A New Control Design with Dead-Beat Behavior for Stator Current Vector in Three-Phase AC Drives

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Abstract *The inner stator current control loop plays a decisive role for the quality of the FOC-based 3-phase AC drive systems using IMSR or PMSM. Although the compensation controllers with finite adjustment time which were inaccurately called by serveral papers as dead-beat controllers, have been successful in industry practices, the improved dead-beat controller is promising to possess advantages for industrial equipments. The paper deals with a new approach to the stator current vector controller which ensures the quality requirements "Dynamic - Accuracy - Decoupling", and was also unpublished in any other studies about control of electrical drives.*

Keywords *Three-phase AC drive, IMSR, PMSM, linear stator current controller, compensation controller, dead-beat controller, finite adjustment time.*

Symbols and abbreviations

A, B	Denominator, numerator of G_S
$\mathbf{G}_h$, $\mathbf{G}_w$	Open, closed loop transfer matrix
G_R, G_S, G_W	Controller, process, closed loop transfer functions
$\mathbf{I}$	Unit matrix
$\mathbf{i}_s$, i_{sd}, i_{sq}	Stator current vector and its *dq* components
L, L_1, L_2	Control functions
$\mathbf{L}$	Matrix of control functions
$\mathbf{R}_I$	Stator current controller
R, P	Numerator, denominator of G_R
$\mathbf{u}_s$, u_{sd}, u_{sq}	Stator voltage vector and its *dq* components
$\mathbf{y}$, y_d, y_q	Output vector of $\mathbf{R}_I$ and its *dq* components
ψ_r, ψ_{rd}	Rotor flux and its *d* component
ω, ω_s	Angle speed of rotor, angle speed of stator-side vectors
ϑ, ϑ_s	Rotor angle, rotor flux angle
$\mathbf{\Phi}$, $\mathbf{H}$, $\mathbf{h}$	System, input, disturbance matrix of process model of stator current $\mathbf{i}_s$
Φ_{11}, Φ_{12}, Φ_{13}, Φ_{14}	Components of system matrix $\mathbf{\Phi}$
FAT	Finite adjustment time
FOC	Field Oriented Control
IMSR	Induction Motor with Squirelcage Rotor
MIMO	Multi-Input/Multi-Output
PMSM	Permanentmagnet Excited Synchronous Motor
SISO	Single-Input/Single-Output

I. Introduction

Field Oriented Control (see [1], [2]), is the most implemented physical control principle in modern three-phase AC drive systems. Today, the FOC-based industrial AC drive systems are nearly perfect. Theoretically, the FOC would be seen as a view point which leads to the conclusion that the three-phase AC motors operate based on the same physical nature (flux and torque forming process) like DC motors. The FOC structure in accordance with the physical nature of the machine together with perfect designed control laws ensuring stability for the system, that combination gives FOC a superior advantage over other methods, most notably the harmonics in electric torque.

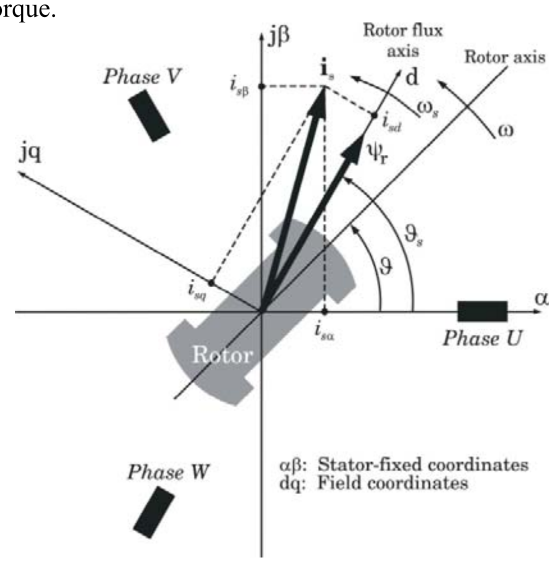


Fig 1: Vector of the stator currents of IMSR in stator-fixed and field coordinates dq

The main idea of the FOC can be summarized as follows:

- The three-phase quantities like currents, voltages and fluxes are represented in form of complex vectors.
- All complex vectors will then be described in a Cartesian coordinate system with *dq* axes, which circulate synchronously with all vectors. The real axis *d* of the coordinate system (Fig 1) is identical with the direction of the rotor flux ψ_r.
- In the coordinate system *dq*, the component:
  + i_{sd} is with delay of first order proportional to the rotor flux $|\psi_r| = \psi_{rd}$. That means, i_{sd} plays the role of field forming current like excitation current of DC motors.
  + If it is possible to control the rotor flux $|\psi_r| = \psi_{rd}$ constantly, then i_{sq} is directly proportional to the electric torque m_M.

From this main idea, a current control method for i_{sd} and i_{sq} fulfilling the three requirements "*Dynamic - Accuracy - Decoupling*" is a must while designing the inner control loop of a three-phase AC drive system. Many linear control designs (very successful in the indutry) which totally fulfill the above requirements can be found in [1], [2] (Fig 2).

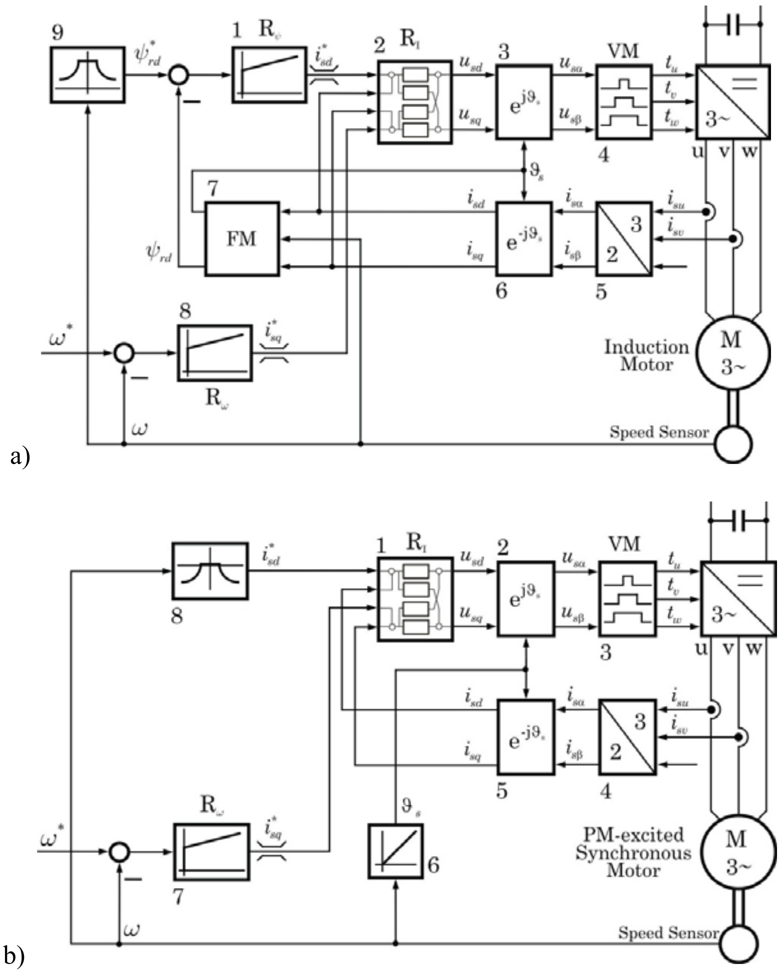


Fig 2: Structure of three-phase AC drive systems using IMSR (a) or PMSM (b) with a inner control loop fullfiling the requirements "Dynamic - Accuracy - Decoupling" a mandatory to ensure the success of the FOC principle [2]

The inner loop controller $\mathbf{R}_I$ in [1] were designed by using the *compensation method* to reach the following two goals. The controller has:

- *Goal* 1: to lead the output variable (the actual value) *along a predefined trajectory* so that …
- *Goal* 2: … the actual value will match the setpoint after exactly a finite *number of N sampling periods* and will hold on to it.

The *pole pair of a control structure with the so called compensation controller is assigned at the origin in the z domain*. This is why many authors mistakenly call this controller a *dead-beat controller*.

Although two compensation and dead-beat controllers have the common feature that every poles are assigned at the origin, the design of the dead-beat controller only includes "*Goal* 2" and not interested in "*Goal* 1" (see [3]).

See [3] when designing the dead-beat controller for SISO systems. This paper proposes a new design of dead-beat controllers for MIMO systems, such as stator current vector $\mathbf{i}_s$.

II. DESIGNS OF DIGITAL CONTROLLERS WITH FINITE ADJUSMENT TIME (FAT)

Arcording to [3], the two design methods - compensation and dead-beat - follow the same "*Goal* 2". Therefore they belong to the group of controllern with FAT. For simplicity, this section briefly describes the two design methods for SISO processes.

A. *Compensation controller design for SISO process*

Based on "*the form of the predefined output trajectory*" following the goal 1 and "*the number of N sampling periods*" following the goal 2, it's easy to point out that the transfer function $G_w(z^{-1})$ of the closed loop must be a polynomial of N-th degree, with the sum of polynomial coefficients equal to 1. After "*compensating the models*" the following controller $G_R(z^{-1})$ would be obtained:

$$G_R\left(z^{-1}\right)=\frac{1}{G_S\left(z^{-1}\right)}\frac{G_w\left(z^{-1}\right)}{1-G_w\left(z^{-1}\right)} \tag{1}$$

Because $G_W(z^{-1})$ is only chosen in depence on "*the form of the predefined trajectory*", that means independent on $G_S(z^{-1})$, it could be that $G_W(z^{-1})$/[1-$G_W(z^{-1})$] can not exactly compensate the poles or zero points of $G_S(z^{-1})$. This is especially dangerous if this method is used to design $G_R(z^{-1})$ for slow processes (poles near the unit circle and far away from the origin in the z domain).

However, because stator currents are process with small inertia (poles near the origin in the z domain), this design method has been successfully applied in industrial devices ([4], [5], [6], [7]).

B. *Dead-beat controller design for SISO process*

Different to compensation controller, the design of dead-beat controllers only follows the goal 2. This leads to the fact that $G_w(z^{-1})$ must also have the form of a polynomial of N-th degree, with the sum of polynomial coefficients equal to 1. According to [3], the problem of the $G_R(z^{-1})$ design now refers to finding a polynomial $L(z^{-1})$. If the transfer function of the process is $G_S(z^{-1})= B(z^{-1})/A(z^{-1})$, the controller would be obtained as follows:

$$G_R\left(z\right)=\frac{R\left(z^{-1}\right)}{P\left(z^{-1}\right)}=\frac{L\left(z^{-1}\right)A\left(z^{-1}\right)}{1-L\left(z^{-1}\right)B\left(z^{-1}\right)} \tag{2}$$

The polynomial $L(z^{-1})$ has to fulfill:

$$L\left(1\right)=\frac{1}{B\left(1\right)} \quad\rightarrow\quad \sum_{i=0}^{s} l_i = 1\Big/\sum_{j=0}^{m} b_j \tag{3}$$

with: l_i coefficients of the polynomial $L(z^{-1})$
b_j coefficients of the polynomial $B(z^{-1})$, numerator of $G_S(z^{-1})$
$L(1)$, $B(1)$ sum of polynomial coefficients of $L(z^{-1})$, $B(z^{-1})$.

The equation (3) is the first condition for finding the coefficients of the polynomial $L(z^{-1})$. If we choose the degree of $L(z^{-1})$ greater than 0, we must rely on the technical specification of the system to generate a sufficient number of equations corresponding to the coefficients of $L(z^{-1})$.

The important differences of the controller (2) compared with (1) are:

- *The 1st difference*: The dynamics of the controller (2) is totally dependent on the dynamics of $G_S(z^{-1})$. The degree of the polynomials in numerator and denominator of (2) is always limited by the degree of the polynomials in numerator and denominator of $G_S(z^{-1})$.
- *The 2nd difference*: The controller (2) do not contain $G_W(z^{-1})$ having the form of a polynomial of N-th degree. As discussed above, the ability to freely choose $G_W(z^{-1})$ involves the risk of instability for systems with high inertia (pole points near the unit circle).

The next part of the article introduces the expansion, applying the idea of designing a dead-beat controller for a SISO to a MIMO process, which is the stator current vector $\mathbf{i}_s$ of the IMSR in the hope of taking advantage of the firstly above described difference.

III. DEAD-BEAT CONTROLLER DESIGN FOR IMSR

A. *General Design*

Given is the following discrete process model of stator current of IMSR (see [2]):

$$\mathbf{i}_s\left(k+1\right)=\mathbf{\Phi}\mathbf{i}_s\left(k\right)+\mathbf{H}\mathbf{u}_s\left(k\right)+\mathbf{h}\psi\left(k\right) \tag{4}$$

with:

$$\mathbf{\Phi}=\begin{bmatrix}\Phi_{11} & \Phi_{12}\\ -\Phi_{12} & \Phi_{11}\end{bmatrix} =\begin{bmatrix}1-\dfrac{T}{\sigma}\left(\dfrac{1}{T_s}+\dfrac{1-\sigma}{T_r}\right) & \omega_s T\\ -\omega_s T & 1-\dfrac{T}{\sigma}\left(\dfrac{1}{T_s}+\dfrac{1-\sigma}{T_r}\right)\end{bmatrix} \tag{5}$$

$$\mathbf{H}=\begin{bmatrix}h_{11} & 0\\ 0 & h_{11}\end{bmatrix}=\begin{bmatrix}\dfrac{T}{\sigma L_s} & 0\\ 0 & \dfrac{T}{\sigma L_s}\end{bmatrix} \tag{6}$$

$$\mathbf{h}=\begin{bmatrix}\Phi_{13} & \Phi_{14}\\ -\Phi_{14} & \Phi_{13}\end{bmatrix}=\begin{bmatrix}\dfrac{1-\sigma}{\sigma}\dfrac{T}{T_r} & \dfrac{1-\sigma}{\sigma}\omega T\\ -\dfrac{1-\sigma}{\sigma}\omega T & \dfrac{1-\sigma}{\sigma}\dfrac{T}{T_r}\end{bmatrix} \tag{7}$$

In (4), the component $\mathbf{h}\psi\left(k\right)$ that characterizes the influence of the flux on the voltage equation is considered a disturbance signal, so the actual control signal of the stator current model (taking into account the dead time due to the hardware) will be:

$$\mathbf{y}\left(k-1\right)=\mathbf{H}\mathbf{u}_s\left(k\right)+\mathbf{h}\psi\left(k\right) \tag{8}$$

The voltage fed to the stator is:

$$\mathbf{u}_s\left(k+1\right)=\mathbf{H}^{-1}\left[\mathbf{y}\left(k\right)-\mathbf{h}\psi\left(k+1\right)\right] \tag{9}$$

The simplified process model with the input signal $\mathbf{y}$:

$$\mathbf{i}_s(k+1) = \mathbf{\Phi}\mathbf{i}_s(k) + \mathbf{y}(k-1) \qquad (10)$$

The equation (10) is transformed into z domain:

$$(z\mathbf{I} - \mathbf{\Phi})\mathbf{i}_s(z) = z^{-1}\mathbf{y}(z) \qquad (11)$$

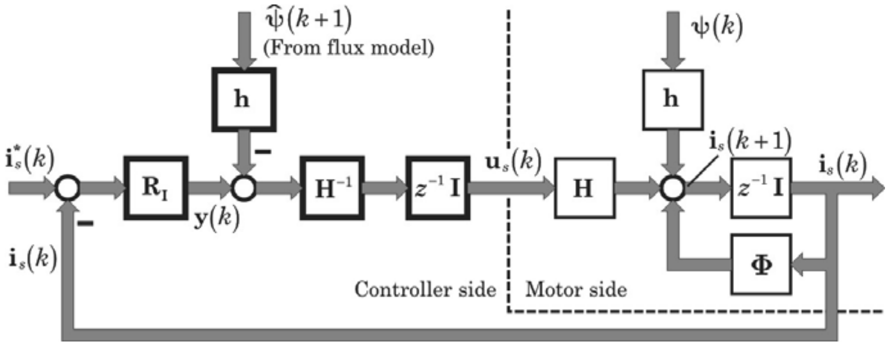


Fig 3: Block structure of the current vector controller for IMSR or PMSM [2]

We replace $\mathbf{A}(z) = z\mathbf{I} - \mathbf{\Phi}; b(z) = z^{-1}$. The control equation (Fig 3) using control deviation is written as follows:

$$\mathbf{y}(z) = \mathbf{R}_I(z)\left[\mathbf{i}_s^*(z) - \mathbf{i}_s(z)\right] \qquad (12)$$

Based on the idea of the dead-beat control design for SISO process in the section 2.2, the following matrix of polynomials must be found:

$$\mathbf{L}(z^{-1}) = \begin{bmatrix} L_1(z^{-1}) & 0 \\ 0 & L_2(z^{-1}) \end{bmatrix} \qquad (13)$$

Additionally it must be proofed, that the closed loop using the following controller:

$$\mathbf{R}_I(z) = \mathbf{A}(z)\mathbf{L}(z^{-1})\left[\mathbf{I} - z^{-1}\mathbf{L}(z^{-1})\right]^{-1} \qquad (14)$$

really has the FAT performance. In fact, matrix **A** is invertible because there is always:

$$\det \mathbf{A} = \left(z - 1 + \frac{T}{\sigma}\left(\frac{1}{T_s} + \frac{1-\sigma}{T_r}\right)\right)^2 + (\omega_s T)^2 > 0 \quad (15)$$

From (11) it will be obtained:

$$\mathbf{i}_s(z) = \mathbf{A}^{-1}(z) z^{-1}\mathbf{y}(z) \qquad (16)$$

Inserting (12) and (14) into (16) it follows:

$$\mathbf{i}_s(z) = \underbrace{z^{-1}\mathbf{L}(z^{-1})\left[\mathbf{I} - z^{-1}\mathbf{L}(z^{-1})\right]^{-1}}_{\mathbf{G}_h(z)}\left[\mathbf{i}_s^*(z) - \mathbf{i}_s(z)\right] \qquad (17)$$

In (17), the transfer function $\mathbf{G}_h(z)$ of the open loop is a diagonal matrix:

$$\mathbf{G}_h(z) = \begin{bmatrix} \dfrac{z^{-1}L_1(z^{-1})}{1 - z^{-1}L_1(z^{-1})} & 0 \\ 0 & \dfrac{z^{-1}L_2(z^{-1})}{1 - z^{-1}L_2(z^{-1})} \end{bmatrix} \qquad (18)$$

The equation (18) indicates that the controller (14) successfully decouples the current components. After some transformations of (17) the closed loop relation will be obtained:

$$\mathbf{i}_s(z) = z^{-1}\mathbf{L}(z^{-1})\mathbf{i}_s^*(z) \qquad (19)$$

The formula (19) means if $L_1(z^{-1})$ and $L_2(z^{-1})$ are polynomials of n_1 -th and n_2 -th degrees, then the current components i_{sd} and i_{sq} will follow their set-points after exactly $n_1 + 1$ and $n_2 + 1$ sampling periods.

Inserting (5) and (13) into (14):

$$\mathbf{R}_I(z) = \begin{bmatrix} \dfrac{(z - \Phi_{11})L_1(z^{-1})}{1 - z^{-1}L_1(z^{-1})} & \dfrac{-\Phi_{12}L_2(z^{-1})}{1 - z^{-1}L_2(z^{-1})} \\ \dfrac{\Phi_{12}L_1(z^{-1})}{1 - z^{-1}L_1(z^{-1})} & \dfrac{(z - \Phi_{11})L_2(z^{-1})}{1 - z^{-1}L_2(z^{-1})} \end{bmatrix} \qquad (20)$$

So that the control laws do not cause future errors (stationary control errors), the polynomials $L_1(z^{-1})$ and $L_2(z^{-1})$ must not contain the coefficient l_0. Additionally, to eliminate the stationary control errors the transfer function $\mathbf{G}_w$ of the closed loop must be equal **I** under stationary conditions ($z = 1$). Therefore, from (19) ist must be:

$$L_1(1) = L_2(1) = 1 \qquad (21)$$

Note: $L_1(1)$, $L_2(1)$ are the sums of the coefficients of the two polynomials $L_1(z^{-1})$, $L_2(z^{-1})$.

B. *Choice of $L_1(z^{-1})$ and $L_2(z^{-1})$ as first degree polynomials*

$$\begin{aligned} L_1(z^{-1}) &= l_{11}z^{-1} \\ L_2(z^{-1}) &= l_{12}z^{-1} \end{aligned} \qquad (22)$$

Inserting (21) into (22), $l_{11} = l_{12} = 1$ will be obtained, and the transfer function of the closed loop will have the following form $\mathbf{G}_w(z) = z^{-2}$. This is the case of the compensation controllers presented in [2].

C. *Choice of $L_1(z^{-1})$ and $L_2(z^{-1})$ as second degree polynomials*

For simplicity, we choose the same two polynomials for L_1 and L_2:

$$L_1\left(z^{-1}\right)=L_2\left(z^{-1}\right)=l_1z^{-1}+l_2z^{-2} \quad (23)$$

After inserting (21) into (23) the first condition is obtained:

$$l_1+l_2=1 \quad (24)$$

The controller (20) can be written in form of difference equations:

$$\begin{aligned}y_d(k)&=l_1y_d(k-2)+l_2y_d(k-3)+l_1e_d(k)\\&+(l_2-l_1\Phi_{11})e_d(k-1)-l_2\Phi_{11}e_d(k-2)\\&-l_1\Phi_{12}e_q(k-1)-l_2\Phi_{12}e_q(k-2)\end{aligned} \quad (25)$$

$$\begin{aligned}y_q(k)&=l_1y_q(k-2)+l_2y_q(k-3)+l_1e_q(k)\\&+(l_2-l_1\Phi_{11})e_q(k-1)-l_2\Phi_{11}e_q(k-2)\\&+l_1\Phi_{12}e_d(k-1)+l_2\Phi_{12}e_d(k-2)\end{aligned} \quad (26)$$

The voltage components can be calculated as follows:

$$u_{sd}(k+1)=h_{11}^{-1}\left[y_d(k)-\Phi_{13}\psi'_{rd}(k+1)\right] \quad (27)$$

$$u_{sq}(k+1)=h_{11}^{-1}\left[y_q(k)+\Phi_{14}\psi'_{rd}(k+1)\right] \quad (28)$$

The initial amplitude value of the voltage can be limited by the accordingly chosen values for l_1 and l_2. In particular, as follows:

$$\begin{aligned}u_{d0}&=h_{11}^{-1}\left(y_{d0}-\Phi_{13}\psi'_{rd0}\right)=h_{11}^{-1}\left(l_1e_{d0}-\Phi_{13}\psi'_{rd0}\right)\\u_{q0}&=h_{11}^{-1}\left(y_{q0}+\Phi_{14}\psi'_{rd0}\right)=h_{11}^{-1}\left(l_1e_{q0}+\Phi_{14}\psi'_{rd0}\right)\end{aligned} \quad (29)$$

The magnetizing current is chosen in the range:

$$0\le i_{md}=\psi'_{rd}\le i_{sdN} \quad (30)$$

From there, in conjunction with condition (24), the parameters l_1 and l_2 can be chosen as follows:

$$\begin{aligned}l_1&=\min\left\{\frac{h_{11}u_{d0}}{e_{d0}};\frac{h_{11}u_{q0}-\Phi_{14}i_{sdN}}{e_{q0}}\right\}\\l_2&=1-l_1\end{aligned} \quad (31)$$

If $l_1=1$ and $l_2=0$ is chosen, then the obtained controller will be the same as presented in [2].

D. Treatment of the limitation of the stator voltage

Although l_1 and l_2 are chosen in accordance with (31), the initial amplitude value of the voltage (usually also the maximum value) is limited, it may occur that the controller requires an amplitude which exceeds the supply capacity of the inverter during operation. Therefore, it is necessary to solve the problem that the stator voltage falls within the limit range. The principle of solving this problem has been fully introduced in section 5.5 of [2], which can be summarized in two parts:

- *First*, define the limit value for each voltage component u_{sd} and u_{sq} from the amplitude limit of the voltage vector (dependent on the inverter). This is the so called *splitting strategy* at voltage limitation.
- *Second*, backward correcting the control errors (named the *correction strategy*) to stop the integral part of the current control algorithm.

For the new controller design in this paper, the *splitting strategy* of the amplitude limit of the voltage vector into components has not changed because it is completely dependent on the inverter. The backward *correction strategy* during voltage limitation is also adopted by [2], the only difference in the calculation expression is that the control law structure has changed.

The control laws using the corrected variables (index "*c*": corrected):

- The *d*-axis component:

$$\begin{aligned}y_d(k)&=l_1y_{d,c}(k-2)+l_2y_{d,c}(k-3)+l_1e_d(k)\\&+(l_2-l_1\Phi_{11})e_{d,c}(k-1)-l_2\Phi_{11}e_{d,c}(k-2)\\&-l_1\Phi_{12}e_{q,c}(k-1)-l_2\Phi_{12}e_{q,c}(k-2)\end{aligned} \quad (32)$$

$$\begin{aligned}y_{d,c}(k)&=l_1y_{d,c}(k-2)+l_2y_{d,c}(k-3)+l_1e_{d,c}(k)\\&+(l_2-l_1\Phi_{11})e_{d,c}(k-1)-l_2\Phi_{11}e_{d,c}(k-2)\\&-l_1\Phi_{12}e_{q,c}(k-1)-l_2\Phi_{12}e_{q,c}(k-2)\end{aligned} \quad (33)$$

$$u_{sd}(k+1)=h_{11}^{-1}\left[y_d(k)-\Phi_{13}\psi'_{rd}(k+1)\right] \quad (34)$$

$$u_{sd,c}(k+1)=h_{11}^{-1}\left[y_{d,c}(k)-\Phi_{13}\psi'_{rd}(k+1)\right] \quad (35)$$

- The *q*-axis component:

$$\begin{aligned}y_q(k)&=l_1y_{q,c}(k-2)+l_2y_{q,c}(k-3)+l_1e_q(k)\\&+(l_2-l_1\Phi_{11})e_{q,c}(k-1)-l_2\Phi_{11}e_{q,c}(k-2)\\&+l_1\Phi_{12}e_{d,c}(k-1)+l_2\Phi_{12}e_{d,c}(k-2)\end{aligned} \quad (36)$$

$$\begin{aligned}y_{q,c}(k)&=l_1y_{q,c}(k-2)+l_2y_{q,c}(k-3)+l_1e_{q,c}(k)\\&+(l_2-l_1\Phi_{11})e_{q,c}(k-1)-l_2\Phi_{11}e_{q,c}(k-2)\\&+l_1\Phi_{12}e_{d,c}(k-1)+l_2\Phi_{12}e_{d,c}(k-2)\end{aligned} \quad (37)$$

$$u_{sq}(k+1)=h_{11}^{-1}\left[y_q(k)+\Phi_{14}\psi'_{rd}(k+1)\right] \quad (38)$$

$$u_{sq,c}(k+1)=h_{11}^{-1}\left[y_{q,c}(k)+\Phi_{14}\psi'_{rd}(k+1)\right] \quad (39)$$

From there, the corrected control errors will be obtained:

$$\begin{aligned}e_{d,c}(k)&=e_d(k)-\frac{y_d(k)-y_{d,c}(k)}{l_1}\\&=e_d(k)-\frac{h_{11}}{l_1}\left[u_{sd}(k+1)-u_{sd,c}(k+1)\right]\end{aligned} \quad (40)$$

$$y_{d,c}(k)\ =h_{11}u_{sd,c}(k+1)+\Phi_{13}\psi'_{rd}(k+1) \quad (41)$$

$$e_{q,c}(k) = e_q(k) - \frac{y_q(k) - y_{q,c}(k)}{l_1} = e_q(k) - \frac{h_{11}}{l_1}\left[u_{sq}(k+1) - u_{sq,c}(k+1)\right] \quad (42)$$

$$y_{q,c}(k) = h_{11} u_{sq,c}(k+1) - \Phi_{14}\psi'_{rd}(k+1) \quad (43)$$

IV. Simulation and Experimental Results

The dead-beat design architecture above has been simulated with MATLAB / Simulink and PLECS (a tool supporting simulation of power electronics and electric motors [8]). The conditions and model parameters are given in Tab.1 below.

Tab. 1 *Simulation parameters and conditions*

1. IMSR parameters	*Symbols*	*Values*
Nominal power	P_{nom}	0.5 kW
Nominal speed	n_{nom}	3000 rpm
Nominal current	I_{nom}	7.4 A_{RMS}
Pole pair	z_p	1
Rotor resistance	R_r	0.42 Ω
Stator resistance	R_s	0.37 Ω
Rotor inductance	L_r	34.25 mH
Stator inductance	L_s	34.41 mH
Mutual inductance	L_m	33.1 mH
Power factor	$\cos\varphi$	0.9
Total leakage factor	σ	0.07
Torque of inertia	J	0.001 kgm^2
2. Simulation conditions		
Modulation frequency	f_{pwm}	5 kHz
Sampling time:		
Inner loop	T_s	200 µs
Outer loop	T_{sw}	2 ms

Some of the typical working modes of the IMSR are investigated through the following simulation scenario:

+ At $t = 0.1$s, the magnetization process.
+ At $t = 0.5$s, acceleration to the nominal value 3000 rpm.
+ At $t = 1.0$s, connection of nominal load (full load).
+ At $t = 1.5$s, reversing process down to -3000 rpm.

Simulation results show that both the flux forming and torque forming currents accurately follow the setpoint trajectories (coming from the magnetic flux controller and the speed controller in the outer loop) in all working modes (Fig 4).

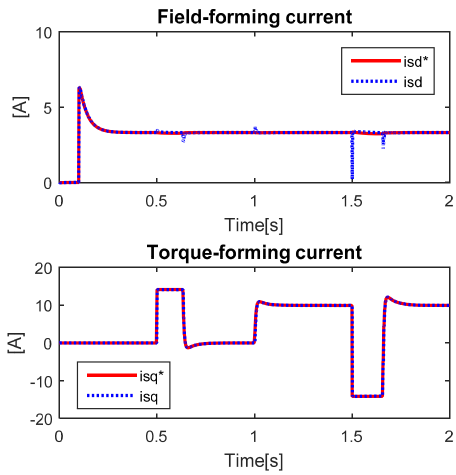
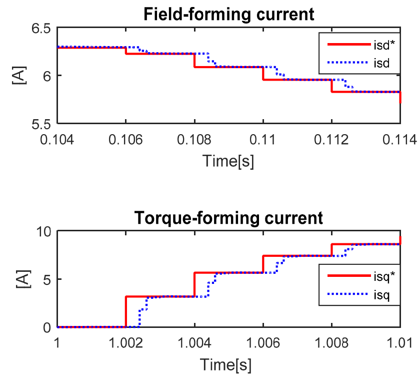
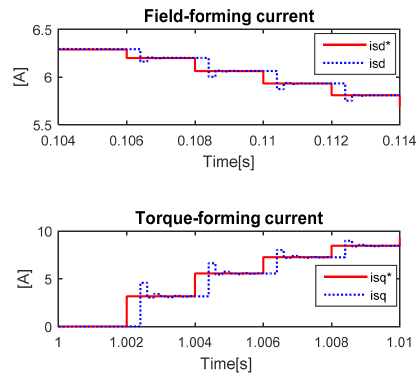


Fig 4: Current setpoint and current actual value curves during the entire simulation process using the new designed dead-beat controller

A zoom of a time 0.01 s (5 sampling periods of outer loop, Fig 5) shows a finer response time of the current loop using the dead-beat controller. When the parameter set l_1, l_2 of polynomial L is varied, the current is driven in different trajectories, but still have the common feature that the actual values will catch the setpoints after a finite number of sample periods, consistent with the theoretical content stated.



a) l1 = 0.6, l2 = 0.4



b) l1 = 1.5, l2 = -0.5

Fig 5: Excerpts from the current setpoint and current actual value curves in Error! Reference source not found. **with different pairs of polynomial coefficients: a) l_1=0.6, l_2=0.4; b) l_1=1.5, l_2=-0.5**

Based on the new current regulator, which meets the criteria "*Dynamic - Accuracy - Decoupling*", the torque can be generated quickly and the speed just for a short time brought exactly to the setpoint (0.18s for the run-up and 0.2s for reversing, Fig 6).

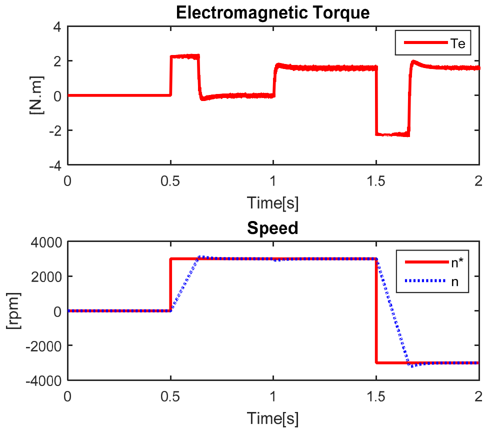


Fig 6: Torque and speed curves during run-up with no-load and reversing with full load

The control algorithm, designed in section 3, is implemented and tested on a test bench using card DS1104 from dSPACE [9]. The experimental system is shown in Fig 7.

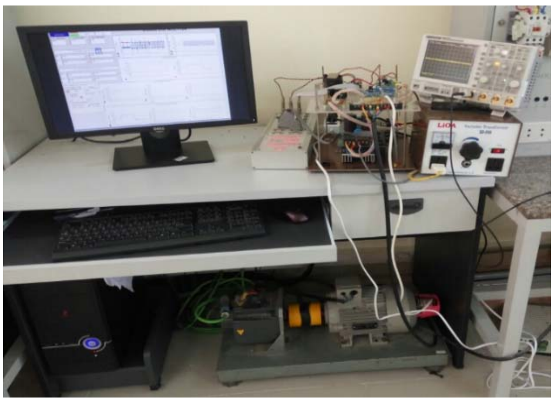
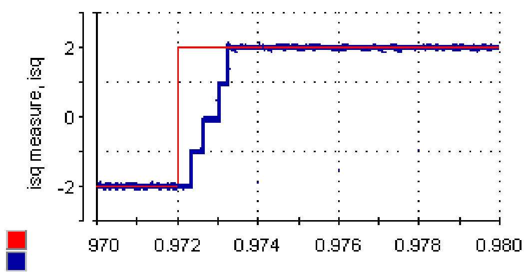


Fig 7: Test bench using card DS1104 from dSPACE [9]

The simulation results achieved in chapter 4 are also confirmed here: the FAT response of the current dead-beat controller (Fig 8) with l_1=0.6, l_2=0.4.

a)



b)

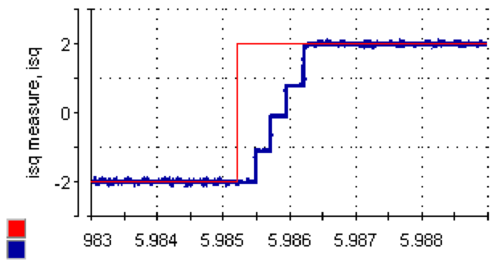


Fig 8: Excerpts from the current setpoint and current actual value curves of i_{sq} at different times

V. CONCLUSIONS

High-quality three-phase AC drives with field-oriented control require a current vector control with the properties "*Dynamic - Accuracy - Decoupling*". This problem is actually solved, and many current control concepts ([2], [6], [7]) were introduced or successfully implemented in industrial plants. However, the paper introduces a possible contribution to the improvement of the required control properties.

Inspired by the idea of a dead-beat controller for SISO processes, a new MIMO dead-beat controller is designed for the stator current vector of the three-phase AC drive using IMSR. The presented design is relatively simple and manageable. Particularly important for the characteristic of the field- and torque-forming current components i_{sd}, i_{sq} is the free choice of the parameters l_1, l_2 of the polynomial L.

The simulation and the experiments show very good results, which let hope that the new control method will one day be used in industrial equipments. In addition, this ultimate current control, combined with one of the possible nonlinear control concepts for speed control in outer loop, such as flatness-based, backstepping-based or exactly linearized concepts, can provide the optimal control structure for three-phase AC drives. The experimental investigations continue at the moment.

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