A comparative study of different methods for Boolean control designing

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Abstract: **-** *S*ystems with Boolean controls appear frequently in power electronics such as in a power converter and electrical machine assembly. In this paper we compare the performance of a Boolean control developed in a previous paper by the authors with the P.W.M approach, and sliding mode control.

The studied non-linear model used for the comparison is a non-linear system and is composed of an induction machine and its power converter. Different procedures of control system is implemented on a non-linear system The control of this system is performed in simulation on the speed

Key-Words: **-** Boolean control, switching, commutation, state space, sliding mode control, PWM.

1 Introduction

Several methods dealing with Boolean control have been developed during the last years. These strategies of Boolean control are generally applied in order to improve the behavior of switching systems. These switching systems are frequently used in industrial applications because they give a high power in output as for example power converters. The power converter is a part of Boolean input systems. The behavior of such systems is controlled by the switching ON (value 1) and OFF (value 0) of components as thyristors or transistors.

These Boolean control methods are separated in two classes. The first class of methods consists to control the process using mean values of inputs, like in the well known P.W.M (Pulse Width Modulation) technique [1][2]. The regulation is often realized by a P.I.D controller. This class of methods does not need any model of the converter or switching device. The second class of control design consists in keeping the binary values of the inputs, and in using different approaches like Sliding Mode Control(SMC) [3][4]. This technique is characterized by discontinuous control action on Variable Structure System(VSS) which changes structure upon reaching a set of switching surfaces. The switching instants are determined by appropriate sliding surface (switching surface). Sliding surface are chosen to achieve a desired dynamical response. Sliding Mode Control for multi-input systems is used to control electronics converter.

The authors have recently proposed a method for designing Boolean controls [5][6]. This technique uses directly the Boolean values to control the system in order to take into account the whole model (plant and switching device), and to act on commutations.

In the first part of this paper we present the studied model. It is a non linear system composed of an induction machine and its power converter. In the second part we will briefly recall the strategy of the different methods we are going to use. Then the different procedures of control system are implemented on the example. The aim is to control the velocity to make it follow a desired trajectory. In the last part some comments on the results are made.

2 Definition of the studied model

The application has been performed in order to evaluate the performance of different Boolean control algorithms. The system is an induction machine coupled to a power converter (figure 1). The power converter is constituted of six switching transistors (S_1, S_2, S_3, S_4, S_5, S_6) and it is supplied by a continuous voltage E_o. The induction machine has three wires at the stator and three at the rotor.

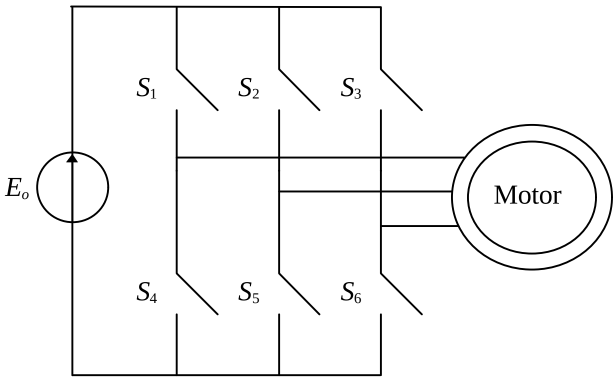

Fig.1:electric scheme of Inverter-Induction Machine

By taking statoric fluxes, rotoric fluxes and the velocity as state variables the mathematical model deduced from a Bond-Graph representation of the system [7][8] is given by :

$$\begin{cases} \frac{df_{as}}{dt} = -R_s a_r f_{as} + R_s b f_{ar} + u_a \\ \frac{df_{bs}}{dt} = -R_s a_r f_{bs} + R_s b f_{br} + u_b \\ \frac{df_{ar}}{dt} = -R_r a_s f_{ar} + R_r b f_{as} - p_o w f_{br} \\ \frac{df_{br}}{dt} = -R_r a_s f_{br} + R_r b f_{bs} + p_o w f_{ar} \\ \frac{dw}{dt} = -\frac{f}{J} w + \frac{p_o b}{J}\left(f_{bs} f_{ar} - f_{as} f_{br}\right) \end{cases} \quad (1)$$

f_{as}, f_{bs} are the equivalent statoric fluxes and f_{ar} f_{br} are the equivalent rotoric ones. w is the velocity of the asynchronous machine shaft. u_α and u_β are the voltage applied to the stator. These voltages are determined by boolean values such as :

$$\begin{cases} u_a = E_o\left(2u_1 - u_2 - u_3\right) \\ u_b = E_o\left(u_2\sqrt{3} - u_3\sqrt{3}\right) \end{cases} \quad (2)$$

The item u_i for $i = 1,2\cdots6$ are the Boolean external control of with the six transistors of switches (S_1, S_2, ..., S_6). $u_i = 0$ if the i^{th} transistor S_i is in OFF state and $u_i = 1$ if it is in ON state. For electrical reasons (see figure 1) they are linked by the relation :

$$u_i + u_{i+3} = 1 \text{ for } i = 1,2,3 \quad (3)$$

Then $u = \left[u_1 \quad u_2 \quad u_3\right]^{\text{T}}$ is the input vector of the system. The other parameters are defined as :

$$a_s = \frac{L_s}{L_r L_s - M_{sr}^2} \quad a_r = \frac{L_r}{L_r L_s - M_{sr}^2} \quad b = \frac{L_s}{L_r L_s - M_{sr}^2}$$

$$L_s = l_s - M_s \ , \ L_r = l_r - M_r$$

$$M_{sr} = \frac{3}{2} m_{sr}$$

R_s : Statoric resistance.
l_s : Statoric inductance.
R_r : Rotoric resistance.
l_r : Rotoric inductance.
M_s : Statoric mutual inductance.
M_r : Rotoric mutual inductance.
E_o : Voltage applied to the power converter.
f, J : friction and inertia of the rotor.
p_o : Number of pole pairs.
m_{sr} : mutual inductances between stator and rotor.

The aim is to control the asynchronous machine for the angular velocity ω to follow a desired velocity ω_d. We can show that to drive the velocity ω, it is adequate to control statoric fluxes by these relations:

$$\begin{cases} f_{as}ref = f_1 \cos(q_s) - f_2 \sin(q_s) \\ f_{bs}ref = f_1 \sin(q_s) + f_2 \cos(q_s) \end{cases} \quad (4)$$

with f_1 and f_2 fluxes.

q_s angular position between α and β axis.

The definition of the reference fluxes $f_{as}ref$ and $f_{bs}ref$ of statoric wires are described in the appendix.

The fluxes of the 2-phased system f_{as} and f_{bs} are measured and compared with the reference flux $f_{as}ref$ and $f_{bs}ref$ .The outputs of these comparators compose error vector ε used in the different methods we are going to compare.

The main objective of this work is for the system to track a trajectory in the state space, imposed here by the flux references $f_{as}ref$ and $f_{bs}ref$. This objective is achieved by using different methods described in the next section.

2 Methods for designing the Boolean control law

Before applying each method to the particular example, we will recall briefly the principle of each of them in the general case

2.1 Boolean control (BC)

Consider the system modeled by the state equation :

$$\dot{x} = f\left(x, u\right) \quad (5)$$

where $x = \left(x_1 \cdots\cdots x_n\right)^T \in \Re^n$ is the state vector and $u = \left(u_1 \cdots\cdots u_m\right)^T \in \{0,1\}^m$ is the input vector composed of Boolean variables.

The input vector u can take any configuration among 2^m different vectors $Config_i(u)$ containing Boolean values. For the motor problem, the number of inputs is $m = 3$, then the $Config_i(u)$ set becomes :

$$\left\{Config_i(u) \,/\, i = 1.8\right\} = \left\{ \underset{1}{\begin{pmatrix}0\\0\\0\end{pmatrix}} \underset{2}{\begin{pmatrix}1\\0\\0\end{pmatrix}} \underset{3}{\begin{pmatrix}0\\1\\0\end{pmatrix}} \underset{4}{\begin{pmatrix}1\\1\\0\end{pmatrix}} \underset{5}{\begin{pmatrix}0\\0\\1\end{pmatrix}} \underset{6}{\begin{pmatrix}1\\0\\1\end{pmatrix}} \underset{7}{\begin{pmatrix}0\\1\\1\end{pmatrix}} \underset{8}{\begin{pmatrix}1\\1\\1\end{pmatrix}} \right\} \quad (6)$$

The method proposed in [1] and [2] uses the scalar product defined as $\langle e, \vec{V}_i \rangle$, where :

- ε is the error vector expressed as $\varepsilon = (x)_d - (x)_p$ with x_p the current position of the system and x_d the desired position in the state space.
- $\vec{V}_i$ the derivative state vector $(\dot{x})_i$, calculated for each input configuration $Config_i(u)$, $i = \{1 \cdots 2^m\}$

$\vec{V}_i = (\dot{x})_i = \left(\dot{x}_p\right)_i = -\left(\frac{de}{dt}\right)_i$ if x_d is a vector supposed to be constant in the calculation interval

then $\langle \varepsilon, \vec{V}_i \rangle = -\varepsilon^T . \left(\frac{d\varepsilon}{dt}\right)_i$.

The control law has to maximize the following criterion J.

$$\max_{i=1;\cdots;2^m} \left\{ \frac{-\frac{1}{2} \cdot \left(\frac{d\left(\|\varepsilon\|^2\right)}{dt} \right)_i}{\|\varepsilon\| \cdot \left\| \left(\frac{d\varepsilon}{dt} \right)_i \right\|} \right\} \quad (7)$$

which corresponds to consider the maximum of the cosine of the angle between ε and the set of vector $\vec{V}_i$. Figure 5 illustrates the strategy.

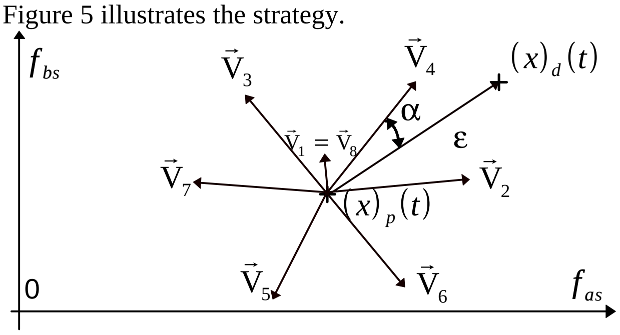

Fig.2 : Control in state space

In that case, the vector $\vec{V}_4$ forms the smallest angle with the vector ε. It corresponds to the control $Config_4(u) = (1 \quad 1 \quad 0)^T$ which will be applied to the system.

The control process is constituted now by two regulation sub-systems, one for the speed control (see appendix) and the second for the fluxes control, as we see on the following simulation scheme :

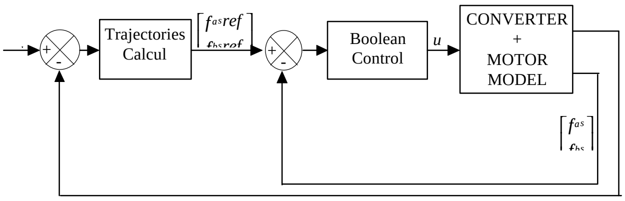

Fig.3: Simulation scheme of the process.

2.2 Sliding mode control (SMC)

This approach is based on switching function of the state variables used to create a sliding manifold [9], the purpose of which is to force the system dynamics defined by the manifold equation. When the state is maintained on this manifold, the system dynamics become insensitive to parameter variations.

Consider the dynamic system described by :

$$\dot{x} = f(x,u) \quad (8)$$

where $x \in \Re^n$. Let us define :

$$u_j = \begin{cases} u_j^+ & \text{si } s_j(x) > 0 \\ u_j^- & \text{si } s_j(x) < 0 \end{cases} \quad (9)$$

$j = 1, \ldots, p$; $s(x)=[\ s_1(x), \ldots\ldots, s_p(x)]^T$ is the sliding manifold. The results depend on the manifold $s(x)$ in the state space. It is known [10] that the sliding condition on the manifold $s(x)$ is :

$$\lim_{s_j \to 0} s_j \dot{s}_j \leq 0 \ \ j = 1,2,\ldots p \quad (10)$$

The aim is to find a control u_{de} called the equivalent control [11], from $\dot{s}_j = 0$.

Let us apply this method on the motor. The state vector is:

$$x = (x_1, x_2, x_3, x_4, x_5)^T = \left(f_{as}, f_{bs}, f_{ar}, f_{br}, w\right)^T \quad (11)$$

The control law will be defined from the sliding surfaces chosen as :

$$s(x) = \begin{pmatrix} s_1(x_1) \\ s_2(x_2) \end{pmatrix} = \begin{pmatrix} x_1 - f_1 \cos(q_s(t)) + f_2 \sin(q_s(t)) \\ x_2 - f_1 \sin(q_s(t)) - f_2 \cos(q_s(t)) \end{pmatrix} \quad (12)$$

from equation (4).

Then the equivalent control $u_{de} = \begin{pmatrix} u_a & u_b \end{pmatrix}^T$ can be determined by derivating of $s(x)$. So the result yield on sliding surfaces intersection is given by :

$$u_{de} = \begin{pmatrix} -R_s b x_3 - \sin(q_s(t))\left(R_s a_r f_2 + f_1 \dot{q}_s(t)\right) \\ \quad + \cos(q_s(t))\left(R_s a_r f_1 - f_2 \dot{q}_s(t)\right) \\ -R_s b x_4 + \sin(q_s(t))\left(R_s a_r f_1 - f_2 \dot{q}_s(t)\right) \\ \quad + \cos(q_s(t))\left(R_s a_r f_2 + f_1 \dot{q}_s(t)\right) \end{pmatrix} \quad (13)$$

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The system's behavior on the sliding surface is described as :

$$\dot{x} = \begin{pmatrix} -R_s a_r x_1 - \sin(q_s(t))\left(R_s a_r f_2 + f_1 \dot{q}_s(t)\right) \\ \quad + \cos(q_s(t))\left(R_s a_r f_1 - f_2 \dot{q}_s(t)\right) \\ -R_s a_r x_2 + \sin(q_s(t))\left(R_s a_r f_1 - f_2 \dot{q}_s(t)\right) \\ \quad + \cos(q_s(t))\left(R_s a_r f_2 + f_1 \dot{q}_s(t)\right) \\ -R_r a_s x_3 + R_r b x_1 - p_o w x_4 \\ -R_r a_s x_4 + R_r b x_2 + p_o w x_3 \\ -\frac{f}{J} x_5 + \frac{p_o b}{J}\left(x_2 x_3 - x_1 x_4\right) \end{pmatrix} \quad (14)$$

Consequently we define the control in the (α-β) frame as follows :

$$\begin{cases} \begin{cases} u_a = E_0 \text{ si } s_1(x_1) < 0 \\ u_a = -E_0 \text{ si } s_1(x_1) > 0 \end{cases} \\ \begin{cases} u_b = E_0\sqrt{3} \text{ si } s_2(x_2) < 0 \\ u_b = -E_0\sqrt{3} \text{ si } s_2(x_2) > 0 \end{cases} \end{cases} \quad (15)$$

which allows the deduction of the Boolean variables as represented for each sliding surface value in the following table :

	s_1<0& s_2<0	s_1<0&s_2>0	s_1>0& s_2<0	s_1>0&s_2<0
u_1	1	1	0	0
u_2	1	0	1	0
u_3	0	1	0	1

Table 1 : Control for different sliding surfaces value

In the same way as for the previous method, the scheme of the control process can be represented as follows:

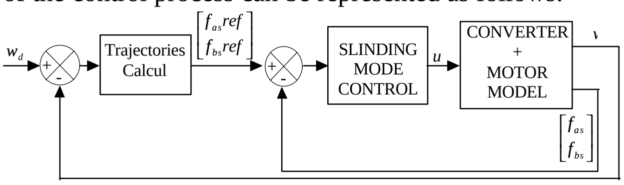

Fig.4:Simulation scheme with Sliding Mode Control

2.3 Pulse Width Modulated

PWM (Pulse Width Modulated) method is frequently used in power electronics and control motor. There are many ways to proceed with. The PWM method the most commonly used compares a reference signal with a high frequency triangular signal. At each instant of equality, the comparator is switched, and consequently the logic state of the signal control is so switched.

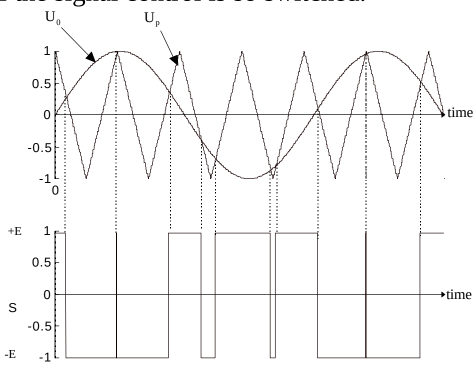

Fig.5:Waveform, one phase.

Figure 5 described the modulation. For the control of the motor, it is necessary to take a controller in order to reduce the steady state error between reference flux and measure flux. The scheme is represented by the following figure :

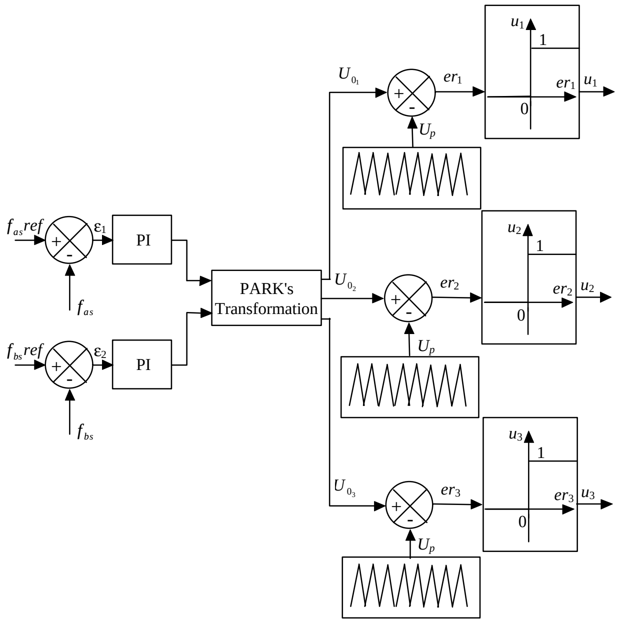

Fig.6 : PWM method

This error should be kept low by choosing a high gain for the PI controllers. For this application the proportional and integral gains are respectively K=1846 and T_i=0.004. The gains are identical for a and b axis.

The control is achieved by the following scheme

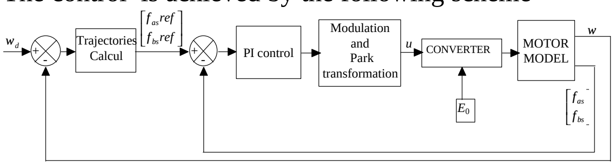

Fig.7 :Simulation scheme with PWM

3 Simulation

The simulation software MATLAB-SIMULINK has been used to study the response of the electrical system. The complete drive was simulated as C-macros, modeling the real time dynamics with a $10ms$ time step. The equations of the complete drive has been solved by using the Runge-Kutta fifth order integration method. The three phased inverter has been simulated considering ideal switches and the asynchronous machine has been simulated by the state equation (1) with the following parameters:

$$\begin{cases} R_s \; : 1.845\ \Omega \quad R_r \; : 1.6\ \Omega \quad M_{sr} \; : 0.227 \\ L_s \; : 236\ mH \quad L_r \; : 236.4\ mH \quad Eo \; : 300\ V \\ f \; : 0.07\ N.s \quad J \; : 0.04\ kg.m^2 \quad p_o \; : 2 \end{cases}$$

The sampling period is chosen according to the system dynamics. Therefore the sampling period for sliding mode control and Boolean control methods is 10^{-4}sec. However the PWM period is a particular case, and it is chosen from the physical limitation of the switch. So the PWM frequency for the modulation signal is 10kHz for the simulation.

3.1 Comparison criterion

A specific speed trajectory is used to study the induction machine behavior on the global speed range figure (8).

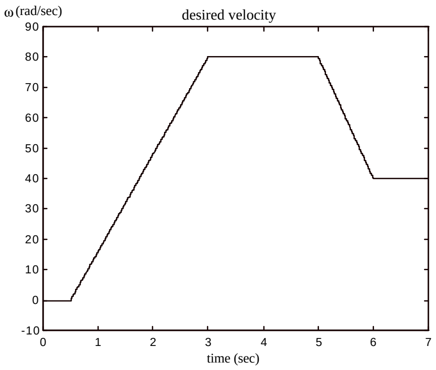

Fig.8 : Desired velocity

As we saw the desired velocity induced the fluxes reference, and so it can be represented in the next figure.

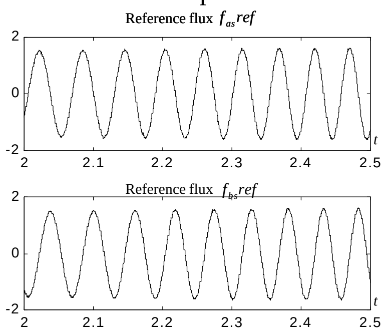

Fig.9 : Fluxes reference.

The measure of the mean absolute error (on speed(rad/s) $\bar{e}_w$ and flux (Wb) $\bar{e}_{f_a}$, $\bar{e}_{f_b}$), maximum error (on speed(rad/s) e_{Mw} and flux(Wb) e_{Mf_a}, e_{Mf_b}), and number of switching (on flux ***nc***) is used for the comparison. In a first case, the normal mode is performed. A robustness test with regards to parameter variations is made in a second study. This robustness test corresponds to an electrical change $\left(\Delta R_s = \Delta R_r = \Delta L_s = \Delta L_r = 10\%\right)$, and a variation on the load torque ($\Delta J = 10\%$). In a last case, the influence of the measure noise effect is tested. Indeed, measures provided by typical sensors are inherently affected by noise, which is generally assumed to be white. The noise magnitude for the flux sensor and speed sensor are respectively supposed equal to 0.01 *Wb* and 0.01 rad/s.

3.2 Simulation results

These different simulation results are summarized as follows in tables 1,2 and 3. For each performing mode, we used the following abbreviation :

NM: Normal Mode.

PV : Parameter Variations.

MNE : Measure Noise Effect.

MODE	NN		PV		MNE	
MEASURE	$\bar{e}_w$	e_{Mw}	$\bar{e}_w$	e_{Mw}	$\bar{e}_w$	e_{Mw}
PWM	0.0958	0.2655	0.0809	0.2330	0.0967	0.2893
SMC	0.1158	0.3360	0.0914	0.2633	0.1175	0.3572
BC	0.1162	0.3408	0.0943	0.2785	0.1165	0.3644

Table 2 : Speed measure for three modes.

MODE	NN		PV		MNE	
MEASURE	$\bar{e}_{f_a}$	$\bar{e}_{f_b}$	$\bar{e}_{f_a}$	$\bar{e}_{f_b}$	$\bar{e}_{f_a}$	$\bar{e}_{f_b}$
PWM	0.0117	0.0115	0.0115	0.0113	0.0209	0.0216
SMC	0.0141	0.02	0.0139	0.0199	0.0193	0.0224
BC	0.0358	0.0364	0.0350	0.0354	0.0355	0.0354

Table 3 : Mean absolute error for the fluxes.

MODE	NN		PV		MNE	
MEASURE	e_{Mf_a}	e_{Mf_b}	e_{Mf_a}	e_{Mf_b}	e_{Mf_a}	e_{Mf_b}
PWM	0.3667	0.0687	0.3667	0.0727	0.3676	0.1623
SMC	0.0980	0.0985	0.0988	0.1009	0.2939	0.1479
BC	0.1551	0.1475	0.1504	0.1406	0.1913	0.2380

Table 4 : Maximum error for the fluxes

MODE	NN	PV	MNE
MEASURE	***nc***	***nc***	***nc***
PWM	279587	279587	279563
SMC	94289	96877	62936
BC	12790	14024	19259

Table 5 : Number of commutations.

As we can see the first control strategy (BC) reduces the number of the commutations in comparison with the others. However the mean absolute errors for flux and speed are much less with the SMC and PWM method. Despite noise measurements the fluxes converge toward the reference flux with a good accuracy. In conclusion on the robustness test, the different control structures have the same performance.

4 Conclusion

A comparison between methods for designing Boolean control law has been studied in this paper. After a short presentation of these different controls, we perform a comparison in simulation on a non linear system, composed of an asynchronous machine associated with a power converter.

The major drawback of the PWM method is in the structure. This structure computes the PI controller or another control without taking into account the converter. The continuous control is then modulated by a triangular signal. So the instant of the commutation is not controlled.

The advantage of SMC and BC method is that the switching is controlled by the sampling period and fixed. At the contrary the PWM method, the switching instant is not regular and depend on the frequency of the triangular signal. Moreover the BC method reduces the number of switching. However the PWM technique is more accurate.

Appendix

The reference flux $f_{as}ref$ and $f_{bs}ref$ are defined by the relations:

$$\begin{cases} f_{as}ref = f_1 \cos(q_s) - f_2 \sin(q_s) \\ f_{bs}ref = f_1 \sin(q_s) + f_2 \cos(q_s) \end{cases} \qquad (16)$$

with f_1 and f_2 fluxes.

q_s angular position between α and β axis.

These flux expressions are based on the Concordia transformation [12]. In fact this transformation matrix is used in electrical machine study in order to achieve axis reference transformation.

Classically in asynchronous machine control, it is necessary to control a magnetization flux f_{ref} which magnetizes the machine. However in this application we have considered that the magnetization flux is fixed and constant $f_1 = f_{ref} = 0.8Wb$

To find q_s we have to go back to the system model in *d-q* reference [12]. In this field oriented frame, one expression of the torque is :

$$C_{em} = p_o \frac{M_{sr}}{L_r}\left(f_{dr} i_{qs} - f_{qr} i_{ds}\right) \qquad (17)$$

As we see, the torque (and consequently the speed) depends on 4 variables. The followed strategy is to find a law with this form $C = kf\,i$ which allows to direct control of the torque by controlling one of the two fluxes. For that, it is sufficient to impose the angular position q_s such that the rotor flux is placed along the *d*-axis at any time, then the flux f_{qr} becomes null. In consequence, the torque equation becomes :

$$C_{em} = p_o \frac{M_{sr}}{L_r}\left(f_{dr} i_{qs}\right) \qquad (18)$$

So by using the model in *d-q* reference and equation (18), we can deduce the derivative of the angular position as follows :

$$\frac{dq_s}{dt} = pow_d + \frac{R_r}{pof_{ref}^{\;2}} C_{em} \qquad (19)$$

According to the fundamental equation of the dynamics for the torque, the angular position become :

$$q_s = \int_0^t pow_d + \frac{R_r}{pof_{ref}^{\;2}}\left(J\frac{dw}{dt} + fw\right) \qquad (20)$$

This algorithm gives the angular position q_s from the measures of the velocity ω.

By now let us consider the flux f_2 define in (16). **This flux is proportional to the torque to impose. Therefore it will be used to tune the dynamics and time response for the motor**. This tuning is performed by a control loop and a controller. The corrector is simply achieved by a Proportional Integral (PI) controller. This compensator allows the system to attain satisfactory performances. The gain of this PI is chosen in order for the response to be fast and with a low overshoot for the nominal conditions. This controller reducing the error between the reference signal ω_d and the process output ω, is expressed as :

$$\left\{ f_2 = K\left(w_d - w\right) + \frac{1}{T_i}\int_0^t \left(w_d - w\right) dt \right. \qquad (21)$$

For the considered application, the PI controller gains are chosen as follows : K = 1.5 and T_{i1} = 0.2. In summarized the computation of the fluxes reference is defined by the following scheme :

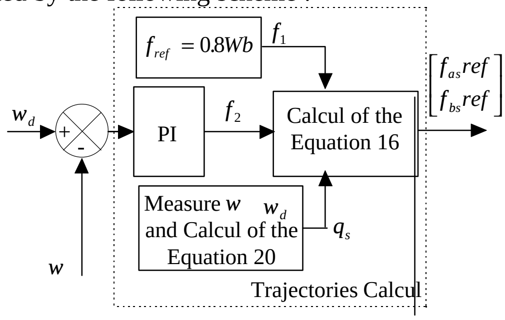

Fig.10: Representation of fluxes reference trajectories

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