all modifications comparison*

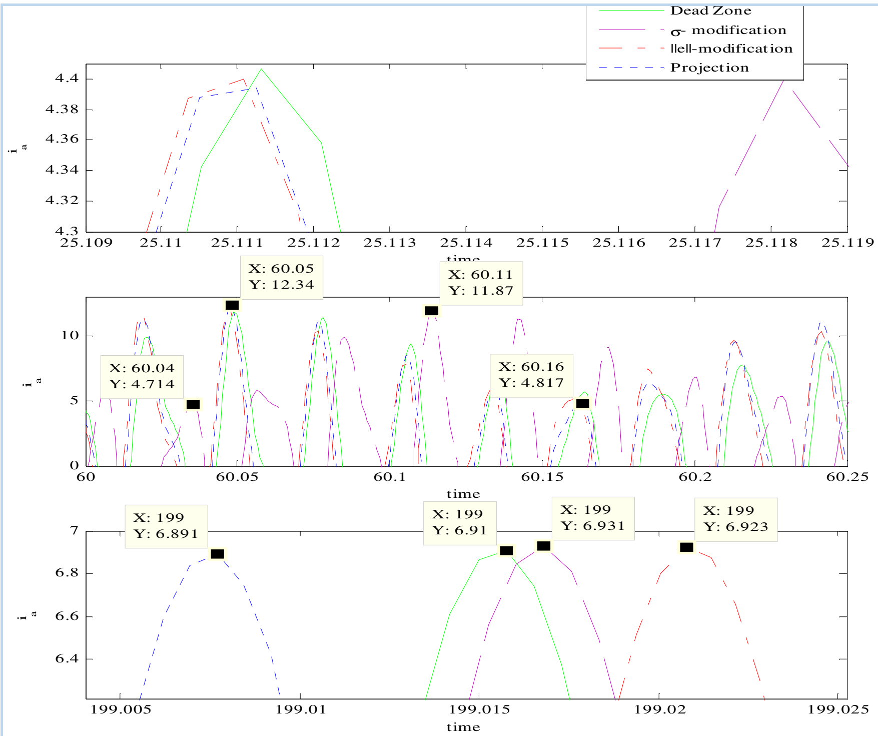


Figure (18) *Supply current of phase* ***a*** *inverter in (Amperes) for all modifications further zoomed*

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