Simplified Predictive Stator Current Phase Angle Control of Induction Motor With a Reference Manipulation Technique

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ABSTRACT Finite control set model predictive control (FCS-MPC) is a simple method and has an appropriate dynamic response for the drive applications. Applying additional control objects, e.g., the maximum torque per ampere (MTPA), is easy in FCS-MPC because of its characteristics. A direct application of FCS-MPC to MTPA is the predictive direct angle control method. Though this method eased the MTPA process, the good result is highly sensitive to the proper selection of the weighting factor. Furthermore, finding the best phase angle needs a complicated optimization process. In this paper, the application of simplified predictive control is proposed for angle control. By this proposed method, not only the weighting factor is eliminated but also the constraints of the motor are considered in the control strategy. In this way, the phase angle is automatically controlled in the proper value due to the torque while tedious computation is avoided. Therefore, the proposed method is valid in a wide range of operating points while no optimization process is performed due to changes in speed and torque. This proposed method is evaluated by simulations and experiments.

INDEX TERMS Motor drives, predictive control, torque control.

I. INTRODUCTION

Finite control set model predictive control has a lot of progress in different fields, and because of the simplicity and good dynamic response, there is a lot of interest in studying it [1], [1]–[3]. The performance of this method is based on minimizing the cost function and considering the discrete nature of the inverter to select the most appropriate voltage vector [4]–[8]. Good dynamic response and controllable switching losses are the advantages of this method over the vector control [9]. One of the features of this method is the capability of considering multi-objective cost functions. One of these objectives that can be considered in the cost function is the maximum torque per ampere (MTPA) criterion [10].

The MTPA is a method to reduce the copper losses. It can increase efficiency especially when the copper loss predominates [11]. In this method, the torque is obtained by a vector of current that is the minimum current vectors [12].

In [10], [13]–[15], the FCS-MPC method is applied to the MTPA in different ways. The main differences between these methods are the type and the number of weighting factors.

In [13], the MTPA procedure is considered for calculations of the references. The reference are calculated based on the operating region. The operating region is divided to the flux-increased and flux-limited regions. The MTPA is performed by considering these regions. The transition between those regions are also covered. In this method, the weighting factor in the cost function is a weighting matrix. This weighting matrix is calculated from the optimization techniques and the inequality of linear matrices. The complexity of the state reference calculation and weighting factor tuning are the drawbacks of this method.

In [14]–[16], the cost function is multi-objective. One of these objectives is dedicated to MTPA control. This
method avoid the problem of lack of coordination in cascaded form of MTPA and predictive control. In [14], [16], three objectives form the cost function. The first part is the torque control. Instead of flux control, the second objective is the MTPA criterion. The third part is a limitation which is applied to avoid wrong convergence. In [15], the first objective is current control and the second one is the current magnitude. This is direct application of MTPA which needs less computation. The drawback of this method is coordination of several weighting factors in the cost function.

To overcome the problem of several weighting factors, a cost function with less number of objects was introduced as predictive direct angle control in [10]. Compared to methods in [13]–[15], predictive direct angle control is the most compact combination between MTPA and FCS-MPC. In this method, the cost function has two objectives. The first is the tracking errors of the torque and the second is the error of the current angle. In each period, to minimize the cost function, the torque and the current angle are predicted for all voltage vectors that can be applied to the inverter. The reference angle of the current in the cost function is equal to the MTPA angle. The MTPA angle for the induction motor is 45° [17]. Though this method is much easier compared to several objective methods, the disadvantages of this method is still a needed weighting factor. The presence of even one weighting factor can affect the quality of the results because it may be dependent on the operating point and motor parameters. Furthermore, the phase angle equal to 45° is not the proper reference when the electromagnetic torque reaches to the flux limitation borders of the motor. These disadvantages are due to the inherent problems of the multi-objective FCS-MPC method.

Lately, the simplified FCS-MPC method is introduced, which is a new form of the FCS-MPC method and reduces the problems of the multi-objective FCS-MPC method. The cost function in this method is single-objective, and the weighting factor is eliminated. This advantage is achieved at the cost of the lack of other constrains in the cost function. Two different types of this method have been proposed to date, i.e., flux vector based cost function, and voltage vector based cost function.

In [18]–[20], the torque is controlled through the flux angle. Therefore, the tracking error in the cost function is just the flux vector. In these methods, the flux vector must be predicted for all applicable voltage vectors. These values are compared to the calculated flux vector reference via a single-objective cost function. The flux vector reference controls the torque automatically. The amount of computation is still high in these methods.

In [1], [21], [22], the methods are based on the dead-beat control. In these methods, one reference voltage vector is calculated to control the torque and the flux magnitude in each interval. The voltage vector reference is directly examined in the cost function. This feature reduces the computation significantly.

Therefore, the simplified FCS-MPC based on dead-beat control has received more attention due to its simplicity and less computation. On the other hand, to increase the features of the simplified FCS-MPC method, criteria such as the MTPA method can be added. In [23], the simplified FCS-MPC with the MTPA method is proposed. In this method, the MTPA trajectory is used to calculate the optimum flux magnitude reference. Although the weight factor has been eliminated, the computation is still complicated because of the MTPA process for the optimum flux calculation and load angle prediction.

Therefore, there is a research gap for applying the MTPA technique in predictive control. The simplified predictive control needs flux optimization and the predictive angle control needs weighting factor optimization. Furthermore, the proper phase angle is not considered for all operating points. In this research, the voltage reference of the simplified predictive control is calculated to automatically control the optimum phase angle. The prominent feature of the proposed method is the elimination of the weighting factor from the cost function which eliminates the need for tedious pre-simulations for tuning the weighting factor. Also, automatically manipulation of the phase angle reference from 45° is achieved by considering the flux limitation based on the electromagnetic torque. Thus, there is no need to calculate the optimal flux or optimal phase angle.

II. PREDICTIVE DIRECT ANGLE CONTROL

In this method, the MTPA is applied to the FCS-MPC with this property that the flux is controlled indirectly. The cost function of this method consists of the errors of the torque and the angle of the current.

In this method, the flux is controlled automatically through the control of the current angle. The current angle in rotor flux oriented frame is controlled at the reference which will minimize the current magnitude [10]. Thus, the MTPA goal is fulfilled.

In the induction motor, the minimum current magnitude is obtained when the angle between the rotor flux vector and the stator current vector is equal to 45° [17]. In this way, this angle is selected as the reference of the current angle in the cost function, which causes the current magnitude to be minimized in each period. To perform this idea, the cost function is as follows:

\[ C_i = |T_{em}^* - T_{em,i,k+1}| + Q |\alpha_{r,i,k+1} - \pi/4| \quad i = 1, 2, ..., 7 \]  

where \( T_{em}^* \) is the reference of the torque, \( T_{i,k+1} \) and \( \alpha_{r,i,k+1} \) are the predicted torque and the phase angle of the current, respectively, \( i \) is the index of seven applicable voltage vectors to the inverter. \( Q \) is the weighting factor that is used to coordinate the current angle error and the torque error.

In [10], \( T_{em,i,k+1} \) and \( \alpha_{r,i,k+1} \) are expressed as follows:

\[ T_{em,k+1} = \frac{3}{2} \frac{L_m}{L_r} |\vec{i}_{r,k+1}| |\vec{i}_{s,k+1}| \sin \alpha_{r,k+1} \]  
\[ \alpha_{r,k+1} = \angle \vec{i}_{s,k+1} - \angle \vec{i}_{r,k+1} \]
where \( p \) is the number of the pole pair, \( L_m \) and \( L_s \) are the mutual and rotor inductance, respectively, \( \lambda_{r,k+1}, \vec{I}_{s,k+1} \) are the predicted rotor flux and stator current vectors in instant \( k+1 \), respectively. Note that (1) and (2) are valid in any reference frame and any time instant. They are used at \( k+1 \) time instant in order to form the prediction model.

In the (2) and (3), \( \vec{I}_{s,k+1} \) and \( \vec{\lambda}_{r,k+1} \) are calculated as follows:

\[
\vec{I}_{s,k+1} = \frac{1}{\sigma L_s} \vec{\lambda}_{s,k+1} - \frac{1}{\sigma L_s} \left( 1 - \frac{T_s}{\tau_r} + j\omega_r T_s \right) \vec{\lambda}_{s,k} + \left( 1 - \frac{T_s}{\tau_r} + j\omega_r T_s \right) \vec{I}_{s,k} \tag{4}
\]

\[
\vec{\lambda}_{r,k+1} = \frac{L_r}{L_m} \left( 1 - \frac{T_s}{\tau_r} + j\omega_r T_s \right) \vec{\lambda}_{s,k} + \left( \sigma L_m (1 + j\omega_r T_s) - \frac{L_m L_r}{L_m} \frac{T_s}{\tau_r} \right) \vec{I}_{s,k} \tag{5}
\]

where \( \sigma = 1 - L_m^2/L_s L_r \), \( \vec{I}_{s,k} \) and \( \vec{\lambda}_{s,k} \) are the estimated current and flux stator, respectively. \( L_s \) is the stator inductance, \( T_s \) is the sampling interval, \( \tau_r \) is the rotor time constant, and \( \omega_r \) is the rotor angular frequency. Note that the proof is presented in the appendix.

Though the cost function is simpler than that of the previous methods with several objectives due to the MTPA, tuning even one weighting factor needs pre-simulations. Furthermore, tuning the weighting factor to coordinate these two particular objectives is harder because both of the angle and torque are quick variable while the flux is slower in conventional FCS-MPC. On the other hand, calculating the phase angle difference for seven vectors increases the needed computational power. Also, in some cases phase angle equal to 45° is not the best solution, e.g., in very low torque condition the magnetizing component of the current will be smaller than the needed current, and in rated torque condition the core will be saturated. In this research, these drawbacks are tried to be improved by using the simplified predictive technique.

III. PROPOSED SIMPLIFIED PREDICTIVE DIRECT ANGLE CONTROL

A. CONCEPTION

The concept of controlling the torque and the current phase angle is applied similar to the basic direct angle control. But how the idea is applied is different for the proposed method. In the proposed method, the reference voltage vector that will fulfill the torque and phase angle control will be predicted in each interval. Then, the voltage-based single-objective cost function is used to choose the switching state. Therefore, the weighting factor is eliminated by this method. Furthermore, the prediction stage is not needed to be performed for seven times which reduces the needed computational power. The needed prediction stage will be the calculation of the voltage vector which produces the torque to the reference value and keeps the current phase angle at 45° in rotor flux oriented frame. By this method, the up and down limitation of the flux can be considered before the voltage calculation.

FIGURE 1. Vector diagram of induction motors.

B. MATHEMATICS FOR THE PROPOSED METHOD

Fig. 1 shows the vector diagram of the induction machine in the stationary frame (α-β) and the rotating frame (d-q). According to this figure and (2), the torque equation in the d-q frame can be rewritten as follows [10]:

\[
T_{em} = \frac{3}{2}\frac{L_m^2}{L_r} I_{sd} I_{sq} \tag{6}
\]

To consider the MTPA method in the induction motor, the following equation should be fulfilled. Note that the automatic flux limitations are not considered at this stage but they will be noticed in the next section.

\[
I_{sd} = I_{sq} \tag{7}
\]

To reach to the above equation, and express it based on the stator current and the current angle, the current angle must be equal to 45°. So:

\[
I_{sd} = I_s \sin 45 \tag{8}
\]
\[
I_{sq} = I_s \cos 45 \tag{9}
\]

By putting (8) and (9) in (6), the torque equation will as follows:

\[
T_{em} = \frac{3}{4} \frac{L_m^2}{L_r} I_s^2 \tag{10}
\]

On the other hand, according to flux current equation of the induction machine, the stator flux equation can be calculated in terms of stator current and rotor flux as follows:

\[
\lambda_s = \frac{L_s L_r - L_m^2}{L_r} I_s + \frac{L_m}{L_r} \lambda_r \tag{11}
\]

Then, by transferring (11) to the rotating frame it can be expressed in a separated form.

\[
\lambda_{sd} = L_d I_{sd} \tag{12}
\]
\[
\lambda_{sq} = L_q L_r - L_m I_{sq} \tag{13}
\]
C. CONTROL LOOP

By expressing (10) in the $k + 2$ instant and setting the torque reference equal to $T_{em,k+2}$, the predicted stator current can be calculated in this instant.

\[ I_{s,k+2} = \sqrt{\frac{4}{3} \frac{T_{em,k+2}}{P} \frac{L_r}{P_m^*}} = \sqrt{\frac{4}{3} \frac{T_{em}}{P} \frac{L_r}{P_m^*}} \]

(14)

where $T_{em}^*$ is the torque reference which is calculated by the proportional-integrator controller of the speed. That note that a current limiter should be used after this prediction. If the predicted current is higher than the maximum current of the motor it will set to the maximum value.

By considering (14), in (12) and (13), the stator flux for instant $k + 2$ is calculated.

\[ \lambda_{sd,k+2} = L_d L_{sd,k+2} \]

(15)

\[ \lambda_{sq,k+2} = \frac{L_d L_r - L_m^2}{L_r} I_{sq,k+2} \]

(16)

If the current conditions (8) and (9) are considered, and the predicted current phase angle is set to 45°, the predicted stator flux will be as follows:

\[ \lambda_{sd,k+2} = L_d L_{sd,k+2} \sin \alpha_{r,k+2} = \frac{\sqrt{2}}{2} L_d I_{s,k+2} \]

(17)

\[ \lambda_{sq,k+2} = \frac{L_d L_r - L_m^2}{L_r} I_{s,k+2} \cos \alpha_{r,k+2} = \frac{\sqrt{2}}{2} L_d L_r - L_m^2 \]

(18)

On the other hand, according to [18], the stator flux equation for instant $k + 1$ is calculated as follows:

\[ \ddot{\lambda}_{s,k+1} = \ddot{\lambda}_{s,k} + T_s \left( \dot{V}_{s,k} - R_s \dot{I}_{s,k} \right) \]

(19)

Since the final applied voltage by the inverter is in the stationary frame, the voltage reference should be expressed in the stationary frame. For this purpose, the discrete form of the first difference equation of the induction motor is applied.

\[ \frac{\ddot{\lambda}_{s,k+2} - \ddot{\lambda}_{s,k+1}}{T_s} = \dot{V}_{s,k+1} - R_s \dot{I}_{s,k+1} \]

(20)

In this equation, $\dot{\lambda}_{s,k+1}$ is calculated by (19) which is in the stationary frame, and $\ddot{\lambda}_{s,k+2}$ is calculated by (17) and (18) which should be transformed into the stationary frame as below:

\[ \frac{\ddot{\lambda}_{s,k+2} + j \lambda_{sq,k+2}}{T_s} = \dot{V}_{s,k+1} - R_s \dot{I}_{s,k+1} \]

(21)

where $\dot{\lambda}_{s,k+1}$ is the phase angle of the rotor flux. The rotor flux is predicted by (5).

Eventually, the decomposed form (21) is applied to predict the proper voltage references in the stationary frame.

\[ V_{a}^* = \frac{\dot{\lambda}_{sar,k+2} - \dot{\lambda}_{sar,k+1}}{T_s} + R_s I_{sar,k+1} \]

(22-a)

\[ V_{\beta}^* = \frac{\lambda_{sar,k+2} - \lambda_{sar,k+1}}{T_s} + R_s I_{sar,k+1} \]

(22-b)

Finally, in each period, the optimal voltage vector is selected by calculating the following cost function for seven possible switching states and selecting the minimum.

\[ G = \left| V_{a}^* - V_{i,a} \right|^2 + \left| V_{\beta}^* - V_{i,\beta} \right|^2 \]

(23)

To further understand how the proposed method works, the block diagram of the proposed method is shown in Fig. 2. In this figure the MTPA block is the application of (14) and
the flux prediction block is using (17) and (18). As it is shown in the block diagram, there is no need for flux optimization. Also, no direct flux control is needed. The flux will be regulated automatically by voltage reference prediction.

The general computational process is shown in the form of a flowchart in Fig 3.

In order to show the difference between the voltage selection of the proposed method and that of the previous angle control method [10], a numerical example is provided. Fig. 4 shows the numerical example in a case of voltage selection situation. In this figure, the vectors of stator and rotor fluxes, and also the current vector are depicted in αβ frame. In this frame, \( \hat{I}_{s,k+1} = 0.2289 + j0.9056 \), \( \hat{\lambda}_{s,k+1} = 0.7371 + j0.1539 \), \( \hat{\lambda}_{r,k+1} = 0.7269 + j0.0488 \). Note that normalized variables are provided. The current vector trajectory that satisfies the torque reference generation is shown in this figure.

In the previous method, all possible voltage vectors are used to predict the next current phase angle and the torque. The phase angle should be close to \( \pi/4 \) and the torque should be close to the reference. By considering Fig. 4, it is clear that the errors are approximately acceptable for two candidates, i.e., \( j = 1, 6 \). The error of the phase angle for \( j = 1 \) is 0.0526 and the torque error is 0.2672. Also, for \( j = 6 \), the error of phase angle is 0.1445 and the torque error is 0.2026. It is seen that the error are slightly different for these two options. Therefore, the weighting factor selection is essential in this case. For example if equal weighted cost function is used in (1), \( C_1 = 0.3198 \) and \( C_6 = 0.3472 \). So, \( j = 1 \) is selected. However, if the weighting factor \( Q = 2 \) is selected, \( C_1 = 0.587 \) and \( C_6 = 0.5498 \) and \( j = 6 \) is selected.

On the other hand, in the proposed method, the next flux reference is predicted, \( \hat{\lambda}_{s,k+2} = 0.7862 + j0.1028 \). The voltage vector that can result in the predicted flux reference is only \( j = 1 \) and \( j = 6 \) is not a close solution while there is no weighting factor to be tuned.

**D. REFERENCE MANIPULATION**

In the proposed method, the reference value for the phase angle of the stator current is set to 45°. There are some cases that this degree can result in improper performance. In the following, the cases that the reference manipulation is performed are explained.

1) VERY LIGHT TORQUE CONDITION

The reference 45° cannot be effective in very light torque condition because both current components will be close to zero and the magnetization of the machine would not be performed. To avoid this condition, the minimum value of the direct component of the flux in (17) is limited [24].

\[
\min (\hat{\lambda}_{s,k+2}) = \sqrt{\frac{8}{3p} \frac{\sigma L_r L_m^2}{L_m^2} T_{em}} \tag{24-a}
\]

\[
\min (T_{em}) = 0.05 T_{nom} \tag{24-b}
\]

where \( T_{nom} \) is the rated torque of the motor. Based on (24-a), there is a minimum value for the stator flux for every torque value. If the calculated value by (17) is lower than that the minimum value will be selected. Also, the minimum accepted torque and \( T_{nom} \) is 5% rated torque.

Therefore, in light load condition, the phase angle would be lower than \( \pi/4 \).
2) NEAR RATED TORQUE CONDITION

When the torque is close to the rated value, the phase angle 45° results in a big value for the direct component of the flux. The outcome of this effect is core saturation. In order to avoid that, the maximum value of the d-axis flux is limited.

\[
\max \left( \lambda_s, k+2 \right) = \frac{\sqrt{2}}{2} L_s I_{s, \text{nom}} \tag{25}
\]

where \( I_{s, \text{nom}} \) is the rated value of the magnetizing current which is considered as the maximum allowed value for the d-axis stator current.

Furthermore, the magnitude of the stator flux vector should also be limited in order to avoid the saturation [24].

\[
\max \left( \lambda_s, k+2 \right) = \max \left( \sqrt{\frac{\lambda_s^2, k+2 + \lambda_q^2, k+2}{2}} \right) = \max \left( \sqrt{L_s^2 \left( I_{s, \text{max}}^2 + \frac{I_{s, \text{max}}^4 - 64L_s^2 I_m^4 T_{e, \text{max}}^2}{9p I_m^4 T_{e, \text{max}}^2} \right) \omega_e} \right) \tag{26}
\]

where \( I_{s, \text{max}} \) and \( \omega_e \) are the maximum values for the stator current and voltage. These values are also tunable based on the limitation and needed response.

If the limit of (26) happens, the q-component of the stator flux is kept at the value calculated by (18) and its d-component will be set at the following value.

\[
\lambda_{sd, k+2} = \sqrt{\max(\lambda_s, k+2)^2 - \lambda_{sq, k+2}^2} \tag{27}
\]

In this situation, the phase angle is set to a value higher than 45° because the q-component is larger than the d-component. Fig. 5 summarizes the reference manipulation algorithm. In this algorithm, first the minimum torque is limited. Afterward, the predicted d-axis flux is checked. If it is lower than the minimum value, it is reserved at the minimum value. If the flux is higher than the maximum value it is limited at the maximum value. At the end, the magnitude of the stator flux is checked and limited in order to avoid core saturation.

IV. RESULTS AND DISCUSSION

The proposed method is evaluated by the simulations experiments. Table 1 shows the specifications of the motor which is used in laboratory tests. The model of that motor is used in the simulations. Also, a two-level inverter by PM25SR5120 package is used in the laboratory set. The inverter is controlled by the TMS320F28335 digital processor.

| Specifications |
|----------------|
| \( P_s = 1.5 \text{[KW]} \) | \( R_s = 5.2 \Omega \) |
| \( V_s = 220 \text{[V]} \) | \( R_s = 4.9 \Omega \) |
| \( I_s = 3.9 \text{[A]} \) | \( L_s = 632 \text{[mH]} \) |
| \( n_s = 1410 \text{[rpm]} \) | \( L_m = 591 \text{[mH]} \) |
| \( P = 2 \) | \( L_0 = 623 \text{[mH]} \) |
| \( T_s = 10.5 \text{[Nm]} \) | \( f_s = 50 \text{[Hz]} \) |

TABLE 1. Specification of induction motor.

To obtain the results, the following conditions are considered for simulation and laboratory testing: 1) The switching interval and sampling time are set to 100 \( \mu s \). 2) The switching operation is done in the middle of the sampling interval to reduce the noise and delay time. 3) The saturation is considered in the simulated model.

A. SIMULATION RESULTS

Simulations are performed to compare the proposed method with the previous direct angle control method [10]. The importance of the elimination of the weighting factor in the proposed method is studied. In other words, the effect of the weighting factor on the result is investigated. For this purpose, the direct angle method was simulated for four values of weighting factors. The study is repeated for three torque load conditions, i.e., 5%, 50%, and 100% rated torque. The responses are shown in Fig. 6, Fig. 7, and Fig. 8, respectively.

The result of the proposed method is also shown for the same condition in the related figure. In all simulations, the speed reference is 50% rated value. By investigating these figures, it is clear that the ripples are sensitive to the weighting factor for all torque conditions. Note that the cost function is normalized in order to ease the sense of comparison. It means that \( Q = 1 \) gives the equal control weight to the torque and the phase angle. When the weighting factor is set to a small value, i.e., \( Q = 0.01 \), which means the phase angle control is less important than the torque control, the torque ripple is also very high because the machine cannot be controlled only by the torque control. If \( Q = 0.71 \), the best result was achieved for 50% rated torque. It happens in \( Q = 0.39 \) for 5% rated torque, and in \( Q = 0.89 \) for 100% rated torque. These results are better than the results of equal weighted form (\( Q = 1 \)). If the weighting factor is \( Q = 1.5 \) again all of the ripples were increased. On the other hand, the results...
of the proposed method showed that the ripples are similar to the best condition of the previous direct angle control method but without the need to tune the weighting factor.

Another conclusion which is achieved by Fig. 8 is the performance at near rated torque condition. This condition is the case which the manipulated boundaries will work for the proposed method to avoid the saturation in the motor core. It is seen that the phase angle for the proposed method is increased to 51° because the d-axis component of the flux was limited based on (25) and (26). However, the previous method tried to keep the phase angle at 45°. This results in a bigger d-axis component of the flux and the motor saturation consequently. On the other hand, phase angle equal to 45° is not the solution of minimum current when the core is saturated and finding the optimum phase angle needs tedious offline and online calculations for the basic method. This problem is avoided by automatically increase of the phase angle in the proposed method. Thus, the stator current amplitude is slightly smaller than that in the previous method.

B. EXPERIMENTAL RESULTS

The experiments are also performed to verify the performance of the proposed method. Fig. 9 shows experimental set.

Fig. 10 shows the responses of the proposed method and the previous predictive direct angle control method [10] at 50% rated speed and 50% nominal load torque. In Fig. 10-a, it can be seen that the torque has risen in less than a few steps, which indicates a fast dynamic response of the torque. The manipulated reference of the phase angle shows that it is increased to a value of more than 45° during the transient state of the torque in order to maintain the needed torque without saturation in the core. After the rise of the torque, the reference manipulation has not occurred because the torque was less than the rated value. The limitation of the magnetization current is $1.44\lambda_n/L_s = 1.063A$ based on (25). Therefore, the torque response contains an overshoot, unlike the conventional predictive torque control. The result of this attitude is the dynamic current minimization.

The same test was repeated for the predictive direct angle control method [10] and the results are depicted in Fig. 10-b. Based on the obtained simulation results, the weighting factor is set to 0.71 for this operating point. The results showed a smaller ripple for the phase angle because it was directly
TABLE 2. Average ripple value at different operating point.

| Operating point | Ripple of \( T_c \) (%) | Ripple of \( \Delta \alpha \) (%) | Ripple of \( \Delta \lambda \) (%) |
|-----------------|--------------------------|--------------------------|--------------------------|
| \( \omega_0 \cdot 20\% \omega_s \cdot T_r < 10\% T_L \) | 8.66 | 29.21 | 1.51 |
| \( \omega_0 \cdot 50\% \omega_s \cdot T_r = 30\% T_L \) | 9.41 | 26.19 | 1.62 |
| \( \omega_0 \cdot 50\% \omega_s \cdot T_r = 30\% T_L , \Delta R = 100\% \) | 10.31 | 31.48 | 1.73 |
| \( \omega_0 \cdot 50\% \omega_s \cdot T_r = 30\% T_L , \Delta R = 100\% \) | 9.84 | 27.96 | 1.65 |
| \( \omega_0 \cdot 50\% \omega_s \cdot T_r = 50\% T_L \) | 9.43 | 26.28 | 1.62 |
| \( \omega_0 \cdot 60\% \omega_s \cdot T_r = 50\% T_L \) | 8.88 | 19.14 | 1.63 |
| \( \omega_0 \cdot 60\% \omega_s \cdot T_r = 60\% T_L \) (step) | 7.76 | 14.58 | 1.69 |
| \( \omega_0 \cdot 80\% \omega_s \cdot T_r = 100\% T_L \) | 7.85 | 20.23 | 1.73 |
| \( \omega_0 \cdot 50\% \omega_s \cdot T_r = 50\% T_L , \Delta L = 100\% \) | 9.42 | 57.48 | 2.15 |
| \( \omega_0 \cdot 80\% \omega_s \cdot T_r = 50\% T_L , \Delta L = 50\% \) | 9.58 | 41.37 | 1.86 |
| \( \omega_0 \cdot 80\% \omega_s \cdot T_r = 50\% T_L , \Delta L = 50\% \) | 11.49 | 27.5 | 1.95 |
| \( \omega_0 \cdot 80\% \omega_s \cdot T_r = 50\% T_L , \Delta L = 50\% \) | 13.91 | 38.03 | 1.96 |

FIGURE 10. Responses of (a) the proposed method (b) method of [10] at 50% rated speed and 50% nominal load torque.

also included in the cost function. However, the torque ripple is higher for the previous method. The quantified results are summarized in Table 2.

Fig. 11 shows the steady-state responses of the proposed and the previous method [10] at 60% rated speed and 90% nominal load torque. In Fig. 11-a which is the result of the proposed method, the phase angle is automatically increased to 50° because of the flux limitation occurred based on (25) and (26) when the torque was close to rated value. The flux magnitude shows flux increase is avoided similar to the simulation result. The same scenario is checked for the previous direct angle control method [10] and the results are shown in Fig. 11-b. The weighting factor is set to 0.87 from the simulation results. The results showed that the achieved flux and current are higher than those of the proposed method.

Also, the current shape showed that the saturation was more probable in this test.

The low speed performance is also studied. Fig. 12 shows the responses of the proposed method and the method in [10] at 20% rated speed and in light load condition. The comparison between the proposed method and the previous method at low speed operating point showed that both torque ripple and phase angle ripple are slightly lower for the proposed method. However, there is no big difference between the results of this test. Note that the weighting factor in the method of [10] is set to 0.5 by performing the try and error in the simulations. The quantitative measures of Figs. 10, 11, and 12 between the proposed method and the previous method [10] are illustrated in Table 3. The results show that the tracking error is improved by the proposed method in all three tests. Also, the torque to current ratio is improved for two cases but it is
equal for both methods in the low-speed test. The impressive improvement is the computational time because seven times prediction is eliminated by the proposed method.

The no-load condition was tested in 80% nominal speed condition and the results are reported in Fig. 13. After the startup, the ripple of the phase angle is increased because the variation estimated torque changes the minimum value of the $\lambda_{sd}$ based on (24). The torque ripple alternated the minimum value of the $\lambda_{sd}$ and the phase angle consequently. However, this phase alternation which is a part of the control algorithm resulted in the stability of the current, torque, and flux. To clearly indicate this effect, the load torque is slightly increased and the result is reported in Fig. 14. It can be seen that the phase angle ripple was reduced in this test.

The effect of the parameter mismatch is studied by Fig. 15 and Fig. 16. In Fig. 15, the uncertainty effect of the stator resistance is studied on the proposed method. It was tested at 50% rated speed and 30% nominal load torque. The results with accurate parameters are reported in Fig. 15-a. When the stator resistance was increased by approximately 100% and the test was repeated, Fig. 15-b reports the results. It is understandable from the comparison of these two figures that the torque dynamic response is slightly deteriorated. But in general, the proposed method retains its stability despite the change of the stator resistance.
FIGURE 17. Apply suddenly load (a) the proposed method (b) the method of [10].

The uncertainty of the stator inductance is also studied in Fig. 16. The steady state results at 50% rated torque and 80% rated speed is studied. In Fig. 16-a, the accurate value of the inductance is used in the prediction model. The results with a 50% error of the inductance are shown in Fig. 16-b. It showed that the torque ripple, flux ripple, and the current distortion were increased but the current minimization was accomplished.

Fig. 17 shows the performances of the proposed method and the method of [10] in applying a sudden load equal to 60% rated torque. In these tests, the speed was 60% nominal speed. As it can be seen in Fig. 17-a, the ripple of the angle was high before the load exertion because of the discussed issue about Fig. 13. Immediately after the load exertion, the phase angle of the current was controlled at 45°. In this test, it is well visible that the proposed method maintains its sustainability after load disturbance. Fig. 17-b shows the result of the same test for the previous method. The comparison showed that the torque ripple was smaller for the previous method because the reference was kept equal to 45°. The result of that fixed reference was a bigger amplitude of the current before load exertion. The current amplitude was 1.81A before load exertion for the proposed method but it was 2.15A for the previous method.

Table 2, shows the average ripple value of the proposed method for the torque, the current angle, and the flux in the steady-state, relative to the nominal value of each at different operating point. It shows that the proposed method provided similar control for the torque and the flux in a wide range of operating point without a need for weighting factor tuning. The phase angle ripple is dependent on the operating point which is part of the angle manipulation scenario in order to keep the torque and flux control. Also, the sensitivity to the parameter mismatch is reported in this table. It can be seen that the proposed method is robust against 100% mismatch for the stator and rotor resistance and also 30% mismatch for the inductance.

V. CONCLUSION

A simplified predictive direct angle control is proposed in this paper. By this method, the features of the direct angle control were improved while there is no need for weighting factor calculation. Due to the use of the MTPA method, the predicted angle is set 45° and the required torque is obtained for the minimum current vector.

In this method, the flux is automatically optimized by controlling the angle between the stator current and the rotor flux. Also, the phase angle reference is automatically manipulated by the limitation of the minimum and maximum value of the direct component of the stator flux. By this technique, the phase angle was automatically decreased in very low torque condition and increased in near rated torque condition. This effect was not possible in the previous version of the direct angle control.

The experimental tests on this method validated the effectiveness of that at different operating points. Also, it was shown that the method kept the stability in the load disturbance and the variation of the stator resistance.

To sum up, there are two advantages over the previous version of the direct angle control about the proposed method:

1) There is no need for weighting factor tuning
2) The optimum phase angle is not fixed on 45° which is not optimum for the light and also the rated torque condition.

APPENDIX PROOF OF (5)

The second difference equation of the induction motor is the origin of (5) as below:

$$\frac{d}{dt} \lambda_r = -R_r I_r + j \omega_r \lambda_r$$

(28)

On the other hand, the rotor current can be expressed by the following equation based on the relationship between the fluxes and currents.

$$I_r = \frac{1}{L_m} (\lambda_s - L_s I_s)$$

(29)

If (29) is applied to eliminate the rotor current from (28) and the discrete form of that is considered (5) will be attained.

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