Control of a three-phase four-wire active filter operating with an open phase

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Abstract: This paper investigates the utilization of a power compensation control scheme that provides fault-tolerant operation for a three-phase four-wire system containing a shunt active power filter. With such a control strategy, when one of the filter converter legs is lost, the power compensation can still be provided partially with only two phases. A suitable pulse-width modulation control is derived to be used for the after-fault operation. The system model after-fault condition is derived. Experimental results are presented to demonstrate the correctness of the proposed solution.

I. Introduction

The strict regulations about the flow of electrical energy has stimulated the use active power compensation schemes [1-4]. The active power compensation is normally achieved with the help of switching converters connected as an active filter to the load in series, parallel or mixed arrangements to improve the power quality.

AC power systems are very sensitive to different types of failure occurring at the power converter. Whenever possible, the isolation of the fault is preferable to the stop for immediate maintenance since it keeps the system partially operative. In recent years different papers have focused schemes of fault-tolerant converter systems [5-8].

This paper investigates what type of control strategies can be applied when one of the legs of a three-phase four-wire active power filter is lost but it is required to keep, at least partially, the system operating under balanced power conditions. Figure 1(b) illustrates this situation for the case where the connection of the leg q1-q4 was lost. With the control strategies proposed, when one of the filter converter legs is lost, the power compensation can still be partially provided with only two phases, until it can be stopped with security. The control strategies permit to maintain the active power compensation except for the homopolar term or the current control for the faulty phase. Besides an appropriate dynamic modelling for controlling the system under fault condition, this paper also presents a specific pulse-width modulation technique to define the switching patterns of the converter of the active power filter for this case.

II. System Description

Figs. 1(a) present an electric power system composed by the grid source, a three-phase load and a voltage source converter. The grid source is composed by three balanced voltage sources (e,i, e,s, eSs) with equal series resistances (Rg) and inductances (Lg). The switching converter of the active power filter is a voltage source converter (VSC) with equal series resistances (Rf) and inductances (Lf). The switching converter of the active power filter is a voltage source converter (VSC) with equal series resistances (Rf) and inductances (Lf). This converter also has a fourth leg connected to the grid neutral. The current at the output of the VSC is controlled to deliver the reactive and harmonic current demanded by the load in order to provide the active power filtering feature. The dc side of the VSC requires a capacitor bank with minimum energy storage capacity. The converter needs active power because the load voltage is different from the grid voltage (due to the finite values of Rs and Ls) and because the losses associated to its operation. The average power of the VSC can be directly supplied from the grid. In this case, the active power filter control can be achieved by controlling the grid current in phase with the grid voltage, while the current amplitude is defined by DC-bus voltage regulation.

Figure 1(a) presents the normal operating condition while Fig. 1(b) illustrates one of the possible faulty configurations. In this case, the connection of the phase 1 is lost. An important and related problem is to detect and locate the fault. This paper will not discuss the fault detecting techniques and then it is assumed that the faulty condition can be identified by using the methods presented in [9].

The most relevant characteristics exhibited by the system can be preserved after the fault. However, after the fault, the active filter cannot control all the three-phase currents but, instead, it can control the components of the dq grid currents to be in phase with the corresponding components of the dq grid voltages or it can control the currents of the healthy phases to be in phase with the corresponding grid voltages.
a particular type of load, the voltage-current relationship of the load must be provided and added to the model given by (1)-(2).

The load phase voltages $u_l (j = 1, 2, 3)$ can be given by

$$u_l = e_{sj} - R_s i_s - L_s \frac{di_{sj}}{dt}$$  \hspace{1cm} (3)

From the above equation it can be seen that the time derivative of the load current is required to compute $u_l$ and must be derived from the given load model.

B. Model in $odq$ coordinates

The $odq$ variables can be determined from the 123 variables by using the transforming equation given by

$$y_{123} = A y_{odq}$$  \hspace{1cm} (4)

with $y_{123} = [y_1, y_2, y_3]^T$, $y_{odq} = [y_o, y_d, y_q]^T$ and

$$A = \begin{bmatrix} \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & 0 \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & \frac{\sqrt{3}}{2} \\ \frac{1}{\sqrt{2}} & -\frac{1}{\sqrt{2}} & -\frac{\sqrt{3}}{2} \end{bmatrix}$$  \hspace{1cm} (5)

This transforming matrix is such that $A^{-1} = A^T$ and the vector $y_{123}$ ($y_{odq}$) can be either grid, filter and load voltage vectors $e_{123}$, $f_{123}$, $i_{123}$ ($e_{odq}$, $f_{odq}$, $i_{odq}$) or grid, filter and load current vector $i_{123}$, $f_{123}$, $i_{123}$ ($i_{odq}$, $f_{odq}$, $i_{odq}$).

By applying the transforming equation given in (4) to the equations (1)-(2), the following $odq$ model for representing the equivalent circuit of Fig. 2(a) can be obtained ($k = o, d, q$):

$$e_{sk} - u_{fk} + u_{ik} = R_t i_{sk} + L_t \frac{di_{sk}}{dt}$$  \hspace{1cm} (6)

$$u_{ik} = R_f i_{ik} + L_f \frac{di_{ik}}{dt}$$  \hspace{1cm} (7)

From (3) and (4) the $odq$ voltage load $u_{lk}$ ($k = o, d, q$) is given by

$$u_{lk} = e_{sk} - R_s i_{sk} - L_s \frac{di_{sk}}{dt}$$  \hspace{1cm} (8)
IV. After-Fault Model

A. Three-phase model

Figure 1(b) illustrates the system configuration when the phase 1 is open ($i_f 1 = 0$). In this case, the circuit equations for phases 2 and 3 ($j = 2, 3$) are

$$e_{sj} - v_{fj} + u_{ij} = R_i i_{sj} + L_i \frac{di_{sj}}{dt} \quad (9)$$

$$u_{ij} = R_f i_{ij} + L_f \frac{di_{ij}}{dt} \quad (10)$$

and for the phase 1

$$v_{11} = e_{s1} - R_s i_{11} - L_s \frac{di_{11}}{dt}; \quad i_f 1 = 0 \quad (11)$$

where $v_{11}$ is the voltage for phase 1 of the load.

Note that $i_{s2}$ and $i_{s3}$ can be controlled by adjusting $v_{f2}$ and $v_{f3}$, respectively. However, $i_{s1}$ cannot be controlled by $v_{f1}$.

B. Model in $dq$ coordinates

Using the transforming equation given in (4) for the case where the phase 1 open, i.e., $i_{s1} = i_{11}$, it follows that:

$$i_{s2} = i_{11} - \sqrt{3/2}i_{sd} + \sqrt{1/2}i_{sq} \quad (12)$$

$$i_{s3} = i_{11} - \sqrt{3/2}i_{sd} - \sqrt{1/2}i_{sq} \quad (13)$$

Under these conditions the $dq$ model for unbalanced operation is given by

$$e_{sz} - v_{fz} + u_{lz} = R_i i_{sz} + L_i \frac{di_{sz}}{dt} \quad (14)$$

$$e_{sz} - v_{fz} + u_{lz} = R_i i_{sz} + L_i \frac{di_{sz}}{dt} \quad (15)$$

in which the new voltages $y_z$ and $y_{z2}$ ($y$ indicates $e_s$, $v_f$, $e_f$), defined by

$$y_{z1} = -\sqrt{1/6}(u_y + u_1), \quad y_{z2} = \sqrt{1/2}(u_y - u_1) \quad (16)$$

with $e_{lz} = \sqrt{2/3}(R_i i_{lz} + L_i di_{lz}/dt)$, have been introduced.

Note that, except for the unbalanced disturbance terms $e_{s1}$, $u_t$ and $e_{sz}$ this model is balanced. It is also possible to merge the disturbance into the model and this leads to a model that will be unbalanced and have parameters that depend on what phase is open.

It was mentioned previously that, in this case, only the currents $i_{s2}$ and $i_{s3}$ can be controlled. In the $dq$ framework this means that only two currents out of $i_{s0}$, $i_{sd}$ and $i_{sq}$ can be controlled. Moreover, since $e_{s0} = 0$, for providing balanced power, even when leg 1 of the voltage source converter is lost, then the natural choice is to control $i_{sd}$ and $i_{sq}$. Then, the homopolar current will be given by $i_{s0} = (i_{11} + i_{s2} + i_{s3})/\sqrt{3}$ or by $i_{s0} = (\sqrt{3}i_{11} - \sqrt{2}i_{sd})$ if (12)-(13) are introduced. The current $i_{s0}$ depends on $i_{11}$ (defined by the load) and $i_{sd}$ (determined by (14)-(15)).

This analysis also helps to understand why $i_{s0}$ cannot be controlled by the active filter.

The modelling for the case when the failure occurs in another phase follows similar lines. The corresponding model is different from model presented above only by the disturbance terms.

V. Control Schemes

The amplitude of the grid current is defined by capacitor DC-bus voltage control. To control the active filter, an appropriate model for designing the current control loop
The current controller can be either based on the three-amplitude of the grid reference current. The current controller scheme is modified to take into account the after-fault condition. The first one is to control the currents of the healthy phases to be in phase with the corresponding grid voltages (Case a). The second alternative is to control the dq components of the grid currents to be in phase with the corresponding dq components of the grid voltages (Case b). The design of the current controller for the Case a is based on the phase components model given in (9)-(10) on the other hand the model given by (14)-(15) can be used to design the current controller for the Case b. The control variables are $v_{f2}$ and $v_{f3}$ for the Case a and $v_{f2}$ and $v_{f3}$ for the Case b. Note that $e_{s2}$, $e_{s3}$, $u_{f2}$ and $u_{f3}$, for the Case a, are considered as disturbance to be compensated by the controller. The same observation holds for $e_{s2}$, $e_{s2}$, $u_{f2}$, $u_{f3}$ and $e_{t2}$ in the Case b.

Figure 3(b) presents the block diagram for illustrating the after-fault strategy when the control scheme is based on healthy phases (Case a). The block (S) represents that one of the reference current is modified to take into account the fault. Figure 3(c) presents the block diagram for illustrating the after-fault strategy when the control scheme is based on the dq coordinates (Case b). In this case the block (T) represents the transforming equation that was modified to take into account in what leg the fault has occurred. For example, the equations (12) to (13) represents the specific implementation of this block when phase 1 is open.

The control blocks shown in Fig. 3 can be implemented using either linear or non-linear control laws. The simplest non-linear solution would be the hysteresis control which is nearly parameter independent but requires high and variable switching frequency. The controller must be able to handle both positive and negative sequence components [10] that are created by the unbalanced disturbances as well as by the unbalancing of the load. As discussed before, the models for describing the dynamics of the system in all the possible after-fault configurations differ only by the disturbance terms. This means that it is possible to have a controller with the same parameters no matter what phase is open.

**VI. PWM Control**

The pulse-width modulation technique for the before-fault configuration is well known [11] and will not be discussed in this paper. On the other hand, the pulse-width modulation technique for the after-fault configuration [Block AF-PWM shown in Figs. 1(b) and 1(c)] will be discussed in this paper.

The pulse-width modulation for the after-fault configuration can be defined either in terms of the phase voltages (scalar modulation) or in terms of the voltage space plane (vector modulation). The two approaches will be discussed in the following.

For the configuration of Fig. 1(b) let us consider that the conduction state of the power switches is associated to binary variables $q_2$, $q_3$, $q_4$, $q_6$, $q_8$ and $q_0$. Therefore, from now on binary '1' will indicate a closed switch and binary '0' an open one. Pairs $q_2$-$q_5$, $q_3$-$q_6$ and $q_0$-$q_8$ are complementary and as a consequence, $q_5 = 1 - q_2$, $q_6 = 1 - q_3$ and $q_8 = 1 - q_0$.

The voltages $v_{f2}$ and $v_{f3}$, depend on the states of the power switches and may be expressed in terms of the previously defined binary variables $q_2$, $q_3$ and $q_6$, as

\[ v_{f2} = v_{f20} - v_{n0} = (2q_2 - 1)E/2 - (2q_3 - 1)E/2 \]  
\[ v_{f3} = v_{f30} - v_{n0} = (2q_3 - 1)E/2 - (2q_6 - 1)E/2 \]

where $E$ is the DC-bus voltage.

The pulse-widths of the gate signals can be calculated given the voltages referred to the DC-bus midpoint. If the desired phase voltages are specified by $v'_{f2}$ and $v'_{f3}$ then, according with (17) and (18), the reference midpoint voltages are given by

\[ v_{f20} = v_{f2} + v_{n0} \]  
\[ v_{f30} = v_{f3} + v_{n0} \]

Note that these equations cannot be solved unless $v_{n0}$ be specified.

Since $v'_{f20} \leq E/2$, $v'_{f20} \leq E/2$ and $|v_{n0}| \leq E/2$, it follows that

\[ |v'_{f20} - v'_{f30}| \leq E \]

which demonstrates that $v'_{f2}$ cannot be specified independently of $v'_{f3}$ and vice-versa.

**A. Scalar PWM**

In order to specify $v'_{f2}$ and $v'_{f3}$ independently we must choose $v_{n0}^* = 0$ and consequently from (17) and (18) it...
follows that

$$v_{f20} = v_{f2}$$ and $$v_{f30} = v_{f3}.$$  (21)

In this case $$v_{f2}$$ and $$v_{f3}$$ are restricted to

$$\left| v_{f2} \right| \leq E/2 \text{ and } \left| v_{f3} \right| \leq E/2,$$  (22)

and corresponds to the same voltage capability as obtained for the converter in which the leg $$q_0 - q_3$$ is not used, and it is replaced by using a capacitor DC-bus midpoint (two-leg converter).

However, it is possible to conceive another strategy that permits to increase the largest value of $$v_{f2}$$ and $$v_{f3}$$ by playing with $$v_{n0}$$. In this case, the value for $$v_{n0}$$ is dependent on the values specified for $$v_{f2}$$ and $$v_{f3}$$. The strategy proposed for increasing the voltage capability is described through the following algorithm:

1. Determine the largest $$v_{\text{max}}^*$$ and smallest $$v_{\text{min}}^*$$ voltage value among $$v_{f2}$$ and $$v_{f3}$$.
2. Check for the existence condition, i.e., $$\left| v_{\text{max}}^* \right| \leq E$$ and $$\left| v_{\text{min}}^* - v_{n0}^* \right| \leq E$$. If this test fails then STOP, i.e., the converter cannot synthesize the specified voltage profile. If this test is OK go to next step.
3. If $$v_{\text{max}}^* = 0$$ then clamp $$v_{n0}^*$$, i.e., $$v_{n0}^* = E/2$$, $$v_{f20} = v_{f2} + E/2, v_{f30} = v_{f3} + E/2$$.  
4. If $$v_{\text{max}}^* > 0$$ then make the midpoint reference voltage ($$v_{f20}^*$$ or $$v_{f30}^*$$) associated the reference phase voltage ($$v_{f2}^*$$ or $$v_{f3}^*$$) that provided $$v_{\text{max}}^*$$ equal to $$E/2$$. From the specified reference phase voltages ($$v_{f2}^*, v_{f3}^*$$), calculate $$v_{n0}^*$$ by using one of the equations given in (19). Now, knowing $$v_{n0}^*$$ the other midpoint reference voltage can be determined by using the remaining equation.

Let’s consider the following example to illustrate the execution of Step 4 of the proposed algorithm. Supposing that $$v_{\text{max}}^* = v_{f2}$$, then $$v_{f20} = E/2, v_{n0} = E/2 - v_{f2}$$ and consequently $$v_{f20} = v_{f3} + E/2 - v_{f2}$$.  

Differently from the case where $$v_{n0}^* = 0$$, the values obtained for $$v_{f2}$$ and $$v_{f3}$$ are now restricted to

$$\left| v_{f2} \right| \leq E \text{ and } \left| v_{f3} \right| \leq E,$$  (23)

respectively. However, in this case, voltages $$v_{f2}$$ and $$v_{f3}$$ are mutually dependent. For example, to obtain $$v_{f2} = E$$, it is required that $$v_{f20} = E/2$$ and $$v_{n0} = E/2 - v_{f2}$$ and consequently $$v_{f30} = v_{f3} + E/2 - v_{f2}$$.  

Given the midpoint reference voltages, the pulse widths for the command signals of the leg switches $$\tau_1, \tau_2$$ and $$\tau_3$$ can be determined by

$$\tau_1 = \frac{T}{2} + \frac{T}{E} v_{f20}^* \quad \tau_2 = \frac{T}{2} + \frac{T}{E} v_{f30}^* \quad \tau_3 = \frac{T}{2} + \frac{T}{E} v_{n0}^*.$$  (24)

It is possible to deal with the free-wheeling periods to minimize the harmonic distortion of the output voltage of the after-fault converter topology. The free-wheeling periods can be distributed at the begin and at the end of the switching period. The free-wheeling time corresponds to the smallest value, $$\tau_{\text{min}}$$, among $$\tau_1, \tau_2$$ and $$\tau_3$$. The distribution of free-wheeling periods along the switching interval can be made by adding $$\tau_{\mu} = (\mu - 1)\tau_{\text{min}}$$ to the pulse widths calculated previously via equation (24). The variable $$\mu$$ in the expression for $$\tau_{\mu}$$ is defined as the apportioning factor and is restricted to $$0 \leq \mu \leq 1$$.

**B. Vector Modulation**

The voltages supplied by the power converter to the system phases can be displayed in the space vector plane. This vector plane is defined such that $$v_{f2}$$ and $$v_{f3}$$ correspond to the real axis (Re) and the imaginary (Im) axis, respectively, that is:

$$v_k = v_{f2} + jv_{f3} \quad \text{for } k = 0, 1, 2, ..., 8$$  (25)

as shown in Fig. 4. The instantaneous voltage vectors generated by the converter are listed in Table 1, for the binary states of the power switches.

<table>
<thead>
<tr>
<th>$$q_2$$</th>
<th>$$q_3$$</th>
<th>$$q_0$$</th>
<th>vector</th>
<th>$$v_{f2}$$</th>
<th>$$v_{f3}$$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0</td>
<td>0</td>
<td>$$v_0 = 0$$</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>0</td>
<td>$$v_1 = E$$</td>
<td>0</td>
<td>E</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>0</td>
<td>$$v_2 = \sqrt{2}Ee^{j\pi/4}$$</td>
<td>E</td>
<td>E</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
<td>0</td>
<td>$$v_3 = Ee^{j\pi/2}$$</td>
<td>0</td>
<td>E</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>1</td>
<td>$$v_4 = Ee^{j\pi/2}$$</td>
<td>-E</td>
<td>0</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>1</td>
<td>$$v_5 = \sqrt{2}Ee^{j3\pi/4}$$</td>
<td>-E</td>
<td>-E</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>1</td>
<td>$$v_6 = Ee^{j3\pi/2}$$</td>
<td>0</td>
<td>-E</td>
</tr>
<tr>
<td>1</td>
<td>1</td>
<td>1</td>
<td>$$v_7 = 0$$</td>
<td>0</td>
<td>0</td>
</tr>
</tbody>
</table>

| TABLE 1. Vector and voltages generated by the converter. |

There are eight vectors: four vectors with amplitude $$E$$ ($$v_1, v_3, v_4$$ and $$v_6$$), two vectors with amplitude $$\sqrt{2}E$$ ($$v_2$$ and $$v_5$$) and two null vectors ($$v_0$$ and $$v_7$$), as shown in Table 1 and Fig. 4. These vectors define six sectors $$K = 1, 2, ..., 6$$. The phase-voltage $$v_{f2}$$ and $$v_{f3}$$ assume only three different values: $$E, 0, \text{ or } -E$$.

Let $$v^*$$ represent the reference voltage to be synthesized by the converter within one cycle time of length $$T$$. According to the vector modulation technique, the reference vector located in sector $$K$$ is synthesized by using a time-weighted sum of the two adjacent vectors that define the
sector [12]. Then, for the sector K, it can be written that
\[ v^* = v_K \frac{t_K}{T} + v_{K+1} \frac{t_{K+1}}{T} + v_0 \frac{t_0}{T} + v_r \frac{t_7}{T} \]  
(26)
with the time positive weights for each vector \( t_K, t_{K+1}, t_0 \) and \( t_7 \) restricted to \( T = t_K + t_{K+1} + t_0 + t_7 = K = 0, 1, 2, ..., 8 \) and \( K + 1 = 1, if K = 6 \). The purpose of adding the zero vectors \( v_0 \) and \( v_7 \) in (26) is to obtain a constant switching period for the converter.

Note that in steady state, \( v^* = v_m \cos(\omega*t - \frac{2\pi}{3}) + jv_m \cos(\omega*t + \frac{2\pi}{3}) \) does not describe a circle in the complex plane.

By using (25), then \( v^* = v_{i2} + jv_{i3} \) and since \( v_0 = 0 \) and \( v_7 = 0 \), it follows from (26) that

\[ \text{If } K = 1 \Rightarrow t_1 = (v_{i2} - v_{i3})T_E; \quad t_2 = v_{i3}T_E \]
\[ \text{If } K = 2 \Rightarrow t_2 = v_{i2}T_E; \quad t_3 = (v_{i2} - v_{i3})T_E \]
\[ \text{If } K = 3 \Rightarrow t_3 = v_{i3}T_E; \quad t_4 = -v_{i2}T_E \]
\[ \text{If } K = 4 \Rightarrow t_4 = (v_{i2} - v_{i3})T_E; \quad t_5 = -v_{i3}T_E \]
\[ \text{If } K = 5 \Rightarrow t_5 = -v_{i2}T_E; \quad t_6 = (v_{i2} - v_{i3})T_E \]
\[ \text{If } K = 6 \Rightarrow t_6 = -v_{i3}T_E; \quad t_1 = v_{i2}T_E \]

(27)

where \( T_E = T/E \).

Given the magnitudes of the components of vector \( v^* \), i.e., \( v_{i2} \) and \( v_{i3} \), the sector can be determined in the following way

\[ \text{If } v_{i2} > 0, \quad v_{i3} \geq 0, \quad v_{i2} > v_{i3} \Rightarrow K = 1 \]
\[ \text{If } v_{i2} > 0, \quad v_{i3} \geq 0, \quad v_{i2} \leq v_{i3} \Rightarrow K = 2 \]
\[ \text{If } v_{i2} \leq 0, \quad v_{i3} > 0 \Rightarrow K = 3 \]
\[ \text{If } v_{i2} < 0, \quad v_{i3} \leq 0, \quad v_{i2} < v_{i3} \Rightarrow K = 4 \]
\[ \text{If } v_{i2} < 0, \quad v_{i3} < 0, \quad v_{i2} \geq v_{i3} \Rightarrow K = 5 \]
\[ \text{If } v_{i2} \geq 0, \quad v_{i3} < 0 \Rightarrow K = 6 \]

(28)

The values for \( t_0 \) and \( t_7 \) are not defined. They can be chosen by introducing the apportioning factor \( 0 \leq \mu \leq 1 \)
\[ t_0 = \mu(T-t_K-t_{K+1}); \quad t_7 = (1-\mu)(T-t_K-t_{K+1}) \]  
(29)

Then, the modulation technique for the scheme presented in Fig. 1(b) can be implemented by executing the following steps:
1. Given \( v_{i2}^* \) and \( v_{i3}^* \), determine \( K \) from (28) and \( t_K \) and \( t_{K+1} \) from (27).
2. Given \( \mu \) compute \( t_0 \) and \( t_7 \) from (29).

The modulation strategy for the other phase open follows similar steps.

VII. SIMULATION RESULTS

The two control schemes employed for the after-fault configuration have been evaluated by numerical simulation. In both cases the standard hysteresis control law was employed to implement the current controller (\( k_{i2,123} \)), and the capacitor voltage control was achieved with standard proportional-integral controllers (\( R_c \)). The parameters employed in the numerical simulation (see Fig. 1) are: \( R_c = 0.05p.u., \quad X_c = 0.2p.u., \quad R_f = 0.05p.u., \quad X_f = 0.2p.u. \). The load current has the following composition: active (0.5pu) + reactive (0.3pu) + negative sequence (0.2pu) + homopolar (0.17pu) + fifth harmonic (0.1pu). The grid voltage amplitude is 1pu and \( \omega = 120\pi \text{rad/s} \).

Figure 5 (Case a) and Figure 6 (Case b) present the simulation results. In these figures the system runs without fault up to \( t = 0.206s \), when the connection of the leg 1 of the active filter is broken. During the time period where there is no fault the two control schemes provide the same performance. The \( dq \) and 123 currents are controlled quite well. The grid power \( (p_g) \) is nearly constant, except for the ripple introduced by the switching of the converter, and the average power \( (p_f) \) of the active filter is zero. However, after the fault the two schemes lead to different behaviors.

Figure 5(a) shows that, in the Case a, the power grid presents an important AC term, the current and voltage are not sinusoidal, except for the currents \( i_{i2} \) and \( i_{i3} \) and voltages \( u_{i2} \), \( u_{i3} \) and \( u_{i23} \). Figure 6 shows that, in the Case b, the ripple of the power grid remains essentially the same and the \( dq \) grid currents are controlled quite well in phase with \( dq \) grid voltage. Also, the \( dq \) and the line load voltages are quite sinusoidal but the grid phase current, the load currents and the load line voltages are not sinusoidal (compensated).

The comparison of the simulation results for Case a and Case b shows that with the scheme a the system can operate near to the ideal condition, i.e., unity power factor and \( dq \) and line load voltages sinusoidal. However, with scheme b the grid currents and load voltages for the phase that is not open have a much better profile than for the Case a.
VIII. EXPERIMENTAL RESULTS

In the experimental set-up, the current and the voltage controllers were implemented by software in a Pentium II – 266MHz microcomputer, equipped with dedicated plug-in boards and sensors, with $h = 200\mu s$, and the switching frequency of the VSC was $5kHz$. The unbalanced load has been implemented by using a three-phase induction machine, with a resistor connected in parallel with one of the machine windings. The control strategy for Case b was employed. The current control law was the same used in the positive-negative sequence controller.
described in [10].

Figure 7 shows the reference and actual \( odq \) and \( 123 \) currents for the system before and after the fault occurrence. The fault was caused at \( t = 0.02 \) s, when both switches of leg 3 are open. Except for the transient caused by the fault, the actual current tracks quite well the reference ones. Note that the reference current \( i_{odq} \) after the fault is equal to \( i_{odq} \).

IX. Conclusions

This paper has demonstrated that is possible to provide fault tolerant properties for a three-phase shunt active power filter by using appropriated control strategies. Since the after fault configuration results in a asymmetric system it was necessary to develop a suitable model and control strategies.

It has been demonstrated that it is possible to maintain power balanced operation of the grid source even when one of the converter legs is lost. However, it is not possible to provide full compensation since the configuration of the active filter when the fault occurs does not permit to control the homopolar component or one of the phase currents. The proposed solution extends the functionality of the system and, consequently, increases its reliability.

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