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C O N T E N T

SPEED CONTROL OF AC MOTORS FOR ELECTRIC VEHICLES USING FIELD ORIENTED CONTROL

TOSHE VELKOV AND VLATKO CHINGOSKI

Abstract. The paper describes the speed regulation of electric vehicles driven by induction motors with the help of a vector control scheme which emulates a DC motor with independent excitation. The proposed speed regulation provides constant torque and constant speed of electric motors at different loads. Since vector control method needs the position of the rotor at any time, there are two types of speed regulation, direct - by measuring the angle of the rotor with the help of a hall sensor, and indirect - by mathematical determination of the rotor position. There will also be a simulation showing that in order to obtain a more adverse outcome, we must have a properly tuned PI regulator, which would otherwise result in delayed response and oscillations of the controlled driving system.

Keywords. Electric Vehicle, AC Motor, Transformations, Torque, Speed, PI Regulator

1. Introduction

The field of controlled electric drives has undergone rapid expansion over the last few years, mainly due to the advantages of semiconductors in both industrial and signal electronics and culminating in microelectronic microprocessors [1]. These technological improvements have enabled the development of a truly efficient AC drive control with less power dissipation and even more precise control structures. Power drive controls are becoming more accurate in the sense that not only are DC current and voltage controlled, but three-phase currents and voltages are also controlled by the so-called power supply vector controls [2].

This paper describes a very efficient form of a vector control scheme based on the field-oriented speed and torque control of AC motors used widely in modern electric vehicles. The proposed scheme is based on these main points: the transformation of three-phase current or voltage and time-dependent system into two coordinate time-invariant systems, decoupling of torque and flux, tuning of the PI controller, and effective generation of broadband impulse modulation. Thanks to these factors, AC control of the AC machine gains every advantage over DC control of the DC machine and gets rid of mechanical offsets.

Two field orientation schemes that require PI controllers have already been presented [3]. In general, the variables to be controlled are speed, torque, flux, voltage, and current [4]. The implementation of the direct type of field-oriented control has been regarded as being difficult in practice by virtue of the sensors needed for the control, such as search coils or Hall-effect sensors. These sensors, besides contributing a considerable amount to

the total cost of the controller and affecting reliability, often impose some additional control problems [7] – [12]. In a feedforward field orientation drive system, a slip calculator is used to coordinate the values of the magnetic flux, torque, and slip commands. In particular, the indirect vector control (sometimes called decoupling control or field-oriented control) system is the most practical system because of its various merits such as no required flux sensor and higher reliability of flux measurement, etc. Such systems, however, have the disadvantage of requiring a shaft angle sensor [13] – [15]. Various pulse wide modulation (PWM) technologies have also been developed and implemented in practical inverters. Some of the most popular methods are based on the concept of space vectors of the inverter voltages [16] – [18], and the utilization of the so-called voltage source inverters (VSI) that use sinusoidal band hysteresis current controllers [19], [20]. Regardless of which control method is used, using adequate tunning of the PI controller results with a very accurate steady-state and transient control that leads to high dynamic performance of the controlled AC motors in terms of timing response and power conversion.

2. Principles of the field-oriented control

The main purpose of the field-oriented control (FOC) is to keep the stator field perpendicular to the rotor field so that it can always produce maximum torque as with DC motors. For the control of the speed and torque of induction motors (IM) and/or permanent magnet synchronous motors (PMSM), the only approach is through phase voltages and currents. Figure 1 shows a simplified block scheme for general field-oriented control of IM and PMSM using mathematical transformations, such as forward and inverse Clarke and Park transformations.
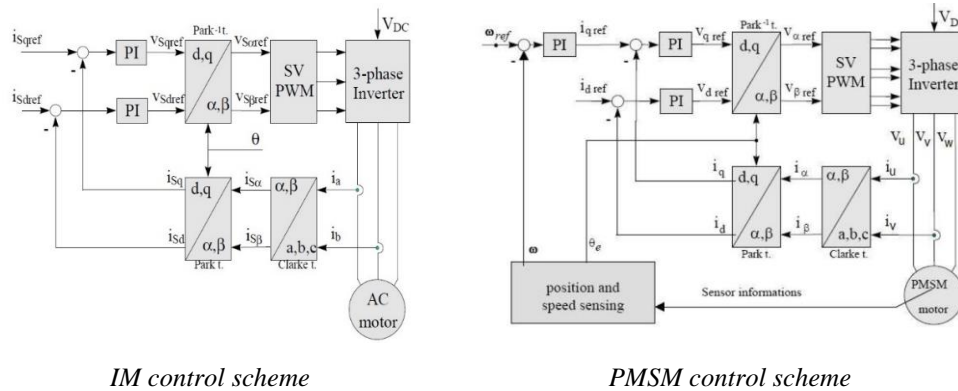
IM control scheme   PMSM control scheme

Figure 1. *The basic scheme of the induction motor with vector control*

The measurements of the feed currents i_a and i_b of the IM, or i_u and i_v for the PMSM, respectively, using the Clarke transformation module, results with output currents so-called designated currents $i_{S\alpha}$ и $i_{S\beta}$, or i_α и i_β, respectively [1, 2].

These two components of the current than become the inputs of the Park transformation that gives the current in the $d - q$ rotating reference frame. The i_{Sd} and i_{Sq} components are compared to the references i_{Sdref} (the flux reference) and i_{Sqref} (the torque reference). At this point, this control structure shows an interesting advantage: it

can be used to control either synchronous or induction machines by simply changing the flux reference and obtaining a rotor flux position. As in synchronous permanent magnet motors, the rotor flux is fixed (determined by the magnets), there is no need to create one. Hence, when controlling a PMSM, i_{Sdref} should be set to zero, as shown in Figure 1. As induction motors need a rotor flux creation in order to operate, the flux reference must not be zero. This easily solves one of the major drawbacks of the *"classic"* control structures: the portability from asynchronous to synchronous drives. The torque command i_{Sqref} could be the output of the speed regulator when we use a speed FOC.

The outputs of the current regulators are v_{Sdref} and v_{Sqref}; they are applied to the inverse Park transformation. The outputs of this projection are $v_{S\alpha ref}$ and $v_{S\beta ref}$, which are the components of the stator vector voltage in the α, β stationary orthogonal reference frame. These are the inputs of the Space Vector Pulse-Wide Modulator (SV PWM) as shown in Figure 1. The outputs of this block are the signals that drive the inverter. Note that both Park and inverse Park transformations need rotor flux position. Obtaining this rotor flux position depends on the AC machine type (synchronous or asynchronous machine) [1], [2]. In general, accurate control of instantaneous torque produced by a motor is required in high-performance drive systems, such as electrical vehicles (EV) and hybrid electrical vehicles (HEV) propulsions. The torque developed in the motor is a result of the interaction between the current in the armature winding and the magnetic field produced in the stator field of the motor. The field should be maintained at a certain optimal level, sufficiently high to yield a high torque per unit, but not too high to result in excessive saturation of the magnetic circuit of the motor [5]. With a fixed field, the torque is proportional to the armature current.

However, independent control of the field and armature currents is highly desirable. In a similar manner to that of a DC motor, the armature winding in IMs is also on the rotor, while the field is generated by currents in the stator winding. However, the rotor current is not directly derived from an external source but it results from the EMF induced in the winding as a result of the relative motion of the rotor conductors with respect to the stator field. In the most used squirrel-cage motors, only the stator current can be directly controlled since the rotor winding is not accessible. Optimal torque production conditions are not inherent due to the absence of a fixed physical disposition between the stator and rotor fields, and the torque equation is nonlinear. In such cases, the FOC can realize the optimal control for transient operation of an induction drive. The FOC can decouple the field control from the torque control [3]. A field-oriented IM emulates a separately excited DC motor in two aspects:

- the magnetic field and the motor torque can be controlled independently, and
- the optimum torque output conditions, resulting in maximum torque per unit, occurring at the motor in both steady-state and transient operating conditions.

In such matters, the field-oriented control of an IM emulates the optimal torque production conditions that are inherently satisfied in a DC motor (see Figure 2). The armature current i_a supplied through brushes is always orthogonal to the flux vector (field flux), λ_f, produced in the stator, and linking the rotor winding. Thus, the developed torque, T, is proportional both to the armature current and the field flux, that is,

$$T = K_T i_a \lambda_f \tag{1}$$

where K_T is a constant depending on the physical parameters of the motor. Therefore, the torque of the separately excited DC motors can be controlled by independently controlling the armature current and flux as previously mentioned [3], [4].
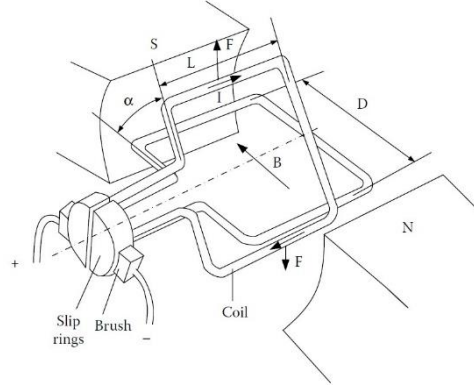

Figure 2. *Principle operational scheme of a DC electric motor*

3. Field-Oriented control scheme

As demonstrated in the above section, the field orientation principle defines the conditions of optimal torque production. The orthogonality of the rotor current and stator flux vectors must always be maintained. This is inherently satisfied in the steady state when the rotor settles down to such a speed that the developed torque matches the load torque. Under transient conditions, however, in order to meet the field orientation principal conditions, special techniques are required to provide an algorithmic equivalent of the actual physical disposition between the stator and rotor fields of the emulated DC motor.

The general block diagram of a vector control system for an IM drive is shown in Figure 3. A field orientation system produces reference signals, i_{as}^*, i_{bs}^* and i_{cs}^* of the stator currents, based on the input reference values λ_r^* and T^* of the rotor flux and motor torque, respectively, and the signals corresponding to selected variables of the motor. An inverter supplies the motor currents, i_{as}, i_{bs} and i_{cs}, such that their waveforms follow the reference waveform, i_{as}^*, i_{bs}^*, and i_{cs}^* [4].

Initially, it is not indicated which vector the excitation reference frame $D - Q$ aligns with. Clearly, anyone of the vectors can be used as a reference with which the excitation frame is to be aligned. In practice, usually it is the rotor flux vector, λ_r^s, along which the excitation frame is orientated. This method is referred to as the rotor flux orientation scheme, and it is shown in Figure 4 [6]. If the angular position of the rotor flux vector in the stator reference frame is denoted by θ_r, the $D - Q$ to $d - q$ transformation of currents and voltages in the described scheme could be expressed as:

$$\begin{bmatrix} i_{ds}^{s*} \\ i_{qs}^{s*} \end{bmatrix} = \begin{bmatrix} cos(\theta_r) & -sin(\theta_r) \\ sin(\theta_r) & cos(\theta_r) \end{bmatrix} \begin{bmatrix} i_{DS}^{e*} \\ i_{QS}^{e*} \end{bmatrix} \tag{2}$$

$$\begin{bmatrix} v_{DS}^e \\ v_{QS}^e \end{bmatrix} = \begin{bmatrix} cos(\omega t) & sin(\omega t) \\ -sin(\omega t) & cos(\omega t) \end{bmatrix} \begin{bmatrix} v_{ds}^s \\ v_{qs}^s \end{bmatrix} \tag{3}$$

It can be observed that this orientation of the orientation frame inherently satisfies the field orientation principal condition. The rotor flux is controlled by adjusting the i_{DS}^e component of the stator current vector—independently of the torque control, which is realized only by means of the i_{QS}^e component. The only requirement for this scheme is the accurate identification of angle θ_r, that is, the position of λ_r^s. This can be done in either a direct or indirect way [5], [6].
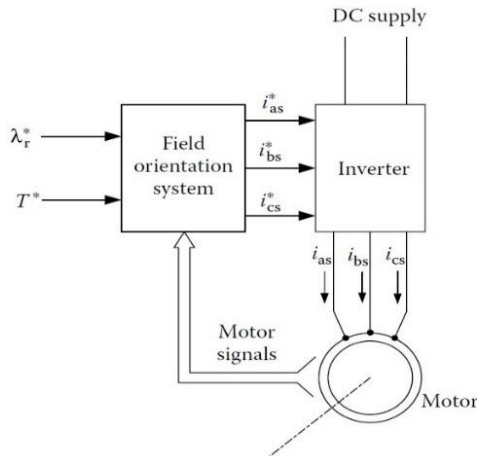
Figure 3. *General block diagram of a vector control system for an induction motor*
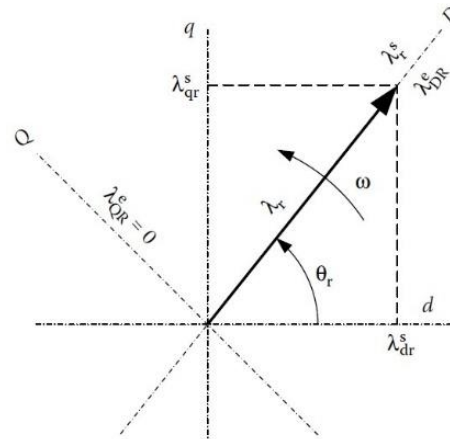
Figure 4. *The orientation of the excitation reference frame along the rotor flux vector*

In a field orientated IM, the i_{DS}^e and i_{QS}^e components of the stator current vector i_S^e in the excitation frame can be used for the independent control of the motor field and torque, respectively. Hence, the field orientation system as shown in Figure 3, first converts λ_r^* and T^* into the corresponding reference signals, i_{DS}^e and i_{QS}^e, of the vector of the stator current and then transforms them into the reference signals, i_{as}^*, i_{bs}^* and i_{cs}^*, of the stator phase current, which are to be produced by the inverter. The stator phase currents, i_{as}^*, i_{bs}^* and i_{cs}^*, can be calculated using $d - q$ to a-b-c transformation if the corresponding reference signals, i_{ds}^{s*} and i_{qs}^{s*}, in the stator reference frame are known. This is a simple scalar or static transformation since the elements of the transformation matrix used to perform this operation are constant [2].

3.1. Direct flux control

In direct field orientation systems, the magnitude and the angular position (phase) of the reference flux vector, λ_r^e are either measured or estimated from the stator voltage and current using flux observers. For example, Hall sensors can be used to measure magnetic

fields. Placing the sensors in the air gap of the motor, on the d and q axes, allows the determination of the corresponding components of the vector λ_m^s of the mutual flux (air gap flux). However, this air gap flux differs from the rotor flux, which is taken as the reference flux vector [7].

Since λ_r^s differs from the λ_m^s by only the leakage flux in the rotor, then:

$$\lambda_r^s = \lambda_m^s + L_{1r}i_r^s = \frac{L_r}{L_m}\lambda_m^s - L_{1r}i_s^s \tag{4}$$

A microprocessor-based rotor flux calculator is shown in Figure 5. It performs algebraic operations as follows:

- signals i_{ds}^s and i_{qs}^s are calculated from the actual stator phase currents, i_{as}, i_{bs} and i_{cs}, using the a-b-c to $d-q$ static transformation, following by
- calculating signals λ_{dr}^s and λ_{qr}^s using (2), [8],

$$\lambda_{dr}^s = \frac{L_r}{L_m}\lambda_{dm}^s - L_{1r}i_{ds}^s \tag{5}$$

$$\lambda_{qr}^s = \frac{L_r}{L_m}\lambda_{qm}^s - L_{1r}i_{qs}^s \tag{6}$$

- finally, the magnitude λ_r and the phase θ_r of the rotor flux vector are determined using the rectangular to polar coordinate transformation.

$$\lambda_{qr}^s = \lambda_{dr}^s + j\lambda_{qr}^s \rightarrow \lambda_r\angle\theta_r \tag{7}$$

It must be pointed out that the orthogonal spacing of the flux sensors in Figure 5 applies only to two-pole machines. In a p-pole machine, the sensors must be placed $180/p$ from each other [9] – [11].
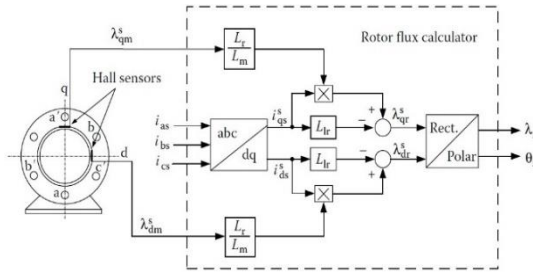
Figure 5. Determination of the magnitude and position of the rotor flux vector using a Hall sensor and a rotor flux calculator

Figure 6. Torque calculator

Since $\lambda_{DR}^s = \lambda_r$, then the output variable λ_r of the rotor flux calculator can be used as a feedback signal in the field control loop. The same variables can also be used for the calculation of the developed torque as shown in Figure 6. The torque calculator computes torque in the following steps:

- the static a-b-c to d-q transformation is performed on the stator currents i_{as}, i_{bs} and i_{cs} to obtain i_{ds}^s and i_{qs}^s;

- the angle θ_r supplied by the rotor flux calculator is substituted into (3) for ωt in order to transfer signals i_{ds}^s and i_{qs}^s into i_{DS}^e and i_{QS}^e components of the stator current vector in the excitation frame;
- the magnitude λ_r of the rotor flux, also supplied by the rotor flux calculator and presumed equal to λ_{DR}^e, is multiplied by i_{QS}^e and by the torque constant K_T to calculate the developed torque.

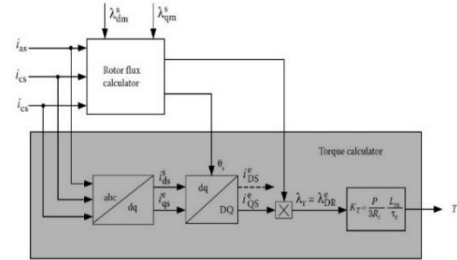Figure 7 illustrates an independent flux and torque control block diagram, based on the vector control of an IM with direct rotor flux orientation. In the system, proportional-plus-integral (PI)-based field and torque controllers are used to generate the control signals i_{DS}^{e*} and i_{QS}^{e*} in the excitation frame by comparing the target rotor flux λ_r^*, and the target torque T^*, with the actual rotor flux λ_r, and torque T. Then, i_{DS}^{e*} and i_{QS}^{e*} from the excitation frame are transferred into i_{ds}^{s*} and i_{qs}^{s*} of the stator reference frame using a rotor flux angle. Furthermore, i_{ds}^{s*} and i_{qs}^{s*} in the stator reference frame are transferred into the phase current signals i_{as}^*, i_{bs}^*, and i_{cs}^* through static transformation.

The phase current signals, as the reference signals, are used to control the power electronics of the inverter to generate the corresponding phase current i_{as}, i_{bs} and i_{cs} [10]. In practice, the ratio of L_r to L_m, and the rotor leakage inductance, L_{1r}, which are used in the rotor flux calculator (see Figure 5), are not significantly affected by changes in the operating conditions of the motor, such as the winding temperature or saturation of the magnetic circuit [8]. Therefore, the described field orientation techniques are the most robust and accurate. However, it requires the placement of vulnerable Hall sensors in the motor's air gap, to the detriment of the cost and reliability of the drive system [9], [11].
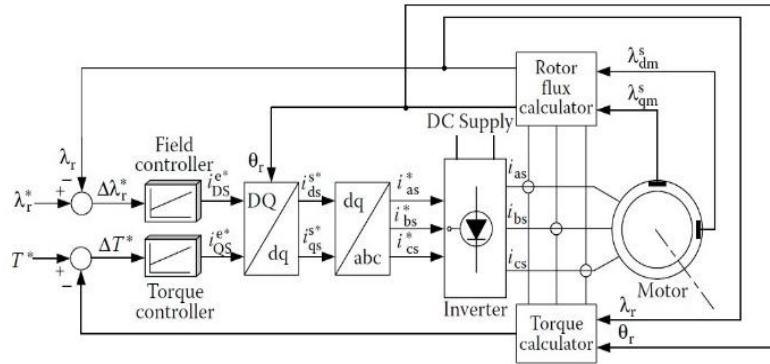
Figure 7. Vector control system for an IM with direct rotor flux orientation

3.2. Indirect flux control

The presence of vulnerable Hall sensors in vector control with direct rotor flux orientation would weaken the reliability and enhance the cost of the motor drive. The indirect approach is to obtain the rotor flux position by the calculation of the slip speed, ω_r, required for correct field orientation, and the imposition of the speed on the motor.

If the synchronous speed necessary to maintain the orthogonal orientation of vectors λ_R^e and i_R^e in the given operating conditions of the motor is denoted by ω^*, the θ_r the angle can be expressed as

$$\theta_r = \int_0^t \omega^* dt = \int_0^t \omega_r^* dt + \int_0^t \omega_0 dt = \int_0^t \omega_r^* dt + \theta_0 \tag{8}$$

where ω^*, ω_r^* and ω_0 are synchronous speed, slip speed and rotor speed respectively, and θ_0 is the angular displacement of the rotor, which is easy to measure using a position sensor [12].

The rotor flux and the developed torque are controlled without the need for feedback. Consequently, the performance of the system strongly depends on the accurate knowledge of the motor parameters, a condition that can be difficult to satisfy in practical applications [13]. The rotor timing τ_r varies substantially during the operation of the IM, generally due to changes in rotor resistance to temperature, the frequency-dependent skin effect on the rotor coils, and the effect of the impeller saturation on the inductor. On the other hand, the great advantage of such a system is that if a standard IM is used, the position of the rotor could be easily measured or estimated by an external sensor [14], [15].
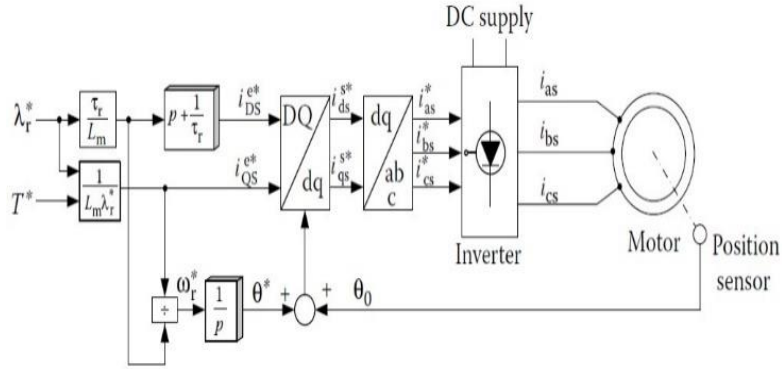
Figure 8. Vector control system for an IM with indirect rotor flux orientation

3.3. PI Controller

An electrical drive based on the FOC needs two constants as control parameters: the torque component reference i_{Sqref} and the flux component reference i_{Sdref}. The classic numerical proportional-and-integral (PI) regulator is well suited for regulating the torque and flux feedback to the desired values as it is able to reach constant references, by correctly setting both, the P term (K_{pi}), and the I term (K_i)) which are respectively responsible for the error sensibility and for the steady-state error [1]. The numerical expression of the PI regulator is as follows:

$$U_k = K_{pi}e_k + K_i e_k + \sum_{n=0}^{k-1} e_n \tag{9}$$

which can be represented by Figure 9:

Figure 9. Classical numerical PI regulator structure

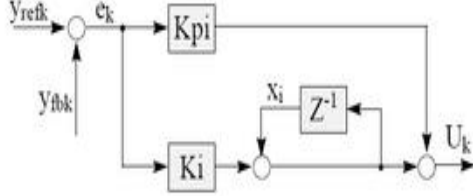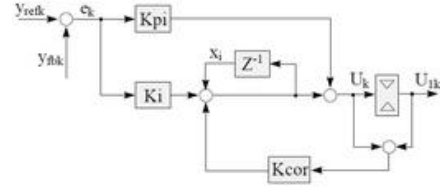Figure 10. Numerical PI regulator with the correction of the integral component

According to (9), the limiting point for the control scheme presented in Figure 9 is that during normal operation, or during tests, large reference value variations or large disturbances may occur, resulting in saturation and overflow of the regulator variables and output. If they are not controlled, this kind of non-linearity damages the dynamic performance of the system. To solve this problem, one solution is to add to the previous structure a correction of the integral component as depicted in Figure 10. The integral term correction algorithm in a high-level language is given below:

Input y_{refk}, y_{fbk}

$$e_k = y_{refk} - y_{fbk} \tag{10}$$

$$u_k = x_i - K_{pi}e_k \tag{11}$$

$$u_{lk} = u_k \tag{12}$$

If $u_k > u_{max}$ then $u_{lk} = u_{max}$
If $u_k < u_{min}$ then $u_{lk} = u_{min}$

Output u_{lk}

$$e_{lk} = u_k - u_{lk} \tag{13}$$

$$x_i = x_i - K_i e_k + K_{cor}e_{lk} \tag{14}$$

where u_{max}, u_{min} are the limitations of the output variables.

4. Results and Discussion

Figure 11 shows the results of an IM simulation with direct rotor flux control in a case when the PI controller is not tuned properly. Before adjusting the PI controller, it is easy to see that at first, we have a high starting torque and a slight slowdown in speed rise from

t=0 to t=0.2 s, which takes time to reach the desired speed and torque. Figure 12 shows the difference between the targeted motor speed given in yellow, and the actual motor speed given in blue color. It is evident that using an untuned PI controller we have oscillation in response, and thus a slow approximation to the desired speed value. Since it is an electric vehicle that carries a heavy load when traveling at low speeds, these oscillations could be very sensitive.
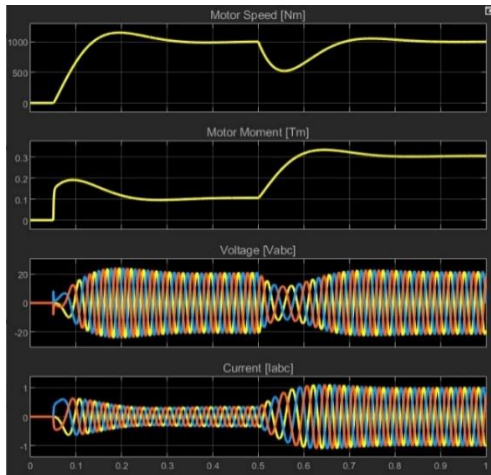

Figure 11. Time diagram of motor speed, moment, current and voltage before tuning the PI controller
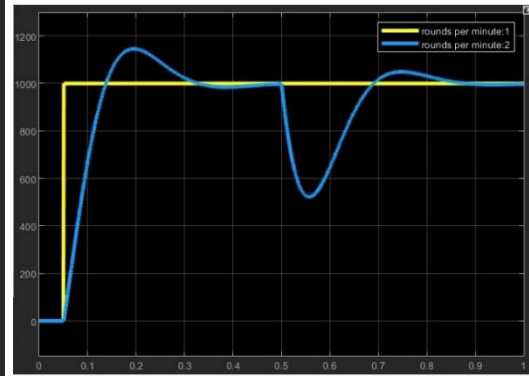
Figure 12. Time diagram of actual and targeted motor speed before tuning the PI controller

Contrary, Figures 13 and 14 show a simulation of an electric vehicle where the PI controller is correctly tuned. With a properly tuned PI controller, we see that the desired speed is reached quickly and there is no deviation. However, a small speed disturbance occurs at t=0.5s. This disturbance occurs when the vehicle starts to move at a certain speed which means that at that time, the motor momentum and speed start to rise sharply and tend to accelerate and reach the default value, this to this end, the engine current is increasing. It should be noted that in this electric vehicle the tuning of the PI controller is automatic and is based on the mathematical model of the system.

5. Conclusions

Based on what has been said so far, we can draw several conclusions. At the beginning of this work, we have seen how transformations are performed and how those transformations allow for the linearization of engine parameters and easier motor exploitation and control. We can conclude that this is an electric drive at which the moment is constant, and therefore the power is variable. We have seen two types of direct and indirect rotor flux control and their preconditions. Therefore we can conclude that the direct control of the rotor flux requires Hall sensors which reduces the reliability of the system and therefore nowadays the indirect vector control technique replaces the direct vector control and is used more.

We have also seen that the PI controller must be properly tuned at all times so that we do not deviate from the desired values. All this tells us that, when powered at low speeds and high loads, the most appropriate choice of control is vector control, but this type of control is also a major challenge as it is quite delicate to perform. From the above discussion, it can be concluded that IM control is essential. Therefore, a well-established motor drive that is simple, solid, low cost and with low maintenance needs could well serve the required purpose in various electric drives including modern electric vehicles.
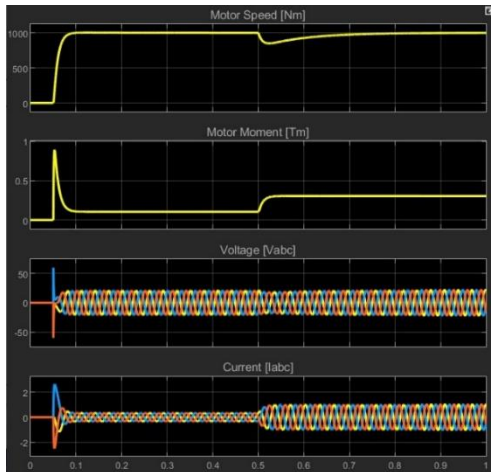

Figure 13. Time diagram of motor speed, moment, current and voltage after tuning the PI controller
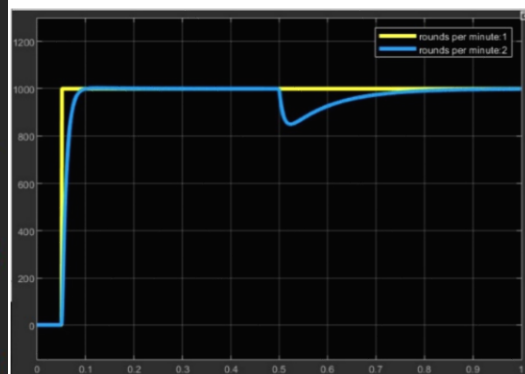
Figure 14. Time diagram of actual and targeted motor speed after tuning the PI controller

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