ACTIVE POWER FILTERS BASED ON A SINGLE CURRENT MEASUREMENT SYSTEM

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Abstract

In this paper shunt active power filters are analyzed in order to define new control schemes characterized by simple control algorithms and a reduced number of current transducers. Standard active filter configurations requiring the measurement of both the load and filter currents are compared with configurations based on a single current measurement system. The comparison is made with reference to current regulators implemented by hysteresis and PWM modulators. It has been verified that the performance obtained by the simplified control schemes is similar to that of standard solutions.

1. - INTRODUCTION

High power non-linear loads, such as ac/dc rectifiers, cause undesirable phenomena in the operation of power systems. Among these, the most important ones are the harmonic pollution and the reactive power demand. As it is known, conventional passive LC filters have been widely used in the past notwithstanding their disadvantages. In order to avoid these problems, various active power filter configurations and control strategies have been proposed and developed in the last decade [1]-[8].

Parallel active filters have been recognized as a viable solution to current harmonic and reactive power compensation. The principle of operation is based on the injection of the current harmonics required by the non-linear loads. As a consequence, the characteristics of the harmonic compensation are strongly dependent on the filtering algorithm employed for the calculation of load current harmonics.

According to their principle of operation, the parallel active filters behave as harmonic current sources. Then, the implementation of a suitable current regulator is required. Hysteresis current regulators have been widely used for active filter applications because of their high bandwidth and simple structure. Alternatively, the implementation of current regulators can be realized by PWM-controlled Voltage Source Inverter (VSI) or by suitable switching rules [9]-[13]. However, with any type of current regulator, two current measurement systems are usually employed to sense the load and filter currents. Furthermore, powerful microprocessors are necessary for the on-line evaluation of the harmonic content of the load current.

In this paper two active power filter configurations requiring very simple algorithms and single current measurement systems are analyzed. The former needs the measurement of the source current only and uses hysteresis current regulators. The latter is based on the relationship between the “source flux” and the “inverter flux”. The reference value of the “inverter flux” is calculated in order to achieve the compensation of the reactive power and load current harmonics, and it is synthesized by using a PWM-controlled VSI.

The standard active power filter configuration with load and filter current measurements is presented in section 3.1. The implementation schemes with current regulators based on hysteresis comparators or PWM-controlled VSI are considered as basic schemes. The principle of operation and the control strategies of active filters requiring the measurement of the source current only or the load current only are discussed in sections 3.2 and 3.3, respectively. These control schemes are compared with the corresponding basic schemes through numerical simulations performed by the “Simulink” tool of Matlab.

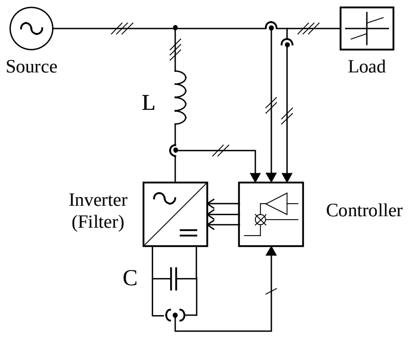

Fig. 1. Active power filter with load and filter current measurements

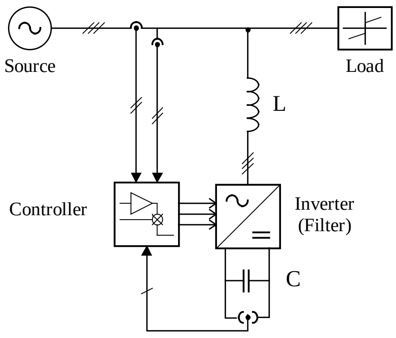

Fig. 2. Active power filter with source current measurement only

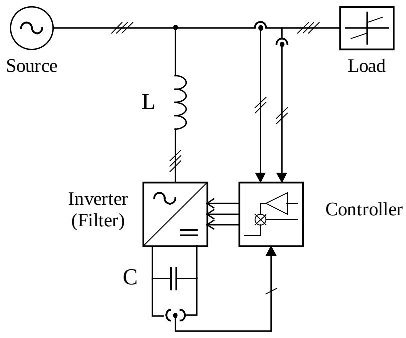

Fig. 3. Active power filter with load current measurement only

2. - ACTIVE POWER FILTER CONFIGURATIONS

As it is known, active power filters are used for compensating the reactive power and the low-order current harmonics generated by a non-linear load. The maximum harmonic order that an active filter is able to compensate determines the filter bandwidth. As a result of the active filter operation, the source currents become almost sinusoidal with unity power factor. Most of the solutions proposed to implement the control system are based on the measurement of both the load and filter currents. The additional measurement of the source currents has been also introduced in [14] leading to a solution with three current measurement systems. The source voltages are also required for the synchronization of the inverter output voltages or for coordinate transformations, depending on the employed control strategy.

A scheme of the standard solution with load and filter current measurements is represented in Fig. 1. The reference currents for the filter can be calculated by using the instantaneous active and reactive power of the load or by filtering the load currents. Sinusoidal currents in phase with the corresponding line-to-neutral voltages are also added to the reference currents to compensate the filter losses and to keep the dc voltage constant. The reference currents so obtained are compared with the measured currents and the error signal is used to control the active filter trough pulse-width modulators or hysteresis modulators.

In this paper the possibility that active filters can operate with a single current measurement system is investigated. Depending on whether the source currents or the load currents are measured, the block diagram shown in Fig. 2 or Fig. 3 must be considered.

For the analytical developments, the three-phase quantities are represented with space vectors according to a stationary d-q transformation.

In order to make a significant comparison, all the schemes analyzed are based on the same principle, i.e. the generation of a reference value for the source current $\bar{i}_S^*$ in phase with the source voltage $\bar{v}_S$, as represented in Fig. 4. The magnitude of the source reference current is obtained by a regulator acting on the instantaneous error between the reference dc voltage V_C^* and the actual dc voltage V_C measured across the dc capacitor C. The regulator has the transfer function R(s) and operates in order to keep the dc voltage close to the reference value. In these operating conditions, the average energy stored in the capacitor is constant (with the exception of the switching ripple) and no active power is absorbed by the active filter (with the exception of the switch and inductor losses). As a consequence, the active power is flowing directly from the source to the load. The reactive power required by the load represents a perturbation quantity that must be taken into account in the overall active filter design. A detailed description of the different control schemes is given in Sections 3.1, 3.2, and 3.3.

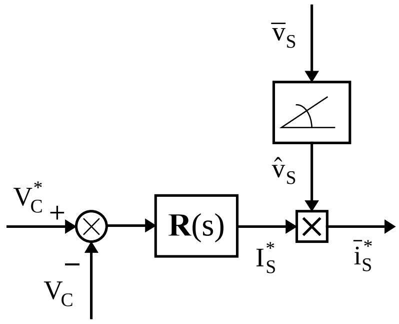

Fig. 4. Block diagram of the dc voltage regulator.

3. - CONTROL CIRCUIT ANALYSIS

In this section, the control circuits of the active filters represented in Figs. 1, 2, and 3 are analyzed and compared. These schemes use the same method for generating the reference value of the source current, whereas the reference signals for the inverter are determined by different methods, depending on whether the source current or the load current or both currents are measured.

It can be noted that the active filter performance is dependent on the time responses of the ac current and dc voltage control loops. However, the dynamic behavior of the filter is mainly affected by the time response of the ac current loop which must be designed in order to track the reference current waveforms closely. The dc voltage control loop has to keep the voltage close to its reference value and may not be very fast. Usually, the time response is several times greater than that of the current loop. Thus, the two control loops can be designed as two independent systems. A PI controller is employed for the voltage control loop since it acts in order to zero the steady-state error of the dc voltage.

3.1 - Control scheme based on load and filter current measurements

In this case the load current $\bar{i}_L$ is measured and the filter reference current $\bar{i}_F^*$ is given by

$$\bar{i}_F^* = \bar{i}_S^* - \bar{i}_L , \tag{1}$$

where the source reference current $\bar{i}_S^*$ is determined as previously discussed.

If the filter current $\bar{i}_F$ is also measured, a hysteresis current regulator acting on the instantaneous current error $\Delta\bar{i}_F = \bar{i}_F^* - \bar{i}_F$ can be employed to determine the inverter switch states. A simplified representation of the control scheme is shown in Fig. 5.

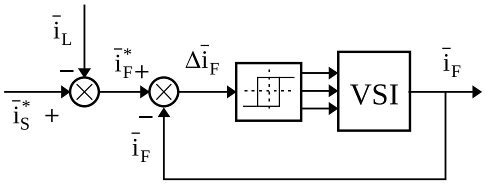

Fig. 5. Block diagram of the controller with hysteresis current regulator. The load and filter currents are measured.

A different solution, which allows hysteresis regulators to be avoided, can be obtained by generating the reference voltage for the PWM inverter through the filter current error. In particular, the filter reference voltage $\bar{v}_F^*$ can be calculated by the voltage equation written across the link inductance L leading to

$$\bar{v}_F^* = \bar{v}_S - L\frac{\Delta\bar{i}_F}{\Delta t} . \tag{2}$$

In (2) the resistance of the link inductance has been neglected. The block diagram of the control circuit related to this method is given in Fig. 6. It can be noted that in both schemes of Figs. 5 and 6 a closed loop control of the filter current is performed. With reference to Fig. 6, it is possible to determine the transfer function of the closed loop system. If we assume that the PWM inverter is able to generate the reference voltage at each cycle period, the following equation is obtained

$$\bar{i}_F = \frac{1}{1+\tau s}\bar{i}_F^* , \tag{3}$$

where $\tau=\Delta t$ is the cycle period. Eq. 3 shows that the response of the ac current loop is represented by a simple first order low-pass filter.

3.2 - Control scheme based on source current measurement only

In this case the control system acts in order to perform a direct regulation of the source currents which must be sinusoidal and in phase with the corresponding line-to-neutral voltages. According to the block diagram shown in Fig. 7, the source current error $\Delta\bar{i}_S$ is obtained through the source reference current $\bar{i}_S^*$, which is determined by the control scheme represented in Fig. 4, and the measured value $\bar{i}_S$.

The current error can now be utilized in different ways to

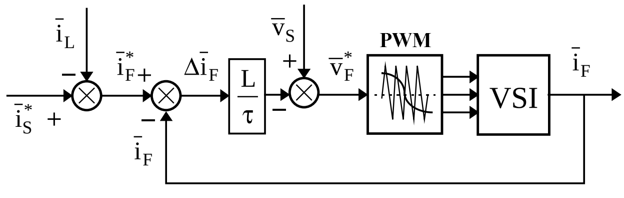

Fig. 6. Block diagram of the controller with PWM current regulator. The load and filter currents are measured.

generate the control signals for the inverter. In [15], the source current error $\Delta\bar{i}_S$ is used as input of a hysteresis regulator which directly provides the inverter with the switch commands. In [16] $\Delta\bar{i}_S$ is the input of an optimized fuzzy controller which is designed in order to give directly the inverter switch states. It can be noted that these current regulators exert a direct influence on the filter current, but they have the source current error as input variable. However, the active filter is able to operate correctly being the rate of change of the filter current usually higher than that of the load current.

3.3 - **Control scheme based on load current measurement only**

If the load current is measured instead of the source current, a different method is used to generate the control signals for the inverter. The block diagram representing the basic control principle is shown in Fig. 8. The voltage equation (2) is integrated leading to the following "flux equation"

$$\bar{\varphi}_F^* = \bar{\varphi}_S - L\,\bar{i}_F^*, \tag{4}$$

where $\bar{\varphi}_F^*$ is the reference "filter flux", and $\bar{\varphi}_S$ is the actual "source flux". The reference filter current is given by

$$\bar{i}_F^* = \bar{i}_S^* - \bar{i}_L, \tag{5}$$

where $\bar{i}_L$ is the measured load current.

It should be noted that $\bar{\varphi}_S$ cannot be estimated by an open-loop pure integration of the measured source voltage $\bar{v}_S$ owing to voltage transducer offset errors and the consequent integration drift. In [13] a special integration structure is introduced to overcome this problem. In order to simplify the control scheme, an alternative method can be employed. Assuming sinusoidal source voltages with a constant angular frequency ω, the steady-state relationship $\bar{v}_S = j\omega\bar{\varphi}_S$ is introduced in (4) yielding the following equation

$$\bar{\varphi}_F^* = \frac{\bar{v}_S}{j\omega} - L\bar{i}_F^*. \tag{6}$$

The actual value of $\bar{\varphi}_F$ can be evaluated by a pure integrator because it is used in a closed loop. Then $\bar{\varphi}_F$ can be expressed as

$$\bar{\varphi}_F(t) = \bar{\varphi}_F(o) + \int_o^t \bar{v}_F\,dt, \tag{7}$$

where the inverter output voltage $\bar{v}_F$ can be calculated on the basis of the dc voltage V_C and the inverter switch states S_A, S_B, S_C by

$$\bar{v}_F = \frac{2}{3} V_C \left(S_A + S_B e^{j\frac{2}{3}\pi} + S_C e^{j\frac{4}{3}\pi} \right). \tag{8}$$

In this way, the non-idealities of the VSI (switch voltage drops and dead-times) are neglected leading to small steady-state errors, but avoiding the use of additional wide bandwidth voltage transducers to sense the inverter output voltages.

From (6) and (7) the error flux $\Delta\bar{\varphi}_F = \bar{\varphi}_F^* - \bar{\varphi}_F$ can be readily calculated and used to generate the reference output voltage of the inverter by $\bar{v}_F^* = \Delta\bar{\varphi}_F / \Delta t$. The corresponding inverter switch states and the on-time ratios can be obtained by the PWM technique.

From Fig. 8 it can be noted that the filter current control is obtained by an open loop system. In this case the feedback variable is the filter flux. Assuming an ideal PWM inverter it is possible to determine the transfer function for the filter current. Taking (2) and (5) into account, the following equation can be derived

$$\bar{i}_F = \frac{1}{1+\tau s}\left(\bar{i}_F^* + \frac{\tau}{L}\bar{v}_S \right). \tag{9}$$

The transfer function contains two terms. The former is the same term of (3) and represents the low-pass filter action applied to $\bar{i}_F^*$. The latter is a perturbation term which does not affect the system stability, but introduces a steady-state error in the filter current. This error can be compensated subtracting the quantity $\bar{v}_S \cdot \tau / L$ to $\bar{i}_F^*$ in the diagram of Fig. 8. It could be noted that the stability of the active power filter depends on its ability to keep the dc voltage close to the reference value.

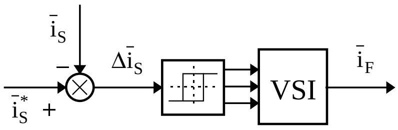

Fig. 7. Block diagram of the controller with hysteresis current regulator. The source current only is measured.

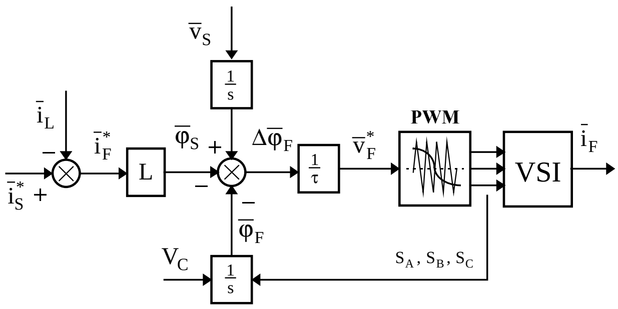

Fig. 8. Block diagram of the controller with PWM current regulator. The load current only is measured.

4. - NUMERICAL RESULTS

The control schemes for active power filters represented in Figs. 5-8 have been simulated by Simulink of Matlab. Ideal sinusoidal source voltages (380 V, 50 Hz) and ideal inverter switches have been assumed. Trapezoidal currents have been considered to represent the behavior of a 30 kVA, three-phase thyristor rectifier, controlled with a firing angle α=30°. As it is known, the ac line inductance increases the commutation overlap angle, reducing the di/dt and the active filter bandwidth requirements. For the numerical simulations a commutation time of 600 μs has been considered at rated conditions. The three-phase link reactor parameters are L= 2 mH, R= 0.1 Ω. The reference dc voltage is V_C^* =750 V and the dc bus capacitance is C=1 mF.

The performance of the various active filter configurations has been evaluated in steady-state and transient operating conditions. The simulation results are shown in Figs. 9-12. The switching frequency is 10 kHz for the schemes with PWM modulators. A current band of about 2 A, corresponding to an average switching frequency of about 12 kHz, is adopted for the schemes with hysteresis modulators. Similar values for both the switching frequency and the current ripple are suitable for a comparison among different solutions. All the schemes have been simulated with the same dc voltage regulator. In particular, the PI controller transfer function was

$$R(s) = P + \frac{I}{s}, \text{ with } P = 0.6 \text{ and } I = 100.$$

During the first 40 ms the active filter is operating in steady-state conditions, then a step change of the load current from 20 A to 40 A (0.5 to 1.0 p.u.) has been applied. The results obtained show that all the control algorithms analyzed are fast enough to respond to the load transient, keeping the source currents in phase with the corresponding line-to-neutral voltages and with a low harmonic content.

After 80 ms the load is switched off, causing a temporary increase in the dc bus voltage. However, the average capacitor voltage is kept within acceptable values, otherwise a clamping circuit should be applied at the dc bus to limit the maximum value. The reference value of dc voltage is recovered in less than 20 ms for the control schemes with two current measurement systems. The active filter topologies with only one current measurement system require about 20 ms using hysteresis modulators and about 40 ms using PWM modulators. It should be noted that the performance of each scheme can be further improved by properly adjusting the parameters of the dc voltage regulator.

5. - CONCLUSIONS

Two control schemes for active power filters which impose sinusoidal source currents in phase with the corresponding line-to-neutral source voltages have been analyzed. The control techniques are based on a single current measurement system, leading to control schemes which are very simple and can be easily implemented. The filter current regulation is obtained by using both hysteresis and PWM modulators. In the last case the inverter switch states are determined by the relationship between the “source flux” and “inverter flux”. The performance of the two control schemes is quite similar to that of standard schemes in steady-state operating conditions. Few differences appear in transient conditions.

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Numerical results in steady state and transient conditions

(a)- dc voltage, V_C **(b)**- line-to-neutral voltage, v_S **(c)**- load current, i_L **(d)**- compensated source current, i_S

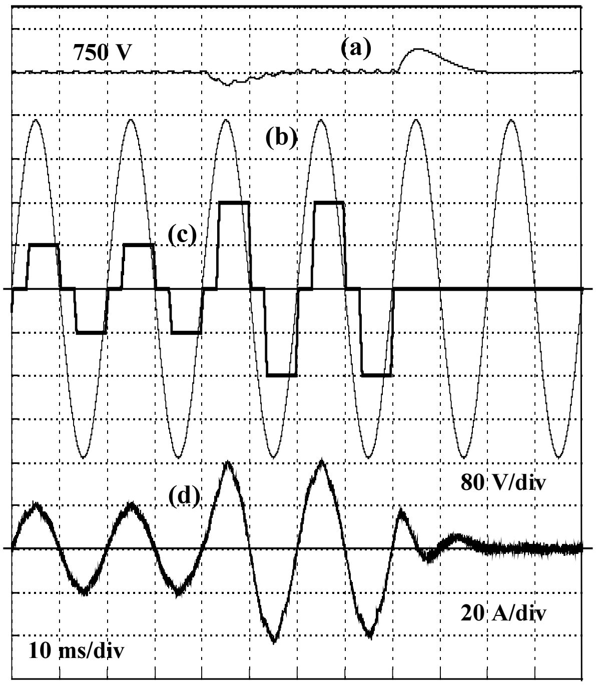

Fig. 9. Active filter with load and filter current measurements. Hysteresis current regulator (filter current as reference).

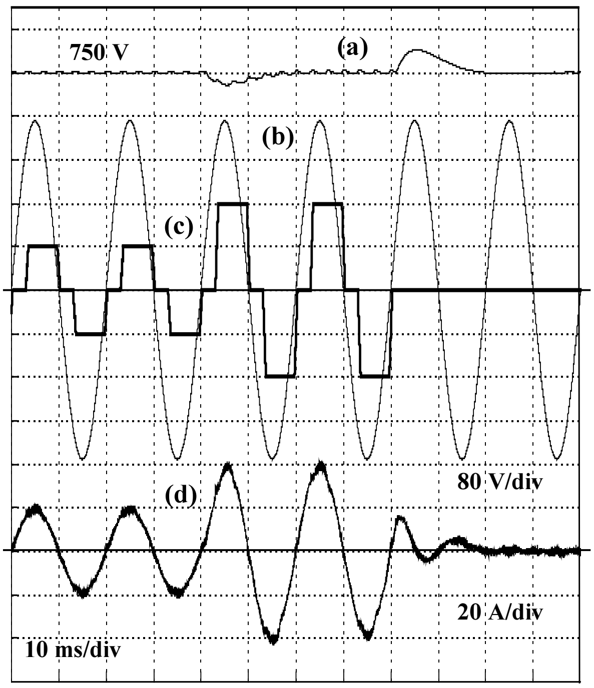

Fig. 10. Active filter with load and filter current measurements. PWM current regulator (voltage method).

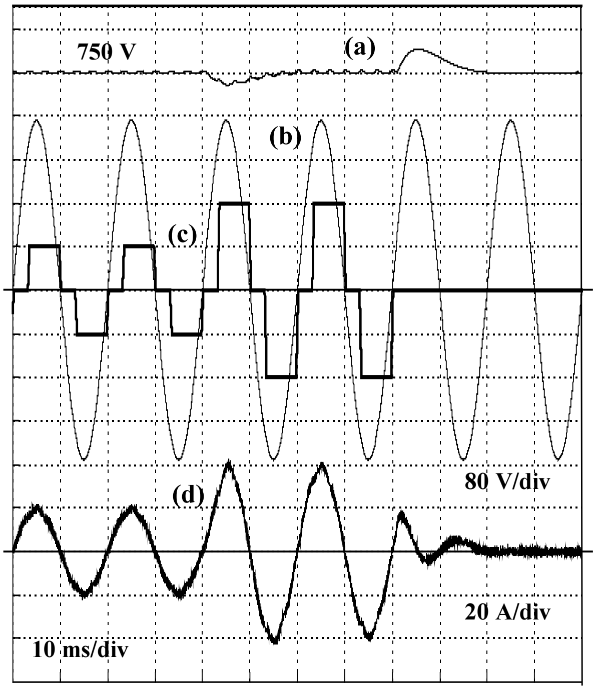

Fig. 11. Active filter with source current measurement only. Hysteresis current regulator (source current as reference).

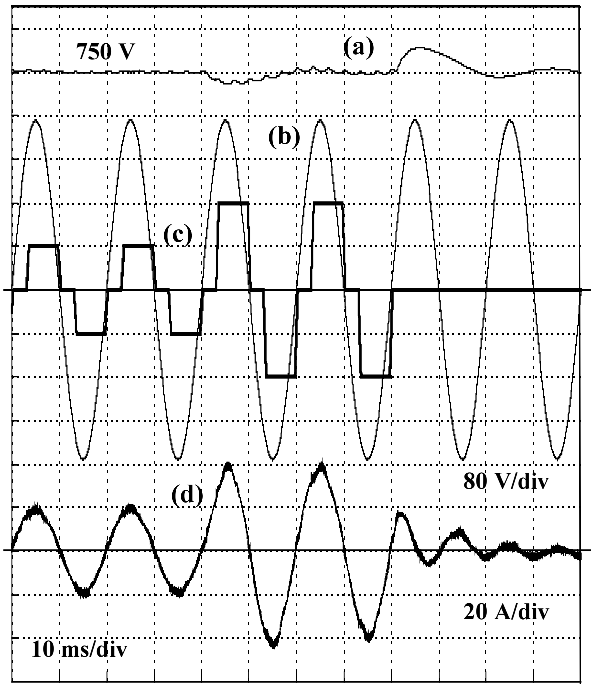

Fig. 12. Active filter with load current measurement only. PWM current regulator (flux method).