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Fault-Tolerant Control of 5-Phase PMSG for Marine Current Turbine Applications based on Fractional Controller

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Abstract: For renewable energy generation and fault tolerant applications, a multiphase PMSG seems to be an interesting solution. In this paper, a fault-tolerant Control of a 5-phase PMSG Non-Sinusoidal EMF for Marine Current Turbine applications under fault operation is proposed. In order to maximize the average torque, to minimize the copper losses and to reduce the torque ripples all harmonics of EMF are exploited under fault opearation. As the optimal current references in the Concordia's frame are not constants, the chosen regulator must have a high dynamic performance even if the machine speed varies. It must ensure reference tracking without affecting the signal magnitudes or introduce phase shift. Hence, a Fractional controller is investigated to control the non-constant current references in the Concordia's frame under fault operation.

Keywords: Fault-Tolerant Control, Fractional Controller, Five-Phase PMSG, Marine Current Turbine.

1. INTRODUCTION

In the context of renewable energy source exploitation and in the case of fault tolerant applications, a multiphase Permanent magnet synchronous generator (PMSG) seems to be an interesting solution. The association of multiphase magnet synchronous generator to AC-DC converter can increase the converted power, segmenting the electrical power and increase the operating of the energy conversion chain. Indeed, this association can operate under fault condition as one opening phase. This paper deals with a 5-phase PMSG non-sinusoidal EMF for Marine Current Turbine Applications under fault operation. The key idea is to exploit all the current harmonics and EMF harmonics which can contribute positively to the optimal power transfer. A fault-tolerant control under fault condition is proposed and it's based on Fractional Controller. Multiphase PMSG for Marine Current Turbine Applications has been subject of studies in (S. Benelghali and al. 2011)-(F. Mekri and al. 2013). But the control strategies proposed in these papers is done in the dq Park frame because the dq current references are constants and a simple PI controller is sufficient for a good control. Nevertheless, in the case of open-circuit fault torque ripples appear. The torque control strategy need to impose optimal current references to reduce the torque ripples and minimize the copper losses. These current references are not constants in the dq frame as highlighted in (H.-M. Ryu and al. 2006)-(B. Sari and al. 2012). Then a robust controller is needed as proposed in (F. Mekri and al. 2013)-(B. Sari and al. 2012). As the new current references are not constants in the dq frames, it is not necessary to develop a dynamical model in view of control of the 5-phase PMSG in the Park frame. An adequate model in the natural base or Concordia's frame is sufficient. In all cases a controller which has a large bandwidth is needed. In this paper, a Fractional controller is investigated. Fig. 1 shows the control scheme of the 5-Phase Permanent Magnet Synchronous Generator under Fault Operation. This work focuses on the association of five-phase PMSG to AC/DC converter under fault condition and the performances achieved thanks to Fractional controller. The first section presents the electrical model of the 5-phase PMSG in the Concordia's frame under fault operation. Then a control strategy of the 5-phase PMSG is introduced. The next section presents the Fractional controller. A synthesis methodology, to determinate its parameters, is given. Finally, the simulation results highlight the performances of the Fractional controller.

2. ELECTRICAL MODEL OF THE 5-PHASE PMSG

The model of the 5-phase PMSG is established using the following assumptions: all phases are identical and regularly shifted by an angle $\alpha=\frac{2\pi}{5}$. The effect of the saturation is neglected and the considered machine is not salient. The homopolar EMF is equal to zero. Under normal operation, in the natural base, the model is simple and it is given by:

$$[E] = [R][I] + [L]\frac{d}{dt}[I] + [V] \tag{1}$$

$$\text{Where}[R] = \begin{bmatrix} r & 0 & 0 & 0 & 0 \\ 0 & r & 0 & 0 & 0 \\ 0 & 0 & r & 0 & 0 \\ 0 & 0 & 0 & r & 0 \\ 0 & 0 & 0 & 0 & r \end{bmatrix}, L = \begin{bmatrix} L_1 & L_2 & L_3 & L_3 & L_2 \\ L_2 & L_1 & L_2 & L_3 & L_3 \\ L_3 & L_2 & L_1 & L_2 & L_3 \\ L_3 & L_3 & L_2 & L_1 & L_2 \\ L_2 & L_3 & L_3 & L_2 & L_1 \end{bmatrix}$$

$$[X] = \begin{bmatrix} X_a \\ X_b \\ X_c \\ X_d \\ X_e \end{bmatrix} \text{And X= V (voltage), I (current), E (EMF).}$$

Under fault operation corresponding for example to the opening of the fifth phase, in the natural base, the model becomes:

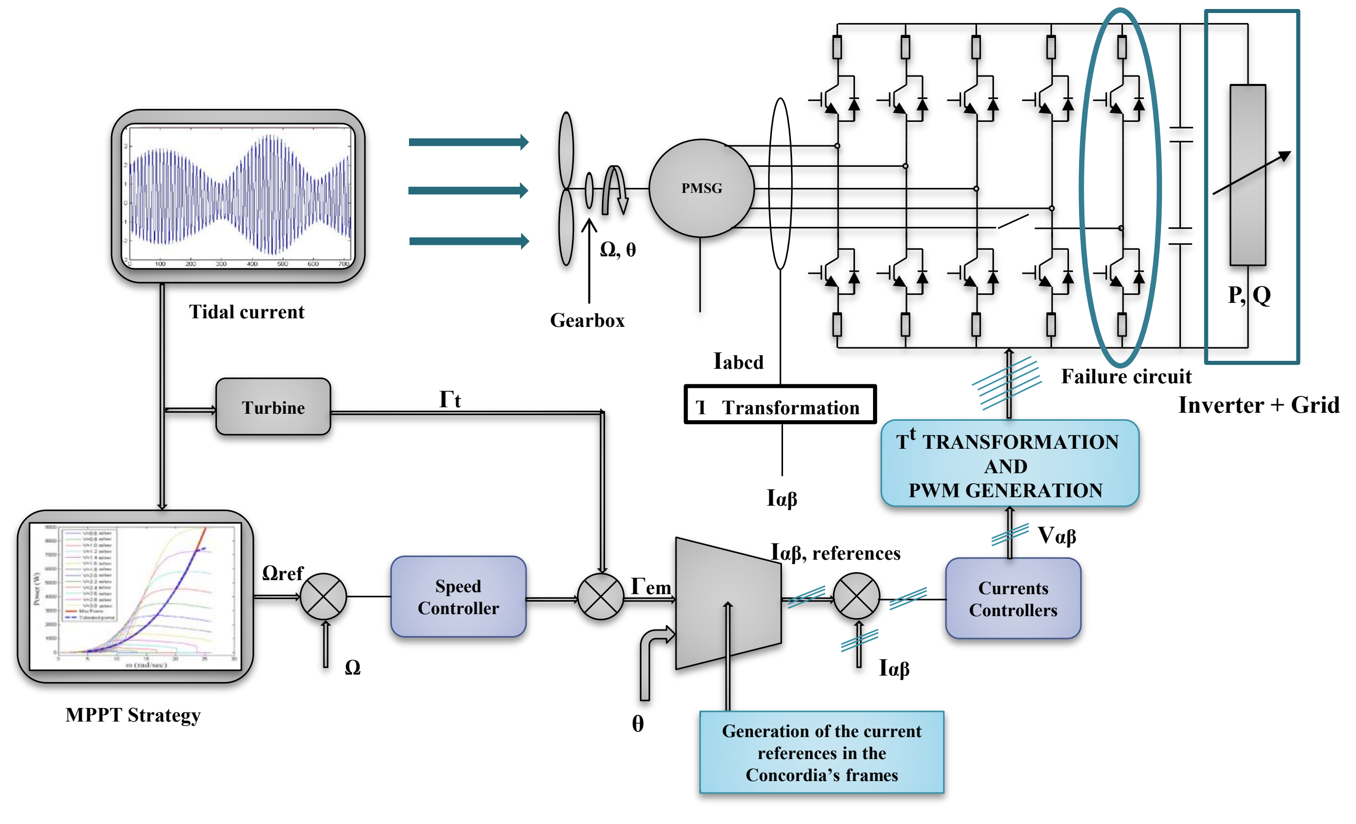


Fig. 1. Control scheme of the 5-Phase Permanent Magnet Synchronous Generator under Fault Operation

$$[E'] = [R][I] + [L']\frac{d}{dt}[I] + [V'] \quad (2)$$

$$\text{Where}[R] = \begin{bmatrix} r & 0 & 0 & 0 \\ 0 & r & 0 & 0 \\ 0 & 0 & r & 0 \\ 0 & 0 & 0 & r \end{bmatrix},\ [E'] = \begin{bmatrix} E'_a \\ E'_b \\ E'_c \\ E'_d \end{bmatrix} = \begin{bmatrix} E_a + 0.25E_e \\ E_b + 0.25E_e \\ E_c + 0.25E_e \\ E_d + 0.25E_e \end{bmatrix}$$

$$L' = \begin{bmatrix} L_1 + 0.25L_2 & L_2 + 0.25L_3 & L_3 + 0.25L_3 & L_3 + 0.25L_2 \\ L_2 + 0.25L_2 & L_1 + 0.25L_3 & L_2 + 0.25L_3 & L_3 + 0.25L_2 \\ L_3 + 0.25L_2 & L_2 + 0.25L_3 & L_1 + 0.25L_3 & L_2 + 0.25L_2 \\ L_3 + 0.25L_2 & L_3 + 0.25L_3 & L_2 + 0.25L_3 & L_1 + 0.25L_2 \end{bmatrix}$$

Nevertheless the model in view of control, in the natural base, under fault condition is not adapted because of the magnetic coupling between the phases. Therefore it is necessary to develop a model where no magnetic coupling between the phases appears. The following transformation matrix is used to diagonalize the inductance matrix L'.

$$[T] = \sqrt{\frac{2}{5}} \begin{bmatrix} \sqrt{2}\cos\left(\frac{2\pi}{5}\right) + \frac{\sqrt{2}}{4} & \sqrt{2}\cos\left(\frac{4\pi}{5}\right) + \frac{\sqrt{2}}{4} & \sqrt{2}\cos\left(\frac{6\pi}{5}\right) + \frac{\sqrt{2}}{4} & \sqrt{2}\cos\left(\frac{8\pi}{5}\right) + \frac{\sqrt{2}}{4} \\ \sin\left(\frac{2\pi}{5}\right) & \sin\left(\frac{4\pi}{5}\right) & \sin\left(\frac{6\pi}{5}\right) & \sin\left(\frac{8\pi}{5}\right) \\ \sin\left(\frac{6\pi}{5}\right) & \sin\left(\frac{2\pi}{5}\right) & \sin\left(\frac{8\pi}{5}\right) & \sin\left(\frac{4\pi}{5}\right) \\ \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{bmatrix}$$

(3)

This transformation allows us to work in orthogonal frame but it is not normed unlike the transformation matrix under normal operation.

The transformation matrix, under fault operation, cannot satisfy the pseudo-orthogonal property as highlighted in (H.-M. Ryu and al. 2006). Another transformation matrix is proposed in (H.-M. Ryu and al. 2006). Using the transformation matrix, the model in the Concordia's frame can be written:

$$[E']_y = r[I]_y + L_y\frac{d}{dt}[I]_y + [V']_y \quad (4)$$

Where y=α_m , β_m , β_s , h

The homopolar EMF is equal to zero and the neutral of the machine is not connected and therefore the homopolar current is equal to zero.

3. CONTROL STRATEGY OF THE 5-PHASE PMSG UNDER FAULT OPERATION

The MPPT strategy is applied. The maximum power available from the tidal current is captured and injected to the grid. The power transfer is optimal when the losses are minimal. In this case an optimal control strategy that minimizes the reactive power and the copper losses is proposed.

The reactive power becomes null when the EMF and current vectors of the machine are collinear:

$$\frac{E'_a}{I_a} = \frac{E'_b}{I_b} = \cdots = \frac{E'_d}{I_d} \quad (5)$$

The optimal current references can be written:

$$I_z = \frac{E'_z}{\sum_{z=a}^{d} {E'_z}^2}\Gamma_{ref}\Omega \quad (6)$$

Where z = a, b, c, d.

Using the transformation matrix under fault operation the optimal current reference can be obtained:

$$[I_{\alpha\beta}] = [T][I_{abcd}] \quad (7)$$

4. FRACTIONAL CONTROLLER (A. Dieng and al. 2012)

In the Concordia's frame the current references are not constants. The controller must ensure reference tracking without affecting the signal magnitudes or introducing phase shift for any operating speed. Hence the controller must have a high dynamic performance. To achieve this goal we compare the performances of the PI controller and the proposed fractional controller.

The PID controller is undoubtedly the most commonly used control law algorithm in the control industry (K. J. Astrom and al. 2005). The fractional order PID controller is the expansion of conventional PID controller based on fractional calculus (K. A. Tehrani and al. 2010). The general form of fractional order PID controller is $PI^{\alpha}D^{\beta}$ and its general transfer function is given by:

$$C(s) = K_P + \frac{K_I}{s^{\alpha}} + K_D s^{\beta} \quad (8)$$

The fractional controller has two more adjustable parameters than traditional PID controller, and they are differential order β and integral order α .Therefore, the design of fractional order $PI^{\alpha}D^{\beta}$ controller is more flexible (R. Gong and al. 2010).

Also, only a fractional order PI controller: PI^{α}can be used. The tuning of PI^{α} controller requires the determination of three parameters (K_P, K_I, and α) and it's done from a given parametric state of the process to be controlled.

In recent years, several methods have been proposed by researchers (I. Podlubny. 1999)-(A. Charef. 2006)-(R. Gong and al. 2010)-(A. Charef and al. 1992)-(C. Yeroglu and al. 2010)-(Y. J. Cao and al. 2006)-(K. A. Tehrani and al. 2010) for the determination of these parameters. From the point of view of performance, all methods are equal, but according to the process to be controlled it is important to fix some objectives. In order to calculate the parameters of PI^{α} controller two methods are proposed:

- The first one is an analytical one and is based on the ideal Bode transfer function (H. W. Bode. 1945),
- The second one uses optimization techniques (particles swarms in our case) (Y. J. Cao and al. 2006).

4.1 Analytic method

The proposed method is addressed to the first order system. The controller parameters are determined using the ideal Bode transfer function which transfer function in open loop is defined by the following fractional order integrator:

$$T(s) = \frac{A}{s^{\beta}} \text{avec } 1 < \beta < 2 \quad (9)$$

The desired phase margin Φm of the loop is fixed constant for all values of gain:

$$\Phi_m = \pi\left(1 - \frac{\beta}{2}\right) \quad (10)$$

The parameters K_P and K_I are firstly determined with α=0 and using classical synthesis methods (pole compensation method, pole placement method…). Then, keeping the values of K_P and K_I and depending on desired performance, α is calculated so that the asymptotic behavior of the overall system in open loop is equivalent to that of the ideal Bode transfer function in a desired waveband:

$$G(s) = \left(K_P + \frac{K_I}{s^{\alpha}}\right)\left(\frac{K}{1+\tau_{bo}s}\right) \approx \frac{A}{s^{\beta}} \quad (11)$$

In this waveband the asymptotic behavior of the fractional controller depends on α and can be written as follows:

$$C(s) = K_P + \frac{K_I}{s^{\alpha}} \approx \frac{K_I}{s^{\alpha}} \quad (12)$$

If in this waveband, the order of asymptotic behavior of the system to be controlled equals to zero, the loop transfer is:

$$G(s) = \frac{K_I K}{s^{\alpha}} \approx \frac{A}{s^{\beta}} \quad (13)$$

By identificationα=β and A=K_IK, as 1<β<2 therefore 1<α<2.

In the opposite case (the order of asymptotic behavior is equal to 1), the transfer function becomes:

$$G(s) = \frac{K_I}{s^{\alpha}}\frac{K}{\tau_{bo}s} \approx \frac{A}{s^{\beta}} \quad (14)$$

By identification α=β-1 and A=$\frac{K_I K}{\tau_{bo}}$, as 1<β<2 therefore 0<α<1.

β is calculated from Eq. 10 by imposing the desired phase margin of the loop:

$$\beta = 2\left(1 - {}^{\Phi m}/_{\pi}\right) \quad (15)$$

Fractional controller can be written as follows:

$$C(s) = K_P + \frac{K_I}{s^{\alpha}} = K_P\left(1 + \frac{1}{\left({}^{s}/_{\omega_N}\right)^{\alpha}}\right) \quad (16)$$

where $\omega_N = \left(\frac{K_I}{K_P}\right)^{1/\alpha}$.

First case 0 <α<1;

For frequencies ω<ω_N, in the desired waveband, the integral effect is dominant then:

$$C(s) = K_P + \frac{K_I}{s^{\alpha}} \approx \frac{K_I}{s^{\alpha}} \quad (17)$$

Otherwise, the proportional effect is dominant. The error will be minimal and the settling time will be better if K_P is high.

Second case 1<α <2; the same observation is done.

Whatever the behavior of the system, if the condition ω<ω_N is not checked, the proportional effect is dominant.

Nevertheless, the value of α will be preserved but it will be necessary to increase the proportional gain to cancel the error and to obtain better settling time. For that, the condition imposed on K_p is (R. Gong and al. 2010):

$$K_P \geq \left(\left({}^{100}/_{\varepsilon}\right) - a_0\right) \quad (18)$$

Where ε is the desired error and a_0 the coefficient of the monomial of smaller degrees of the denominator of the system to be controlled.

4.2 Particle Swarm Optimization

Particle Swarm Optimization (PSO) is a new population-based evolutionary computation (Y. J. Cao and al. 2006). Unlike genetic algorithms, the PSO updates populations without any genetic operators such as crossover and mutation (Y. J. Cao and al. 2006). The PSO algorithm attempts to mimic the natural process of group communication of individual knowledge, which occurs when such swarms flock, migrate, forage, etc., in order to achieve some optimum property such as configuration or location (Y. J. Cao and al. 2006).

For conditioning the performance of optimization it is necessary to define the objective function and the constraints related to this one.

First of all, one fixed the optimization goal, and then encodes the parameters to be searched. PSO algorithm is running until the stop condition is satisfied (Y. J. Cao and al. 2006). Knowing the transfer function of PI^{α} (Eq. 16) and based on the approximation $s^{\alpha} = \omega^{\alpha}\left(\cos\frac{\pi}{2}\alpha + j\sin\frac{\pi}{2}\alpha\right)$(C. Yeroglu and al. 2010), the overall transfer function of closed loop system is deduced:

$$G_F(s) = \frac{N(s)}{D(s)} \tag{19}$$

Where

$$N(j\omega) = K_I K + \omega^{\alpha} K_P K \cos\frac{\pi}{2}\alpha + j\omega^{\alpha} K_P K \sin\frac{\pi}{2}\alpha \tag{20}$$

$$D(j\omega) = A(j\omega) + B(j\omega) \tag{21}$$

$$A(j\omega) = \omega^{\alpha}\cos\frac{\pi}{2}\alpha - \tau_{bo}\omega^{\alpha+1}\sin\frac{\pi}{2}\alpha + K_I + \omega^{\alpha}K_P K\cos\frac{\pi}{2}\alpha \tag{22}$$

$$B(j\omega) = j\left(\omega^{\alpha}\sin\frac{\pi}{2}\alpha + \tau_{bo}\omega^{\alpha+1}\cos\frac{\pi}{2}\alpha + \omega^{\alpha}K_P K\sin\frac{\pi}{2}\alpha\right) \tag{23}$$

The optimization goal is to find the three parameters K_P, K_I, α such that the modulus of $G_F(j\omega)$ is unitary ($G_{desired}$) and the argument of $G_F(j\omega)$ is zero ($\varphi_{desired}$) while minimizing the following objective function:

$$\sqrt{(G_{desired} - G)^2 + (\varphi_{desired} - \varphi)^2}$$

4.3 Approximation of the operator fractional integrator

Once the regulator is calculated, the implementation of PI^{α} controller requires the approximation of the fractional integrator operator. One of the well-known continuous approximation approaches is called Charef (A. Charef. 2006)-(A. Charef and al. 1992). The fractional-order integrator $s^{-\alpha}(0 < \alpha < 1)$ was modeled by a fractional power pole (FPP) (A. Charef. 2006) in a given waveband. Next, this FPP is approximated by a rational function, using the method given in (A. Charef. 2006).

In a given waveband $[\omega_b\,,\ \omega_h]$, this fractional-order operator can be modeled by an FPP whose transfer function is given as follows (A. Charef. 2006)-(A. Charef and al. 1992):

$$G_I(s) = \frac{k_i}{\left(1+\frac{s}{\omega_c}\right)^{\alpha}} \tag{24}$$

Suppose that $\omega \in [\omega_b\,,\ \omega_h]$ and $\omega \gg \omega_c$, Eq. 24 becomes:

$$G_I(s) = \frac{k_i}{\left(\frac{s}{\omega_c}\right)^{\alpha}} = \frac{k_i {\omega_c}^{\alpha}}{s^{\alpha}} = \frac{1}{s^{\alpha}} \tag{25}$$

With $k_i = \frac{1}{{\omega_c}^{\alpha}}$ and ω_c is the -3dB cutoff frequency of the FPP, which is obtained from the low frequency ω_b, $\omega_c = \omega_b\sqrt{10^{\frac{\gamma}{10\alpha}} - 1}$, where γ is the maximum permitted error between the slopes of the fractional order integrator and the FPP (Eq. 24) in the given waveband $[\omega_b\,,\ \omega_h]$ (A. Charef. 2006)-(A. Charef and al. 1992).

Hence, the approximation is given by (A. Charef. 2006)-(A. Charef and al. 1992):

$$G_I(s) = \frac{k_i}{\left(1+\frac{s}{\omega_c}\right)^{\alpha}} \simeq K_I \frac{\prod_{i=0}^{N-1}\left(1+\frac{s}{z_i}\right)}{\prod_{i=0}^{N}\left(1+\frac{s}{p_i}\right)} \tag{26}$$

Using a simple graphical method that began with a specified error ε in decibels and a frequency ω_{max} which is equal to $100\omega_h$ (A. Charef. 2006)-(A. Charef and al. 1992), the parameters z_i, p_i, and N can be calculated as (A. Charef and al. 1992):

$$N = \text{Integer}\left\{\frac{\log\left(\frac{\omega_{max}}{p_0}\right)}{\log(bc)}\right\} + 1 \tag{27}$$

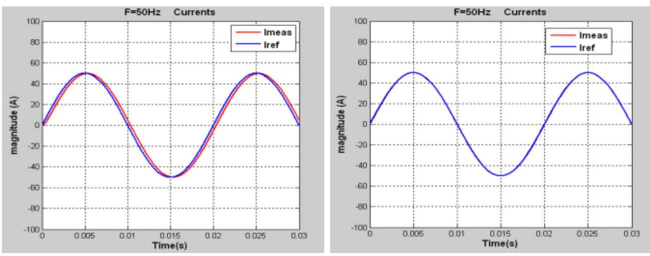
$$\begin{Bmatrix} p_i = (bc)^i p_0, \quad i = 0,1,\dots,N \\ z_i = (bc)^i b p_0, \quad i = 0,1,\dots,N-1 \\ b = 10^{\frac{\varepsilon}{10(1-\alpha)}}, \quad c = 10^{\frac{\varepsilon}{10\alpha}} \\ p_0 = \omega_c\sqrt{c}, \quad z_0 = b p_0 \end{Bmatrix} \tag{28}$$

In our case, by imposing ω_b=100π rads/s ,ω_h=1000π rads/s and by solving Eqs. 27 and 28 the parameters of the fractional controller (Eq. 26) are obtained:

N=3, b=1.5849, c=1.5849.

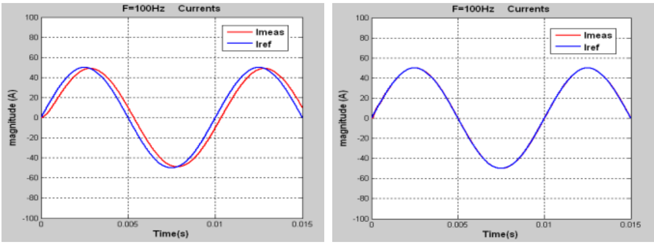
5. RESULTS

To show the performances of the fractional controller comparing to the classic PI controller, the load current of single phase inverter is regulated based on the two approaches. The range of frequency variation is [50Hz-100Hz]. Figs. 2, 3 and 4 present the results obtained.



(a) Classic PI (b) Fractional Controller

Fig. 2. Load Currents, F=50Hz



(a) Classic PI (b) Fractional Controller

Fig. 3. Load Currents, F=100Hz

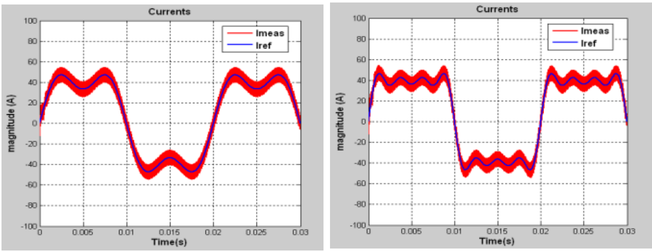


Fig. 4. Load Currents, Fractional controller

For the classic PI (its parameters are constants), if the frequency varies, then the phase shift between the reference signal and the output signal increase (Figs. 2(a) and 3(a)). This is not the case of the fractional controller which maintains this phase shift equal to zero in a large frequency range (Figs 2(b), 3(b) and 4).

Now the fractional controller synthetized in section 4 is applied to control the current of 5-phase PMSG non-sinusoidal EMF under the fifth phase open-circuit fault.

Matlab Simulink is used for the simulation. For the marine current turbine simulator, the model in (S. Benelghali and al. 2011) is used. The switching frequency of the rectifier is fixed and equal to 10 kHz. To better show the performances achieved by the inner current control loop it is only interested in an operating point. In this case the tidal current is supposed constant.

Figs. 9, 10, 11 and 12 show that the performances of the fractional controller to control the non-constant current references under fault operation. No disturbances have been observed in the generator torque, the generator speed, the generator power, the DC bus voltage (Figs. 5, 6, 7 and 8).

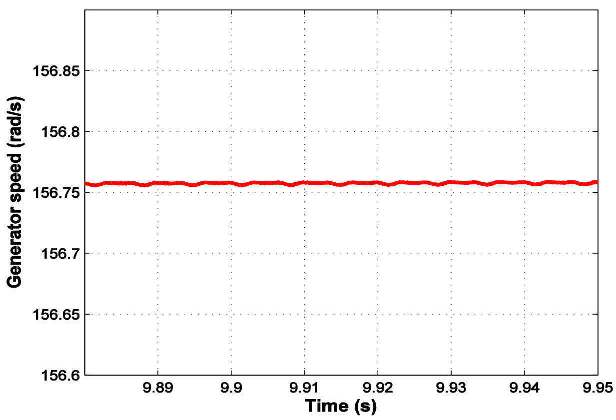


Fig. 5. Generator Speed under fault operation

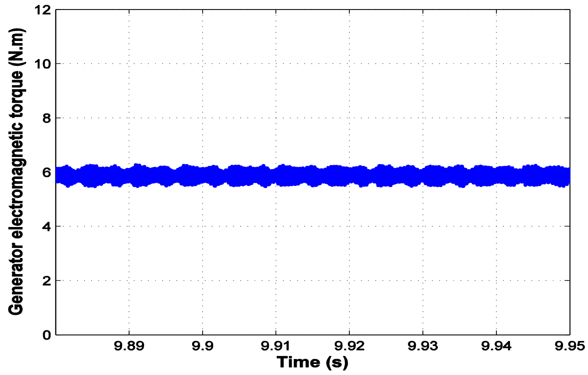


Fig. 6. Generator Torque under fault operation

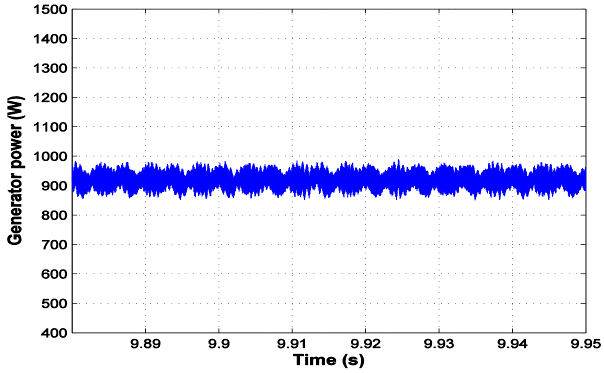


Fig. 7. Generator power under fault operation

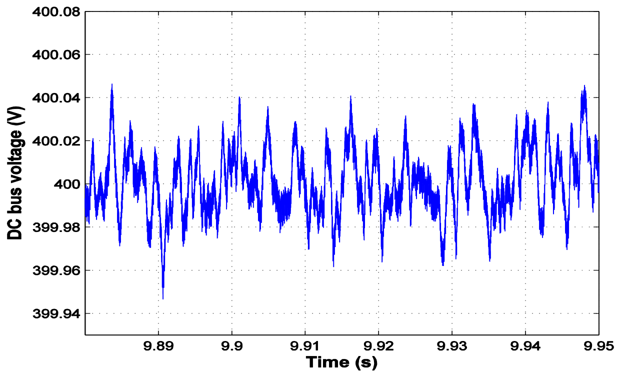


Fig. 8. DC bus voltage under fault operation

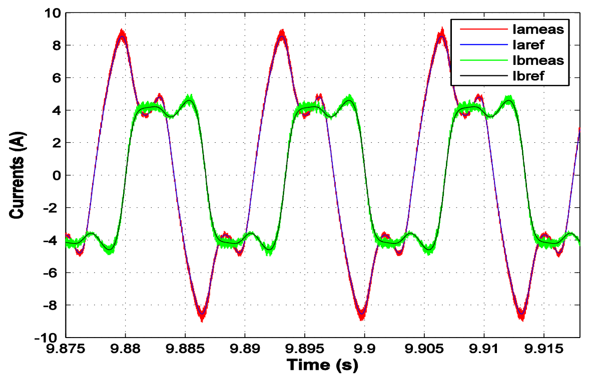


Fig. 9. Currents in the phase a and b under fault operation

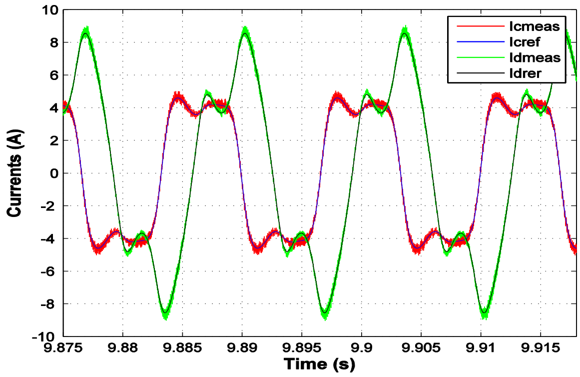


Fig.10. Currents in the phase c and d under fault operation

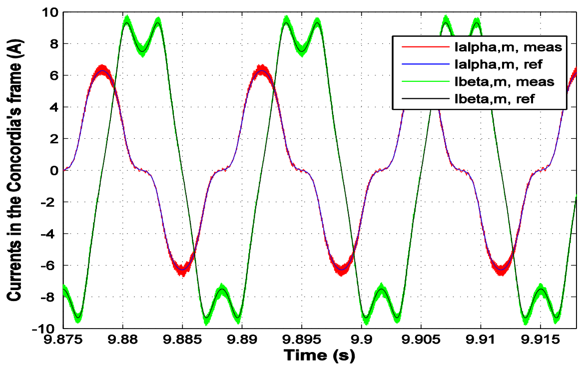


Fig.11. Currents in α_m β_m frame under fault operation

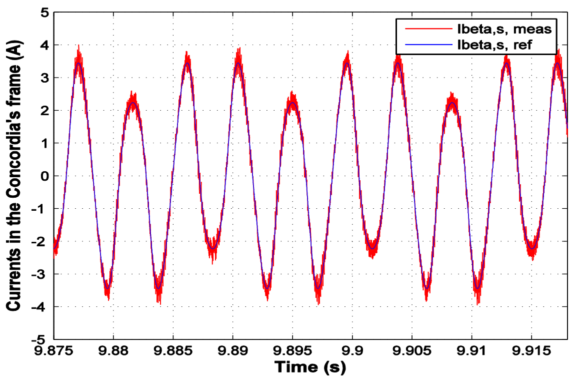


Fig.12. Currents in β_s frame under fault operation

6. CONCLUSIONS

In this paper, a fault-tolerant Control of 5-Phase PMSG under Fault Operation in the Concordia's frame is proposed. The relevance of Fractional controller to control the non-constant current references has been presented. Simulation results prove the effectiveness of the fractional controller as regards transient performances.

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