Using Nonlinear Load DC Variables as Inputs for Harmonic Current Reduction

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Abstract. This paper introduces and discusses new inputs for a single-phase shunt active power filter (SAPF) using nonlinear load equivalent resistance estimation control strategy. The inputs are an average DC voltage and current of a current source nonlinear load. By using a MATLAB Simulink simulation tool, a model of single-phase AC system connected to the SAPF and nonlinear load is designed and its vital voltage and current variables are simulated. Various simulation results are presented and analysed in order to validate the effectiveness of proposed inputs. As a conclusion, the proposed control strategy with new inputs successfully achieved its objective to mitigate harmonic current and to compensate the reactive power with good performance.

1. Introduction

Active power filter (APF) especially in parallel or shunt configuration (SAPF) has been the main tool in mitigating harmonic current distortion recently. The SAPF is connected in between the source and the nonlinear load at point of common coupling (PCC) as depicted in Figure 1. It is the typical configuration for single or three phase system implemented for harmonic current reduction and reactive power compensation. Basically, some advantages of using SAPF reported in literatures are based on its basic configuration, good reactive power and voltage flicker compensations capability etc [1]. In addition, a voltage source inverter (VSI) based SAPF is widely utilized [2-5]. Normally, the choice of topology for SAPF is H-bridge or multilevel VSI which bidirectional switches such as combination of MOSFETs/diodes or IGBTs/diodes are frequently selected as shown in Figure 1. Besides, many of the aforementioned topologies also used an electrolytic DC capacitor as their energy storage element. On the other hand, a ripple or filter inductor is applied to interface the VSI and PCC [6, 7].

Nowadays, many loads are considered nonlinear loads. Typical single-phase nonlinear load choice is a full bridge diode rectifier with inductive or capacitive load [8]. The nonlinear load is known as a source of harmonic current, and it draws reactive power from the line network. As a result, a source current, i, changes from purely sinusoidal waveform into non-sinusoidal waveform. In addition, the phase angle, θ gap between the i, and its reference, a source voltage, v, becomes wider which result in much lower power factor for the system that will greatly affect the system efficiency.
There are some poor impacts occurred due to harmonic current distortion, e.g. heating of conductor, shorten apparatus life expectancy, higher peak current etc. In order to overcome the problems created by harmonic current, a reliable and efficient control strategy for SAPF is required. Before a final product of control strategy which is a proper gating signal, \( g \) being fed into the switch, three different control stages are required to be passed through as depicted in Figure 2 [9].

The first stage is called harmonic detection method where the selected variable, load current, \( i_L \) which contains rich harmonic component is being filtered using low pass filter (LPF) to develop the fundamental component only. Next, to synchronize the produced signal with the \( v_s \), both are multiplied together to form a sinusoidal reference current, \( i_s^* \). Later, the \( i_s^* \) subtracted by the actual \( i_L \) to generate compensated reference current, \( i_c^* \).

The following stage is the control design where the newly generated \( i_c^* \) minus its actual signal \( i_c \) to yield an error, \( e \). This error needs to be compensated using a controller to obtain a control signal, \( u \). The desired response of \( u \) should have a zero steady state error, stable closed loop system, no percentage overshoot as well as fast settling time. In other word, the final value of \( u \) should be identical to its reference without any error \( (u \approx i_c^*) \).

The final stage of control strategy adopted by the SAPF is switching scheme technique. Typically, a linear bipolar pulse width modulation (BPWM) is selected. The \( u \) is compared with the high frequency triangular (carrier) signal to yield the \( g \). Sometimes other than the BPWM, hysteresis PWM and space vector PWM were also chosen as the preferred switching scheme technique [8].

The step by step procedure in determining the best option for simplified control strategy has been discussed briefly. Nevertheless, the issue is, almost all the inputs of APF control strategy is a load current, source current and DC-link voltage [1-8]. For many years, in particular for many literatures [1-10], the inputs for SAPF control strategy are the same, without any effort to change it. The exceptional characteristic for the input of SAPF control strategy is that it should have contained a huge amount of
harmonic component apart from the fundamental component. Previously, the inputs are dominated by the aforementioned variables which have proven their capabilities for SAPF control strategy, and is categorized as a source of harmonic. Another identical characteristic that is imposed by the previous inputs is prior to be inserted in LPF or other type of filters, its initial AC signal should be transformed into a DC signal first. Particularly, this happened to synchronous reference frame (SRF) and instantaneous reactive power reactive (PQ) theory control strategies for APF [1-5].

Therefore, if the typical source of harmonic is replaced by a DC load voltage and current due to some concrete reasons and as the alternatives for the previous one, which will be explained next. Rapid variation identified on the DC load voltage and current waveforms show that they comprise of high frequency component which in turn will provide harmonic current to the AC system. Besides, since they are already the DC signal, there is no requirement for transformation from AC to DC signal. Moreover, the nonlinear load will also draw reactive power from the supply. Therefore, based on these arguments, the DC voltage and current of nonlinear load has a merit to be implemented as the input of control strategy.

The objective is to implement two new inputs e.g. DC load voltage and current for SAPF control strategy using the nonlinear load equivalent resistance estimation technique proposed in [6]. Next section will provide the methodology and mathematical expression used in suppressing harmonic current and improve the system’s power factor. Results and analysis will be given after that, in order to verify the proposed method performance. Finally, the paper ends with conclusion and acknowledgement to the project grant provider.

2. Methodology
Methodology applied in the proposed control strategy can be divided into three sections, where the first section will elaborate about the harmonic detection method, then the control design, and finally the choice of switching scheme.

2.1. Modelling harmonic detection method
By applying the current controlled mode, the compensating current \( i_c \) is injected into the line at point of common coupling (PCC) used to force the source current \( i_s \) to become pure sinewave and makes in phase with the source voltage \( v_s \), which in turn improves the system power factor to unity. Basically, the currents of SAPF system can be expressed as follows:

\[
i_s = i_L + i_c
\]

where \( i_L \) represents the nonlinear load current. According to [9], since the H-bridge VSI configuration is adopted by the SAPF, the relationship between DC link voltage \( v_{dc} \) and the source inductor voltage \( v_s \) when the average inductor voltage \( v_L = v_{Lc} + v_{Ls} = 0 \) for one switching cycle can be summarized as follows:

\[
v_{dc} = \frac{v_s - v_R}{1 - 2D}
\]

\[
\frac{dv_L}{dt} = \frac{(v_{Lc} - v_R) + (2D - 1)v_{dc}}{L}
\]

where \( v_{Lc}: \) filter inductor voltage; \( v_{Ls}: \) source inductor voltage; \( v_R: \) \( v_L \) ESR voltage and \( D: \) duty ratio. Equation (3) presents the dynamic representation ofthe SAPF.

Since nonlinear load is the main source of harmonic generation, hence the mathematical expression that relates the nonlinear load and harmonic current is discussed first. Nonlinear load such as H-bridge diode rectifier with inductive load is selected as the source of harmonic which has been depicted in Figure 1. Thus, the full-wave rectified sinusoidal voltage can be expressed as a Fourier series [9]:

\[
v_o(t) = V_o + \sum_{n=2A...}^{\infty} V_n \cos(n\omega t + \pi)
\]

where the latter represents an even harmonic series; \( V_m: \) peak AC voltage, \( \omega: \) fundamental frequency (rad/s), and \( n: \) harmonic number. The average DC term voltage \( V_o \) is represented by:
\[
V_o = \frac{2V_m}{\pi}
\]

which high frequency component voltage \(V_n\) is:
\[
V_n = \frac{2V_m}{\pi} \left( \frac{1}{n-1} - \frac{1}{n+1} \right)
\]

Using superposition method, the DC load current \(i_o\) is calculated by taking each frequency separately before combining the results. Hence, the DC load current is the combination of two components, i.e. the average DC-term current \(I_o\) and high frequency component current \(I_n\).

\[
I_o = \frac{V_o}{R_o}, \quad I_n = \frac{V_n}{Z_n} = \frac{V_n}{|R_o + jn\omega L_o|}
\]

where \(Z_n\) denotes the load’s impedance with respect to \(n\). Based on nonlinear load equivalent resistance \(R_e\) estimation technique, the \(i^*_c\) generation can be easily developed. A simple technique to generate the \(i^*_c\) is depicted in Figure 3 since it does not require any tedious mathematical transformation which will save a lot of elapsed time needed for calculation.

\[
\text{Figure 3. The control strategy with the proposed inputs}
\]

Another vital component that has great impact in separating the harmonic component and fundamental component of the load voltage and current is the LPF application. Typical LPF allows a signal below its cut-off frequency \(f_{co}\) to pass through, albeit blocks the signal above its cut-off frequency. In this case, the choice of \(f_{co}\) is 1 Hz. An expression of a first order Butterworth LPF in Laplace transform is presented as follows:

\[
LPF(s) = \frac{1}{\tau s + 1}
\]

where \(\tau\) = a time constant, and \(\tau = 0.159\) s.

2.2. Controller design and switching scheme

The generated \(i^*_c\) signal is used as the reference for the actual compensated current, \(i_c\) signal to follow it traits exactly. Therefore, the difference between both signals yield the error, \(e\). After that, the \(e\) is controlled through a PI controller such that its product, a control signal \(u\) tracks the reference signal exactly. To generate a gating signal \(g\), a bipolar pulse width modulation (BPWM) switching scheme is selected, where the \(u\) signal is being compared with the carrier signal in order to form the \(g\). Then, it is fed to the gate of MOSFET switches in two states to operate the SAPF circuit.

Normally, to design the controller for the SAPF, the time domain mathematical representation of SAPF initially is changed to a Laplace transform. Figure 4 illustrates the closed loop control for the compensated current \(i_c\). Hence, the forward transfer function \(G(s)\) and the closed loop transfer function \(C(s)\) (assuming a unity feedback) for the \(i_c\) is expressed as:

\[
G(s) = \frac{e(s)}{i_c(s)} = PI(s) \times PWM \times \frac{1}{L_c s + R_c}
\]

\[
C(s) = \frac{G(s)}{1+G(s)} = \frac{5(K_p s + K_i)}{L_c s^2 + (K_p + R_c)s + K_i}
\]
\[ C(s) = \frac{5K_p}{L_c} \frac{(s + K_i/K_p)}{s^2 + (K_p + R_c)s/L_c + K_i/L_c} \]

where,

\[ PI(s) = \frac{K_p s + K_i}{s} = \frac{K_p}{s} \] (PWM = 5)

\[ \omega_n = \frac{1}{\sqrt{(L_c + L_d)c_d}} \] (11)

A linear BPWM has been chosen as the switching scheme, where its carrier frequency is set at 20 kHz. The \( u \) signal is being compared with the carrier signal to generate a series of pulse train which also known as the gating signals \( g \), using logic circuit. This signal is then being split into two opposite pulses, when one at ON state, the other will be OFF by inverting the \( g \) signal. Both signals are fed to each complement pair of switches’ gates in order to turn-on and turn-off the switches according to the provided gating signal state. To measure the quality and total harmonic distortion (THD) of source current as well as the system power factor PF, the following expression is implemented.

\[ \text{THD} = \frac{\sum_{n=1}^{\infty} |I_{n,rms}|}{I_{1,rms}} \] (12)

\[ \text{PF} = \cos \theta \cdot \frac{1}{\sqrt{1+\text{THD}^2}} \] (13)

where \( \cos \theta \) is referred to a displacement power factor and the second notation of (13) represents a distortion factor (DF).

3. Simulation results and discussion

The MATLAB Simulink simulation tool is implemented to investigate and verify the proposed inputs of control strategy for harmonic current mitigation and reactive power compensation using SAPF. The model of test system is designed based on Figure 1 where the system parameter is listed in Table 1. By
employing fixed step discrete solver and the sampling time was set at 5 µs, the simulation ran about 1 s. A Fast Fourier Transform analysis has been used to analyse the source current THD.

Table 1. System parameter for simulation

| Parameter            | Value                                    |
|----------------------|------------------------------------------|
| Source voltage       | \( V_s = 110 \, \text{V}_{\text{rms}} \)  |
| System frequency     | \( f_1 = 50 \, \text{Hz} \)                |
| Line impedance       | \( L_s = 0.5 \, \text{mH}, \, R_s = 0.05 \, \text{Ω} \) |
| Diode rectifier      | \( L_o = 50 \, \text{mH}, \, R_o = 20 \, \text{Ω} \) |
| SAPF                 | \( L_s = 3 \, \text{mH}, \, R_s = 0.1 \, \text{Ω}, \, C_{\text{dc}} = 470 \, \mu\text{F} \) |
| Switching frequency  | \( f_s = 20 \, \text{kHz} \)            |
| Proportional gain    | \( K_p = 3.2 \)                          |
| Integral gain        | \( K_i = 1824 \)                         |

The main objective of SAPF application is to establish pure sinusoidal source current as well as to fix it becoming in phase with the source voltage. In order to show the effectiveness of the proposed control strategy in compensating reactive power and reducing the total harmonic distortion (THD) under 5%, the \( i_s \) waveform and THD spectrum are depicted in Figure 5. As a result, after SAPF connection, the \( i_s \) becomes exactly a sinewave with its THD around 4% from 28% previously and in phase with the \( v_s \). In addition, the system PF also improved from 0.931 to 0.999. The reduction of \( i_s \) THD and the increment of system PF caused the system active power increased from 542 W to 606 W, while the system reactive power reduced from 151 Var to nearly 0 Var. Therefore, the SAPF successfully compensated the system reactive power.

Moreover, the THD spectrum measured until the 40th harmonic order is depicted in Figure 6 shows a significant improvement after SAPF connection. The peak magnitude of fundamental harmonic current has increased significantly to 7.8 A after filtering, from 7.5 A before filtering. However, the rest of harmonic orders indicate significant magnitude reduction especially the dominant harmonics after filtering, which is expected for successful suppression of dominant harmonics after SAPF implementation. Careful observation shows that at the vicinity of highest harmonic order, its particular magnitude is greater when the SAPF is connected than the disconnected one. The reason could be the injection of high frequency \( i_s \) that incorporates with overmodulation PWM switching.

![Figure 5. Steady state response of the \( i_s \) and \( v_s \)](image-url)
As the SAPF connected to the mains, the waveforms of $i_L$, $i_c$, $v_o$ and $i_o$ are illustrated in Figure 7 and 8 respectively. The $i_L$ maintained its feature except the waveform becomes thicker than the previous one. One reason behind this is due to the injection of high frequency component of the $i_c$. From the observation, the THD of $i_c$ is higher than the $i_L$, but the magnitude of $i_c$ is lower than the $i_L$. The $i_c$ waveform is identical to its reference $i_c^*$ and owns rich of harmonic content.

On the other hand, the waveforms of $v_o$ also indicates an identical characteristic like the $i_L$, e.g. its line variation is thicker than before the connection of SAPF. Again, the reason is because the magnitude of higher harmonic order (40th harmonic order) has been increased a little bit from its origin like depicted by THD spectrum of Figure 6. Moreover, the high switching frequency of 20 kHz injected by the SAPF also contributes to the noisy waveform of $v_o$, besides the absent of a shunt capacitor across the load which can act as a natural filter for the $v_o$. Nevertheless, the waveform of $i_o$ is unchanged, since its waveform is being filtered by the $L_o$ all the time.

**Figure 6.** THD spectrum of the $i_c$ before and after filtering

**Figure 7.** $i_L$ and $i_c$ waveforms after SAPF connection
The output signals responses of LPF and their final product, the $R_e$ signal are depicted in Figure 9. The filtered signals are smoother and slicker compared to the original waveforms of $v_o$ and $i_o$ respectively. Their ripple magnitudes have been reduced significantly which only small variation left on the $V_o$ signal, while the $I_o$ signal features an almost constant DC signal. The settling time for $V_o$ took 0.5 s while the $I_o$ took only 0.2 s. The reason for slow transient response of $V_o$ in particular is due to the time delay introduced by the first order Butterworth LPF. Nevertheless, the $R_e$ signal indicates much faster response by having reached the value of 20 at less than 0.02 s. Since the $R_e$ is a ratio of $V_o$ and $I_o$, even during transient response, it correctly estimates the value of load resistance $R_o$, the purpose of estimation.

In order to have good control design, the controller must have a correct and accurate estimation of sinusoidal reference current $i_s^*$ and compensated reference current $i_c^*$. Using nonlinear load equivalent resistance estimation technique, the harmonic detection method is made simple with short mathematical expression. Therefore, Figure 10 depicts the $i_s^*$ and $i_c^*$ waveforms appearance which are periodical sine waveforms after SAPF connection

![Figure 8. $v_o$ and $i_o$ waveforms after SAPF connection](image1)

![Figure 9. $V_o$, $I_o$ and $R_e$ signals responses](image2)
wave and highly distorted waveform respectively, same as expected. By comparing the $i_s^*$ with $i_s$ and $i_c^*$ with $i_c$, clearly shown that the actual waveforms are identical to their reference signals in term of magnitude, phase and shape.

![Figure 10. $i_s^*$ and $i_c^*$ signals at steady state](image)

Finally, Figure 11 shows the signal response of $u$ and $e$ respectively during SAPF operation. Both signals exhibit a constant variation within the predetermined boundary which signifies the actual $i_c$ keeps tracking its reference signal accordingly and the control system is stable. As a result, the $u$ signal response indicates that the designed PI controller successfully compensated the $e$ and produced a good control signal for SAPF application. The $K_p$ and $K_i$ tuning definitely give a great impact on the SAPF dynamic response, accuracy and stability. Without accurate $K_p$ and $K_i$ tuning, the SAPF cannot achieve its goal to reduce current harmonic distortion and compensate the reactive power.

![Figure 11. $u$ and $e$ signals responses](image)

4. Conclusion
The new inputs of SAPF control strategy based on nonlinear load equivalent resistance estimation technique has been presented and discussed. The proposed control strategy has a simple design in term of harmonic detection method calculation, introducing the DC load voltage and current as the input variables of harmonic detection method [11], utilizing a PI controller, a bipolar PWM as well as a H-
bridge VSI configuration. Besides it used the basic control strategy developed in [6] but with different source of inputs. Thus, the cost to develop the SAPF prototype is low albeit the performance is on par with the other control strategies. It is validated by the simulation result, and has been shown that the equivalent resistance \( R_e \) can be estimated accurately which lead to attain a pure sinewave for source current and in phase with the source voltage. As a conclusion, the objective of the proposed control strategy has been achieved after the source current THD successfully obtained below the limit specified by the IEEE-519 standard.

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