Sliding mode control of a three-phase parallel active filter based on a two-level voltage converter

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ABSTRACT
This paper presents the sliding mode control of a three-phase parallel active filter based on a two-level voltage converter to compensate for the harmonic currents of the pollutant loads. In order to calculate the reference harmonic currents, the p-q algorithm is used and the PWM is used to generate the control pulses of the inverter. Simulations in the Matlab-Simulink environment are provided to validate the theoretical study. The results obtained seem satisfactory in the harmonic compensation quality and the correction of the power factor. The selected comparison criteria are the transient regime and the Harmonic Distortion Rate in the line current.

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Harmonics; parallel active filter; two-level inverter; algorithm p-q; sliding mode control

1. Introduction
The development of power electronics and the increased power involved and the flexibility of the use of semiconductors has encouraged power systems engineer and scientist to undertake important combinations of static power converters with electrical machines (Sahbani, Labben-Ben Braiek, Dami, & Jemli, 2005). Such devices absorb a non-sinusoidal current and act as harmonic generators.

These disturbed currents cause damage to the quality of the feed. This not only explains the increase in the harmonic ratio and the unbalance of the currents and the voltages, but also a significant consumption of the reactive power (Zahzouh, Khochmane, & Haddouche, 2015). These harmonic disturbances have catastrophic consequences on the performance of all the receivers connected to the electrical networks and to the power source. Therefore, it is necessary to find a suitable solution to reduce these disturbances to the lowest level (Zahzouh, Khochmane, et al., 2015). These harmonic disturbances have catastrophic consequences on the performance of all the receivers connected to the electrical networks and to the power source. Therefore, it is necessary to find a suitable solution to reduce these disturbances to the lowest level (Zahzouh, Khochmane, et al., 2015). One of the solutions used is passive filtering: it consists in trapping the harmonic currents in LC circuits, tuned on the rows of harmonics to be filtered. Rows 5 and 7 are commonly filtered (Dixon, Garcia, & Moran, 1995). However, this solution is of average efficiency.

The remarkable progress made in the last few years in the field of electronic power devices has made it possible to design self-adapting harmonic eliminating devices called active harmonic compensators or active filters (Sahbani et al., 2005).

The idea of the active power filter presents a solution well suited to these problems encountered in power lines (Zahzouh, Khochmane, & Haddouche, 2015). The compensation of harmonics with the active filters in the electrical network is done in two steps. The first stage discovers the harmonic components of the current and the second step injects these harmonics into the electrical network in phase opposition (Morsli, Tlemcani, Boucherit, & Ould Cherchali, 2012). The active filters have developed rapidly and can be structured in parallel (Zahzouh, Khochmane, et al., 2015), in series (Peng, 1998), or hybrid (Singh, Al-Haddad, & Chandra, 1999), in the network.

Several techniques are used to detect disturbances in electrical networks. Frequency detection techniques are performed by the Discrete Fourier Transform (DFT) which can be used to analyze non-sinusoidal voltage or signals. Also, the Fast Fourier Transform (FFT) and Discrete Recursive (DRFT) (George & Bones, 1991; Ogata, 1995) represent effective computational methods. However, direct application of these methods requires a large calculation time which delays the control response of the filter. Nevertheless, there are other techniques, such as the Notch filter (Quinn, Mohan, & Mehta, 1993), the (ANN) artificial neural network technique that has been developed for the optimal identification of harmonic signals.
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(Bose, 2007; Sindhu, Manjula, & Nambiar, 2014) and the instantaneous power method (Akagi, Kanazawa, & Nabae, 1983; Pouresmaeil, Montesinos-Miracle, Gomis-Bellmunt, & Sudrià-Andreu, 2011) which is used constantly.

In this study we propose the sliding mode control of a three-phase parallel three-phase parallel active filter based on a two-stage inverter with instantaneous powers for the identification of harmonics. The originality of our control strategy is in the definition of the sliding surface to which we have added another term which is the integral of the error. The advantage of this approach is that the sliding surface is a plane passing through the origin. In case where the system is of the second order $r = 2$, the solution is obtained in a plane whereas for the classic sliding mode control, the solution is obtained on a line.

2. Principle of parallel active filter

This filter is connected in parallel on the network. The principle of the parallel active filter consists in generating currents in opposition of phase to the harmonic currents existing on the network and created by the nonlinear loads (Sahbani et al., 2005). In this way, the current supplied by the energy source remains sinusoidal. The parallel active filter consists of two blocks: the power part and the control-command part (Figure 1).

The network connection of the polluting charge and the compensator is effected via the short-circuit impedance, represented by the inductance $L_s$ and the resistor $R_s$. The diode rectifier bridge delivers on a load ($R_L$). The time constant of the inductive dipole will be chosen to be sufficiently large, so as to ensure sufficient smoothing of the rectified current $I_L$ and preventing the rectifier operating in batch mode (Sahbani et al., 2005).

Figure 1. Principle of the parallel active filter (Hind Djeghloud, 2007).

Figure 2. Two-stage three-phase voltage inverter (Hind Djeghloud, 2007).
The power circuit of the parallel active filter consists of the voltage inverter, the coupling filter and the energy storage system. The structure of the voltage inverter used in this article is shown in Figure 2 below.

This structure of the two-level inverter comprises three arms consisting of six reversible current switches (bipolar transistors, IGBTs, GTOs) controlled on opening and closing in antiparallel with a diode.

3. Development of the identification method

This method introduced by H. Akagi which is a temporal method is used to avoid the difficulties due to the high number of computation when implementing frequency methods (Hafsia, 2015). It leverages Concordia’s transformation of simple network voltages and load line currents, in order to calculate instantaneous real and imaginary powers. It makes it possible to transform the fundamental component into a continuous component and the harmonic components into alternative components. This method of identifying harmonic currents consists of eliminating the DC component of the instantaneous active and reactive power.

The vectors of the simple voltages at the connection point \(v_s\) and of the charge currents \(i_c\) are noted respectively.

\[
[v_s] = \begin{bmatrix} v_{sa} \\ v_{sb} \\ v_{sc} \end{bmatrix}, \quad [i_c] = \begin{bmatrix} i_{ca} \\ i_{cb} \\ i_{cc} \end{bmatrix}
\]

In the space \(\alpha - \beta\) the tensions and the currents are expressed by the following expressions (Hafsia, 2015):

\[
\begin{bmatrix}
v_{sa} \\
v_{sb} \\
v_{sc}
\end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix}
1 \\
-\frac{1}{2} \\
-\frac{1}{2}
\end{bmatrix} \begin{bmatrix}
v_{sa} \\
v_{sb} \\
v_{sc}
\end{bmatrix}
\]

\[
\begin{bmatrix}
i_{ca} \\
i_{cb} \\
i_{cc}
\end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix}
1 \\
-\frac{1}{2} \\
-\frac{1}{2}
\end{bmatrix} \begin{bmatrix}
i_{ca} \\
i_{cb} \\
i_{cc}
\end{bmatrix}
\]

The instantaneous real and imaginary powers denoted respectively \(p\) and \(q\), are defined by the following matrix relation:

\[
\begin{bmatrix}
p \\
q
\end{bmatrix} = \begin{bmatrix}
v_{sa} & -v_{sb} & 0 \\
-v_{sb} & v_{sa} & -\frac{1}{\sqrt{2}} \\
-\frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}}
\end{bmatrix} \begin{bmatrix}
i_{ca} \\
i_{cb} \\
i_{cc}
\end{bmatrix}
\]

From the expression (4), and by positing:

\[
\Delta = v_{a}^2 + v_{b}^2
\]

we have:

\[
\begin{bmatrix}
i_{ca} \\
i_{cb}
\end{bmatrix} = \frac{1}{\Delta} \begin{bmatrix}
v_{sa} & -v_{sb} \\
v_{sb} & v_{sa}
\end{bmatrix} \begin{bmatrix}
p \\
q
\end{bmatrix}
\]

In the general case, each of the powers \(p\) and \(q\) has a continuous and an alternating part, which allows us to write the expression below (Alali, 2002):

\[
\begin{cases}
\rho = \bar{p} + \tilde{p} \\
q = \bar{q} + \tilde{q}
\end{cases}
\]

With \(\bar{p}\): Continuous power linked to the active fundamental component of current and voltage, \(\bar{q}\): Continuous power related to the reactive fundamental component of current and voltage, \(\tilde{p}\tilde{q}\): Alternative powers related to the sum of the harmonic components of the current and the voltage.

Considering Equations (6) and (7), we can separate the current in the reference \(\alpha - \beta\) into three components, active and reactive at the fundamental frequency and the sum of the harmonics. This leads to (Morsli et al., 2012):

\[
\begin{bmatrix}
i_{ca} \\
i_{cb} \\
i_{cc}
\end{bmatrix} = \frac{1}{\Delta} \begin{bmatrix}
1 & 0 & 0 \\
0 & v_{sa} & -v_{sb} \\
0 & -v_{sb} & v_{sa}
\end{bmatrix} \begin{bmatrix}
\bar{p} \\
\bar{q} \\
\bar{q}
\end{bmatrix}
\]

According to relation (8), to identify the harmonic currents, the alternating parts of the real and imaginary powers must be separated from the continuous parts (Morsli et al., 2012). This separation can be carried out using an extraction filter as illustrated in Figure 3.

The three-phase disturbing currents which represent the identified currents, referred to as reference currents \(i_{ref}\), are calculated from the inverse transformation \(\alpha - \beta\) inverse (transformation \(C_{2-3}\)) given by the relation:

\[
\begin{bmatrix}
i_{refa} \\
i_{refb} \\
i_{refc}
\end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix}
1 \\
-\frac{1}{2} \\
-\frac{1}{2}
\end{bmatrix} \begin{bmatrix}
i_{ca} \\
i_{cb} \\
i_{cc}
\end{bmatrix}
\]

With

\[
\begin{bmatrix}
i_{ca} \\
i_{cb} \\
i_{cc}
\end{bmatrix} = \frac{1}{\Delta} \begin{bmatrix}
v_{sa} & -v_{sb} \\
v_{sb} & v_{sa}
\end{bmatrix} \begin{bmatrix}
\bar{p} \\
\bar{q}
\end{bmatrix}
\]

Figure 3. Extraction filters (Hafsia, 2015).
4. p-q algorithm for extracting harmonic currents

The diagram of Figure 4 illustrates the various steps making it possible to obtain the harmonic components of the current of a non-linear load.

5. Dynamic equation of shunt active power filter and controller design

The equation of the voltage per phase of the three-phase parallel active filter is (Moussa, 2016):

\[
v_{sk} = v_{fk} - v_{Lfk} - v_{Rfk} = v_{fk} - L_f \frac{di_{fk}}{dt} - R_f i_{fk}
\]

With \( k = a, b, c \) The phase index.

Hence, the equations of the three phases are expressed by the following matrix equation:

\[
\begin{bmatrix}
L_f \frac{di_{fa}}{dt} & -R_f i_{fa} \\
L_f \frac{di_{fb}}{dt} & -R_f i_{fb} \\
L_f \frac{di_{fc}}{dt} & -R_f i_{fc}
\end{bmatrix}
= \begin{bmatrix}
v_{fa} \\
v_{fb} \\
v_{fc}
\end{bmatrix} - \begin{bmatrix}
v_{fa} \\
v_{fb} \\
v_{fc}
\end{bmatrix}
\]

The system of equation defining the active filter in the three-phase reference frame is given by:

\[
\begin{align*}
L_f \frac{di_{fa}}{dt} &= -R_f i_{fa} + v_{fa} - v_{sa} \\
L_f \frac{di_{fb}}{dt} &= -R_f i_{fb} + v_{fb} - v_{sb} \\
L_f \frac{di_{fc}}{dt} &= -R_f i_{fc} + v_{fc} - v_{sc}
\end{align*}
\]

From Equations (14), (17) et (18) we find:

\[
\begin{align*}
\frac{d}{dt} [i_{fa} f_{f\beta}] &= \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} \\
0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2}
\end{bmatrix} \begin{bmatrix} v_{fa} \\
v_{fb} \\
v_{fc}
\end{bmatrix}
\end{align*}
\]

Current regulator \( i_{fa} \).

The follow-up error of this variable is given by:

\[
e(i_{fa}) = i_{refa} - i_{fa}
\]

The sliding surface chosen for this variable is defined by:

\[
s(i_{fa}) = k_1 e(i_{fa}) + k_1 \int e(i_{fa}) dt
\]

During the slip mode, we have:

\[
\begin{align*}
s(i_{fa}) &= 0 \\
\dot{s}(i_{fa}) &= 0
\end{align*}
\]

Figure 4. Algorithm p-q for extracting harmonic currents.
\[
\dot{s} = k_1 \left( \frac{1}{L_f} \dot{i}_{ta} + \frac{R_f}{L_f} i_{ta} - \frac{1}{L_f} v_{fa} + \frac{1}{L_f} v_{sa} \right) + \frac{k_1}{k_1} e(i_{ta}) = 0
\]

(20)

Table 1. Simulation parameters.

| Parameters                          | Numerical values |
|-------------------------------------|------------------|
| Power supply                        | Effective voltage $E_s$ 230 V |
|                                    | Frequence $f$ 50Hz |
|                                    | Line Resistance $R_s$ 0.15 mH |
| Non-linear load (Graetz bridge      | Line Resistance $R_L$ 0.45 mH |
| with 6 diodes)                      | Line Inductance $L_s$ 3.55 mH |
| Linear load (downstream to          | Load Resistance $L_d$ 602 |
| non-linear load)                    | Load Inductance $L_d$ 20 mH |
| Reference Continuous Feeding $V_{dc-ref}$ | 750 V |
| Capacities of the inverter two levels $C_1 = C_2$ | 20 $\mu$F |
| Controller parameters by            | $k_1$ 5e4 |
| sliding mode                        | $k_2$ 5e10 |
|                                    | $k_{11}$ 5e4 |
| Current regulator $i_{f\beta}$      | $k_{12}$ 5e10 |
| The follow-up error of this variable is given by: | $k_3$ 1e4 |
|                                    | $c_1$ 0.4 |
|                                    | $c_3$ 5e-5 |
|                                    | $L_f$ 0.23 mH |
|                                    | $R_f$ 1e-6 |
|                                    | $T_s$ 5e-6 |
|                                    | $f_{MLI}$ 15 kHz |

The equivalent command is then defined by:

\[
v_{feq} = \frac{k_1}{k_1} L_f e(i_{ta}) + v_{sa} + R_f i_{ta} + L_f \dot{i}_{ta}
\]

(21)

Discontinuous control ensures the attractiveness of the sliding surface, it is defined by:

\[\Delta v_{fa} = U_{max} \text{sgn}(s(i_{ta})) \]

(22)

The command $v_{f\alpha}$ is given by:

\[v_{f\alpha} = v_{feq} + \Delta v_{fa} \]

(23)

Current regulator $i_{f\beta}$.

The sliding surface chosen for this variable is defined by:

\[s(i_{f\beta}) = k_2 e(i_{f\beta}) + k_{12} \int e(i_{f\beta}) dt \]

(25)

During the slip mode, we have:

\[
\begin{cases}
  s(i_{f\beta}) = 0 \\
  \dot{s}(i_{f\beta}) = 0
\end{cases}
\]

(26)

Figure 5. Block diagram of the control by sliding mode.
From Equations (14), (24) et (25) we find:

$$\dot{s} = k_2 \left(i_{ref\beta} + \frac{R_f}{L_f} i_{f\beta} - \frac{1}{L_f} v_{f\beta} + \frac{1}{L_f} v_{s\beta}\right) + k_2 e(i_{f\beta}) = 0$$\hspace{1cm} (27)

The equivalent command is then defined by:

$$v_{f\beta eq} = \frac{k_2 L_f}{k_2} e(i_{f\beta}) + v_{s\beta} + R_f i_{f\beta} + L_f i_{ref\beta}\hspace{1cm} (28)$$

Using Equations (14), (24) et (25), after all calculations, we have the Equation (27) which is the time differential of our sliding surface defined at Equation (25).

Discontinuous control ensures the attractiveness of the sliding surface, it is defined by:

$$\Delta v_{f\alpha} = U_{\text{max}} \text{sgn}(s(i_{f\beta}))\hspace{1cm} (29)$$

At Equation (28), we have used the second condition of Equation (26) that is $s(i_{f\beta}) = 0$ to determine the expression of the equivalent control $v_{f\beta eq}$ presented at Equation (28).

6. DC bus voltage control

The DC bus voltage control of the proposed system can be improved by adjusting the small active power rate in the capacitors. Thus, it compensates losses by conduction and switching (Akagi, Nabae, & Atoh, 1986). The voltage regulation loop is designated to be smaller than the current loop. The DC voltage control circuit must be fast and responsive only to steady-state conditions. Transient DC voltage variations are not permitted and are taken into consideration when selecting the appropriate capacitor value. At steady state, the fundamental component is not

Figure 6. Source current and its harmonic spectrum before filtering.
included in the reference current (Zahzouh, Khochman, et al., 2015).

The dynamic equations of the voltages $V_{dc1}$ and $V_{dc2}$ at the terminals of the capacitors $C_1$ and $C_2$ are (Moussa, 2016):

$$\begin{align*}
\frac{dV_{dc1}}{dt} &= \frac{P_{dc1}}{V_{dc1}C_{dc1}} \\
\frac{dV_{dc2}}{dt} &= \frac{P_{dc2}}{V_{dc2}C_{dc2}}
\end{align*}$$

(31)

$V_{dc}$ controller.

The follow-up error of this variable is given by:

$$e(V_{dc}) = V_{dc} - V_{dc}^{ref}$$

(32)

The sliding surface chosen for this variable is defined by:

$$s(V_{dc}) = k_3 e(V_{dc}) + \frac{d}{dt} e(V_{dc}) + k_{i3} \int e(i_{ra}) dt$$

(33)

In the slip mode, we define the following switching functions:

$$y_1 = \begin{cases} 1 & \text{si } s(V_{dc}) e(V_{dc}) \geq 0 \\ -1 & \text{si } s(V_{dc}) e(V_{dc}) < 0 \end{cases}$$

(34)

$$y_2 = \begin{cases} 1 & \text{si } s(V_{dc}) \dot{e}(V_{dc}) \geq 0 \\ -1 & \text{si } s(V_{dc}) \dot{e}(V_{dc}) < 0 \end{cases}$$

(35)

Hence, the output of the regulator is defined by:

$$u = P_{dc1} + P_{dc2} = c_3 e(V_{dc})y_1 + c_{i3} \dot{e}(V_{dc})y_2$$

(36)

With $c_3$ et $c_{i3}$ positive constants.

The block diagram of the sliding mode command is given in the figure below.

7. Simulation parameters

For this simulation, a rectifier bridge with three-phase diodes with an RL load is used as a non-linear load. Table 1 summarizes the simulation parameters. The study is carried out only in phase a, knowing that the other two phases b and c are offset respectively by 120° and 240° with respect to phase a (Figure 5).

8. Results and discussion

8.1. Before filtering

Figure 6 shows the source current and its harmonic spectrum before the filtering.

The feed stream obtained before the filter is completely deformed and its THD is 25.93%. This is higher than the standard standard (THD < 5%).

8.2. After filtering

Figure 7 shows the current injected by the active filter and its harmonic spectrum.

With regard to the curves obtained, before the connection of the active filter to 0.04 s, the non-linear charge pollutes the network and the waveform of the network is identical to that of the load. A study of the harmonic spectrum of phase A of the network reveals a THD equal to 25.93%. Once the active filter is connected (at 0.04 s), the waveform of the charge does not change, but there is a clear visual improvement of the waveform of the network which returns to sinusoidal shape thanks to the current injected by the Active parallel filter. We can easily verify this result with a study of the harmonic spectrum of phase A of the network after the connection of the filter, we note that after the connection of the filter, the THD rose from...
25.93% to 0.95% which complies with the imposed standards. The $i_{sa}$ and $v_{sa}$ quantities of phase A of the network are well in phase. The currents injected by the parallel active filter must, according to the principle of operation of the active filters, be a polluted current with harmonics of the same rank as that of the load but of opposite phase in order to be able to compensate them. A study of the harmonic spectrum of the current injected in phase A of the network shows us that the harmonics present in this current are the harmonics of ranks 5, 7, 11, 13, 17 and 19 with a THD of 68.0178% (Figures 8–10).

9. Conclusion

The work presented in this paper is the sliding mode control of a parallel three-phase active filter, which is the main solution to the problems caused by harmonic pollution in distribution networks.

The simulation results show that the filter performance is significantly improved. Indeed, a significant reduction of the total harmonic distortion rate (THD = 0.95%) is obtained according to IEEE standards. Similarly, a good compensation of the reactive power in the electrical distribution network is obtained with a power factor of less

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**Figure 8.** Source current and its spectrum before and after filtering.

**Figure 9.** Power factor.

**Figure 10.** Three-phase source currents after filtering.
than unity. The currents in the three phases have the same amplitudes with sinusoidal waveform. The main difficulty was found during the simulation, due to the important programming that was needed. In future work we can envisage implementation on FPGA systems or arm cortex systems. We can also use direct nonlinear model predictive control strategies to control this filter taking into account the nonlinearities of the model and some constraints, envisage the use of MPC-SVPWM control strategy.

**Disclosure statement**

No potential conflict of interest was reported by the authors.

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