A Novel Current Control Strategy for PWM Inverters using the Sliding Mode Techniques

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Abstract— Classical PWM Techniques are replaced by Current Source Sliding Mode Control (SMC) deciding the IGBT’s status in real time by using a FPGA. Utility synchronization and inverter management is guaranteed by Park Vector Transformation. Three novel sliding strategies for Current Controlled PWM Inverters are described, and several simulations show the performance of the inverter control.

I. INTRODUCTION

Pulse Width Modulation PWM techniques have been studied for the last decades [1]. Classical methods allow a simple and economic control of inverters, but it does not guarantee high performance because of, for example, load variations.

Current Source PWM Inverters are widely applied in the control of three-phase asynchronous machines, rectifiers, and as power conditioners of DC Energy Sources in parallel with the utility. These systems behaves nonlinearly and many techniques have been developed for handling it. Park Vector Transformation and the Sliding Mode Control are used in this paper [2], [3]. The former guarantees the analysis of the system almost linearly, while the latter, being part of a Variable Structure Control (VSC), produces a better transient response than PWM methods [4], and is insensitive to parameter variations and load disturbances [5].

The load will be the mains, and since it behaves as a voltage source, the best strategy is to control the inverter in Current Source Mode. An Utility Observer obtains its voltage, frequency and phase for synchronization.

II. INVERTER LAYOUT

The inverter structure is the usual in this kind of power conditioners: the DC Energy Source connected to the inverter through a link capacitor, a LC filter, and a transformer to isolate the equipment from the utility. See figure 1.

III. SYSTEM EQUATIONS

In order to control the inverter in a Current Source Mode, it must be studied the relationship between $U_{dc}$ and $U_{ac}$ in the inductances:

$$u_{uv} - u_{rs} = L \frac{d(i_u - i_v)}{dt} = L \frac{d i_{uv}}{dt}$$
$$u_{vw} - u_{sf} = L \frac{d(i_v - i_w)}{dt} = L \frac{d i_{vw}}{dt}$$
$$u_{wu} - u_{tr} = L \frac{d(i_w - i_u)}{dt} = L \frac{d i_{wu}}{dt}$$

(1)

The control of the IGBT’s gives different values in $U_{dc}$ $(u_{uv}, u_{vw}, u_{wu})$, which means different evolutions in the inverter currents $I$ $(i_u, i_v, i_w)$. As the currents are known, measured in real time, it can be decided the best combination for the IGBT’s status that guides the real currents to the desired sinusoidal references.

From the above equations there are only two freedom degrees, as the sum of voltages and currents is zero. Park Vector Transformation (2) [5], also known as “Field Orientation” in Motor Drive environment, can be applied to handle this kind of variables:

$$\overline{X}(t) = \frac{2}{3} (x_1 + \overline{a} \cdot x_2 + \overline{a}^2 \cdot x_3)$$

(2)

where:
The vectorial representation of $x_1, x_2, x_3$ leads to the vector $X$ that rotates in an anticlock-wise sense. This vector can be expressed in the complex plane (Park space) in terms of three fixed axes: $a, b, c$, each one rotated 120° to the other, two fixed axes: $\text{Re}(\alpha)$ and $\text{Im}(\beta)$, and/or two moving axes: $d, q$ (see figure 2, and (4) and (5)).

\[
\begin{bmatrix}
X_\alpha \\
X_\beta \\
0
\end{bmatrix} = \begin{bmatrix}
\frac{2}{3} & -\frac{1}{3} & -\frac{1}{3} \\
0 & \frac{\sqrt{3}}{3} & -\frac{\sqrt{3}}{3} \\
\frac{1}{3} & \frac{1}{3} & \frac{1}{3}
\end{bmatrix} \begin{bmatrix}
x_1 \\
x_2 \\
x_3
\end{bmatrix}
\]  \hspace{1cm} (4)

\[
\begin{bmatrix}
X_d \\
X_q
\end{bmatrix} = \begin{bmatrix}
\cos \theta & \sin \theta \\
-\sin \theta & \cos \theta
\end{bmatrix} \begin{bmatrix}
X_\alpha \\
X_\beta
\end{bmatrix}
\]  \hspace{1cm} (5)

IV. INVERTER REGULATION AND CONTROL

There are only eight possible combinations in the IGBT’s state (000, 001, 010, 011, 100, 101, 110, 111). Each one imposes different values in the inverter voltages and so, different evolutions in the current vector. The error between the measured currents and the given sinusoidal references in each phase is defined as:

\[e_{1,2,3} = i_{uv,vw,wu} - i^*_{uv,vw,wu} \]  \hspace{1cm} (6)

Transforming this error into the Park Space and expressing it by the three fixed axes $a, b, c$, as said above, leads to $e_{a,b,c}$.

Dividing the Park space into six control intervals, the transformed error in each axis is then compared with a hysteresis band (HB) according to the three strategies shown in figure 3. Therefore, the IGBT’s status can be selected in each phase independently [6].

In order to adjust a constant PWM frequency, a PI regulator modifies the hysteresis band (HB) according with the measured PWM frequency and the desired mean value [4].

This hysteresis band leads to an hexagon in which the evolution of currents is free, depending only on the previous state [7]. Outside the hexagon the inverter imposes the direction displayed in figure 4, assuring convergence and stability.

Transforming the sinusoidal utility voltages into Park Vector Space ($U$) and adjusting the moving $d$ axis with the first phase ($U_q = 0$) guarantee inverter-utility synchronization [4]. In this paradigm, transforming the sinusoidal inverter currents into the vector $I$, active power corresponds with $I_d$ and the reactive one with $I_q$ (7) [8]. The second component is regulated to zero or to a constant to obtain stable power factor. Meanwhile, the $d$ component is used to regulate the DC voltage to extract the maximum instantaneous power available in the DC energy source. See figure 5.

\[
P = \frac{1}{2}(U_d \cdot I_d + U_q \cdot I_q) = \frac{1}{2}U \cdot I_d
\]

\[
Q = \frac{1}{2}(U_q \cdot I_d - U_d \cdot I_q) = -\frac{1}{2}U \cdot I_q
\]  \hspace{1cm} (7)
Fig. 5. Block Diagram of the Inverter Control using Sliding.

Here, two ideal current sources simulate the current control carried out in the Inverter, while the DC Energy Source and the link capacitor are replaced by an ideal voltage source for simplicity.

V. SIMULATIONS

To show the performance of the proposed strategy, using Sliding in Current Source Mode, several simulations have been made. Following it is considered an inverter connected in parallel with the utility and switching at 10 kHz.

First, figure 6 displays the evolution of the inverter and utility magnitudes during the first 160 ms. Connection is made at 16 ms and a -50 A step in $I_q$ is made at 80 ms which leads to a lower power factor, as can be seen in the figure. A sudden disconnection of the inverter from the utility and the DC Energy Source is made at about 145 ms. It can be seen that phase to phase voltages lay down to zero almost instantaneously, as well as the active power transferred to the utility.

The current in the q axis is represented in the upper zone. Next, the phase to phase voltage and current at the inverter, as well as its reference, the phase to phase voltage in the LC filter, and also the voltage and current of the utility, all of them for the first phase. It must be pointed out that the sliding technique proposed in this paper obtains near three level outputs, but this law is broken whenever is needed. Finally, the values of active and reactive powers, and the regulated value of the PWM frequency are shown. It can be seen a positive value in the reactive power after 80 ms due to the -50 A step in $I_q$, which corresponds to the lower power factor as said before.

Figure 7 shows the resulting situation in Park Vector Space for the direct sliding strategy, while figure 8 shows the same situation for the optimal one.

The inverter (slightly filtered), capacitor and utility voltages are represented at left, whereas the utility and inverter currents are at right: utility currents are smooth while the inverter ones are noisy due to their switching nature. With more time, the trajectories will conform circumferences.
Fig. 8. Resulting situation in Park Vector Space for the optimal sliding strategy.

Fig. 9. Evolution of the controlled current error vector for the direct sliding strategy.

On the other hand, figures 9 and 10 reflect the evolution of the controlled current error vector, using the direct and the optimal sliding strategies. Here it can be seen six of the eight possible combinations mentioned above, while the other two: V7 (111) and V8 (000) are at the origin and impose no force on the I vector evolution. For the direct strategy the current error vector rotates in a clock-wise sense, while for the optimal one it converges to the origin of coordinates.

Fig. 10. Evolution of the controlled current error vector for the optimal sliding strategy.

Note that the current error vector double the hysteresis band due to the worst case when the vectors V7 (111) or V8 (000) are imposed [7].

Figures 7, 8, 9 and 10 show that the optimal sliding strategy has a less harmonic distortion than the direct one. This situation arises from the fact that when the current error in one phase exceeds the hysteresis band there is a stronger force, in the opposite direction to this error, than in the direct case. This force will hold the current error vector closer to the hexagon. That is, choosing the optimal sliding strategy leads to a reduced value in the inductances for the same current ripple.

Finally, it can be said that the inverse sliding strategy has a similar behavior than the direct one, but in this case the current error vector rotates in the opposite direction.

VI. CONCLUSIONS

The whole inverter is controlled in Current Source Mode while synchronization is achieved by using an Utility Observer in Park Vector Space. Though a Vectorial Control Strategy is used, each phase is controlled jointly with Sliding.

Park vector is used to control the IGBT’s; meanwhile a PI regulator can be used to assure the power transfer from the DC Energy Source. Near unitary power factor is always obtained.

The optimal sliding strategy reduces the harmonic distortion of the inverter currents in comparison with the direct one.

REFERENCES