Abstract: This paper introduces both RAC and FLC and strategy of shunt active power filters (SAPF) for power quality improvement. The proposed control schemes are implemented without load harmonic detection. The compensation constraints are obtained by regulating indirectly the currents of the power mains. The reference currents of SAPF are generated by the dc-link voltage controller based on the active power balance of system. In RAC the current control strategy is implemented by an adaptive pole-placement control strategy integrated to a variable structure control scheme. In FLC is based on defined linguistic rules and does not require any mathematical model of the system unlike the other traditional controller.

I. INTRODUCTION

The rapid use of power electronic controlled equipment in electrical distribution system offers highly nonlinear characteristics and produces voltage and current waveforms distortions called as ‘harmonics’ [1]. Some major problems in electrical distribution system due to harmonics are well enumerated in [2-6]. Distorted current and voltage waveform further affects other consumers connected to the same point of common coupling (PCC) by propagating these distortions in their premises. In order to overcome these power quality problems, passive and active compensation (filters) approaches is used. Passive compensation though simple approach for achieving above goals but they are associated with several drawbacks such as, resonance problems, bulky, tuned for particular frequency component, component ageing, etc. [7], [8]. For overcoming the drawbacks of passive filters and reducing power quality problems, a number of attempts have been made on the design, analysis and optimal control development of active power filter (APF). APF allows compensation of voltage harmonics (series APF), compensation of current harmonics (shunt APF) with reactive power compensation required by nonlinear load. This paper emphasis on elimination of current harmonics and reactive power compensation, hence shunt APF topology has been considered. Shunt APF harmonic current is equal but opposite in-phase with the required harmonic component of load current. Computation of reference current signal and switching pulse generation for voltage source inverter (VSI) is main task. Various control approaches such as instantaneous reactive power theory [2], synchronous reference frame theory, closed loop PI control [9], etc. are used for generation of reference current signal by sensing combinations of source voltage, load current, filter current and source current. More number of sensor increase complexity and not cost effective solutions. Hence, control algorithm should be such that which give it’s best performance under steady-state as well as transient load conditions with less sensor count. Duke et al. [10] have proposed simple synthetic sinusoidal generation technique by sensing load current which further modified by sensing source current only [11]. The conventional methods use PI controller to regulate the DC bus voltage of VSI. Nemours methods have been proposed to replace PI controller scheme such as optimal regulator control [12], sliding mode control [13], and model Reference adaptive control. However, the design of these mentioned controllers depends on derived accurate mathematical models which are difficult to obtain. Also, these models do not give satisfactory operation under varying load condition with increased nonlinearity [8], [14], [15]. Artificial intelligence is one of the key areas to solve such system complexity and make control more robust for Transient conditions. Neural network, Fuzzy logic, expert system, various other optimization methods are used for the improvement of power quality [15]. Fuzzy logic control (FLC) is one of the significant tools in control design originated by Zadeh [16]. The advantages of FLC over conventional controllers are high robustness, insensitivity to parameters variations, handling of non-linearity and independent on mathematical models [8], [15], [17]-[19]. These solutions are based on controllers designed for SAPF whose dynamic model has fixed parameters. However, the interaction between the load and line impedances may modify the dynamic model of the SAPF system. Moreover, the model parameters can vary especially when the load has random behavior. Therefore, a suitable solution for compensating these power quality problems consists in the use of current controller whose gains are adjusted by adaptation. Recently, some adaptive approaches have been introduced for dealing with the load parameters variation. Apart from the conventional SAPF control schemes, this paper deals with a robust adaptive control strategy and fuzzy logic controlled shunt APF for the elimination of current harmonics and reactive power compensation of nonlinear load. The control schemes are based on indirect current control scheme in which only source current is sensed for avoiding switching spikes [20]. Three phase voltage and current signal is sensed using two voltage and current sensors. The effectiveness and validity of the proposed FLC and RAC are verified through MATLAB simulation.
II. SYSTEM DESCRIPTION AND MODELING

Fig. 1. Basic diagram of an SAPF system.

Fig. 2. Equivalent circuit of the active power filter.

Fig. 2 shows the equivalent circuit of the SAPF system considering impedances in both the power grid and the load. In this system, the power grid is represented by the internal voltage $E_s$ connected in series with impedance $Z_s(n_s + jI_s)$. The load is represented by a Norton equivalent circuit, where the current generator $I_o$ represents the purely distorting current load and the impedance $(\eta + jI_s)$ models its associated passive components. The active filter is composed of a voltage source $V_f$ connected to the PCC by means of the filter impedance $Z_{sf}(\eta + jI_f)$. From the equivalent circuit of Fig. 2, considering the ideal case in which the power-grid impedance is negligible and the load is composed of purely current source $I_o$, that means $Z_s = 0$ and $Z_{l} = \infty$, the current $I_s$ can be calculated by

$$i_s = (V_s - V_f + Z_f I_o) Z_f$$

(1)

Where $V_s$ is the voltage of PCC. Notice that the voltage differences between the split point of dc-link of the VSI and power mains neutral point $n$, or load neutral point $n'$ can be eliminated by using that account that

$$\sum_{k=1}^{3} i_{sk} = \sum_{k=1}^{3} i_{lk} = \sum_{k=1}^{3} i_{nk} = 0$$

(2)

If the grid impedance $Z_s$ is not negligible, the grid current $i_s$ can now represented by

$$i_s = \frac{E_s - V_f + Z_f I_o}{Z_f + Z_s}$$

(3)

On the other hand, if the grid impedance is negligible ($Z_s = 0$) and the load impedance $Z_l$ is considered, the SAPF model given earlier can be rewritten as

$$i_s = \frac{(Z_f + Z_s)E_s - V_f + Z_f I_o}{Z_f}$$

(4)

Finally, if we consider the interaction between the power-grid impedance and load impedance, the currents of the power grid can be determined as

$$i_s = \frac{(Z_f + Z_s)E_s - V_f + Z_f I_o}{Z_f + Z_s + Z_l}$$

(5)

In (4) and (5), the behavior of the current $i_s$ is described by a second-order dynamical system. The filter impedances do not usually vary. Thus, in (4), the location of the system poles is modified only by the load impedance $Z_l$. In (5), the poles locations are affected by both grid and load impedances. Defining the polynomial $\Delta(s)$ given by

$$\Delta(s) = Z_f Z_1 + Z_f Z_1 + Z_f Z_1$$

(6)

Equation (5) can be rewritten as

$$i_s = \frac{(Z_f + Z_s)E_s}{\Delta(s)}$$

(7)

From (7), it is possible to verify that for regulating $i_s$ it is necessary to generate a suitable filter voltage $V_f$. The disturbance terms can be redefined as

$$i_s = \frac{(Z_f + Z_s)E_s}{\Delta(s)}$$

(8)

$$i_s = \frac{Z_f E_s}{\Delta(s)}$$

(9)

Making $I_{s} = t_i = I_s - I_o$ and introducing the impedances $Z_s = \eta + jI_s$, $Z_e = \eta + jI_f$, and $Z_{l} = \eta + jI_l$, the resultant SAPF model now can be written as

$$i_s = \frac{V_f Z_l}{\Delta(s)} = -\frac{1 + \frac{1}{\eta}}{\gamma_2 + \frac{1}{\gamma_1}}$$

(10)

where $\gamma_2 = l_f l_1 + l_1 l_2 + l_2 l_3$, $\gamma_1 = l_f \eta + \eta l_1 + l_1 \eta + \eta l_2 + l_2 \eta$, $\gamma_0 = \eta \eta$, $\eta = \eta \eta$, and $\eta = \eta \eta$. The three-phase transfer function in (10) can be transformed into equivalent orthogonal dq components through the $0dq$-123 conservative transformation, in the stationary reference frame, this results in the following transfer function for the $d^q$ system model

$$i_s = \frac{1 + \frac{1}{\eta}}{\gamma_2 + \frac{1}{\gamma_1}}$$

(11)

Where $I_{s} = I_{s} - I_{s} - I_{s}$ and the superscript $s$ refers to the stationary reference frame. Notice that the filter
model given by (11) has two poles and one zero. If we consider the very usual case in which \( z_2 < z_f < z_1 \), the following simplification can be done

\[
y_2 \approx l_f l_1 + l_1 l_2 \tag{12}
\]

\[
y_1 \approx l_f l_1 + \eta l_1 + I_f \eta + \gamma l_1 \tag{13}
\]

\[
y_0 = \eta \gamma + \frac{\gamma I_f}{l_f} \tag{14}
\]

Therefore, the order of (11) could be reduced, which results in

\[
\frac{i_f^2}{V_f^2} = \frac{b_y}{\ell_f + i_f} \tag{15}
\]

Where \( b_y = \frac{1}{\ell_f + i_f} \) and \( \ell_f = \frac{r_f + \gamma}{l_f + i_f} \).

The system model represented by (15) is the most general model in which all the impedances are considered. However, if we take into account the scenario in which the grid impedances are fixed and the load impedance can vary, the system model may change. Depending on how the load impedance varies, instability problems can be verified if the current controller does not employ any mechanism of adaptation. It must be worse if the load varies randomly. In this model, parameters \( \gamma, l_f, y_f, \), and \( \gamma_0 \) can vary as a function of the random behaviour of the nonlinear load or the grid impedances. Moreover, the SAPF system model has also un modelled disturbances \( I_{sdq}^2 \) and \( I_{sdq}^2 \), whose dynamics are imposed together by the nonlinear load and by the power mains.

### III. DESIGN OF SAPF DEVICES

To obtain a suitable design template for the power converter and passive components, it is necessary to determine the compensation requirements imposed by the load. The first step is to calculate the power rating for the power converter used in SAPF. This power rating can be obtained as a function of load characteristics as

\[
P_{\text{load}} = \frac{\sqrt{\sin 0_2^2 + \text{THD}^2}}{\sqrt{1 + \text{THD}^2}} S_{\text{load}} \tag{16}
\]

Where \( 0_2 \) is the minimum angle of the load power factor and “THD” is the maximum total harmonic distortion of the load currents. If only harmonic compensation is required, the power rating of the SAPF can be determined by

\[
P_{\text{SAPF}} = \frac{\text{THD}}{\sqrt{1 + \text{THD}^2}} S_{\text{load}} \tag{17}
\]

The design of the dc-link capacitor can be developed by computing the energy balance of the SAPF. Therefore, the active power that is injected into the SAPF system by the power converter is expressed as

\[
P_{\text{conv}}(t) = P_c(t) - P_1(t) = \bar{P}_{\text{conv}} + P_{\text{conv}}(\cos \omega_{\text{sw}}) \tag{18}
\]

Where \( \bar{P}_{\text{conv}} \) and \( P_{\text{conv}} \) are the dc and ac components of \( P_{\text{conv}}(\cos \omega_{\text{sw}}) \) respectively. Considering that the converter power loss can be approximated by a constant term \( P_{\text{loss}} \), the magnitude of \( P_{\text{conv}} \) can be given by

\[
P_{\text{conv}} = \frac{1}{2} (V_i I_i - V_{i(1)} \cos \omega_{\text{sw}}(1)) \tag{19}
\]

Where \( V_i, I_i, \) and \( I_{1(1)} \) are the rms of the grid voltage, the load current, and the fundamental of load current, and \( \cos \omega_{\text{sw}}(1) \) is the load power factor. Therefore, the energy stored in the dc-link capacitor can be determined as

\[
\frac{1}{2} c \Delta V_c^2 = (\bar{P}_{\text{conv}} - P_{\text{loss}}) \Delta t \tag{20}
\]

Thus, the voltage ripple of the dc-link is expressed by

\[
\Delta V_c = \sqrt{(V_i I_i - V_{i(1)} \cos \omega_{\text{sw}}(1)) P_{\text{loss}} \Delta t} \tag{21}
\]

The aforementioned equation shows that the ripple of the dc-link voltage can be computed as a function of the energy stored in the capacitor. Based on this ripple, it is possible to design the capacitors of the dc-link. The designs of filter inductors are based upon the following criteria.

1. Limiting the high-frequency components of injected currents.

2. The instantaneous \( \frac{di_f}{dt} \) generated by the active filter should be greater than the maximum \( \frac{di_f}{dt} \) of the load.

Based on these assumptions, the maximum ripple of the filter currents can be determined as

\[
\Delta i_f(t) = \frac{1}{2 \omega_{\text{sw}}} \left( \frac{1}{\ell_f} (v_{f1}(t) - v_{f2}(t)) \right) dt \tag{22}
\]

Where \( \ell_f \) is the filter inductance, \( v_{f1}(t) = q_{\text{sw}}(t) 2 \), and \( q_{\text{sw}}(t) \) is the switching function, which defines the duty cycle of power switches. Solving (22) for the worst case, the filter current ripple can be given by

\[
\Delta i_f(\text{max}) = \frac{2(2P_{\text{sw}})}{2 \omega_{\text{sw}} \ell_f} \tag{23}
\]

Where \( \omega_{\text{sw}} \) the amplitude of the power mains voltage is is vector and \( \omega_{\text{sw}} \) is the switching frequency of the power converter. This equation determines the value of filter inductance based on the first design criterion. The second determines the filter inductance based on the maximum derivative of the load current. This restriction can be expressed analytically by

\[
\left[ \frac{di_f}{dt} \right]_{\text{max}} \approx \frac{1}{\ell_f} \frac{P_{\text{sw}}}{2 \omega_{\text{sw}}} \tag{24}
\]

The restriction on the dc-link voltage for achieving the desired maximum derivative filter current can be established as
Based on the equations mentioned earlier, it is possible to design the power rating of the power converter and the main passive components employed on the implementation of a SAPF.

IV. CONTROL SCHEMES

Fig.3 presents the block diagram of the proposed robust control scheme for the SAPF. In this block diagram, the dc link voltage is regulated by a PI controller with anti-windup. It is done by generating the reference current \( i_{dc}^e \), which determines the system active power component. The phase angle of the power-grid voltage vector \( \theta_2 \) is determined by using a PLL. Thus, the \( dq \) reference phase currents can be obtained by \( i_{dc}^e = i_{dc}^e \cos \theta_2 \) and \( i_{dc}^e = i_{dc}^e \sin \theta_2 \), respectively. The reference current \( i_{dc}^e \) is defined in a manner to guarantee the active power balance of the SAPF system. The phase currents of the power grid are indirectly regulated by the proposed VS-APPC current controllers by generating proper active filter phase voltages \( v_{f1}^e \) and \( v_{f2}^e \), respectively. These VS-APPC current controllers will be described next.

In order to estimate the parameters \( a_z \) and \( b_z \) consider the SAPF filter \( dq \) current–voltage first order plant [see Eq. (15)] described by:

\[
\frac{d i_{dq}}{dt} = a_z i_{dq} - b_z v_{f dq}
\]

An adaptive law can be obtained for generating estimates \( \hat{a}_z \) and \( \hat{b}_z \) by using the observed signals \( v_{f dq} \) and \( i_{dq} \). Considering an arbitrary positive constant \( \alpha_m > 0 \), we can rewrite (31) by subtracting the term \( \alpha_m i_{dq} \) as:

\[
\frac{d \hat{e}_{dq}}{dt} = \alpha_m \hat{e}_{dq} + (\alpha_m - \hat{a}_z) i_{dq} - \hat{b}_z v_{f dq}
\]

The amplitude \( \alpha_m \) defines the convergence speed of the estimated currents \( \hat{e}_{dq} \) [28]. The estimation error can be defined by:

\[
\varepsilon_{dq} = i_{dq} - \hat{i}_{dq}
\]

In the traditional indirect APPC scheme, adaptive laws driven by the errors \( \varepsilon_{dq} \) are used for generating estimates \( \hat{a}_z \) and \( \hat{b}_z \). Here, the parameters \( a_z \) and \( b_z \) can be estimated by using the following switching laws:

\[
\hat{a}_z = -\alpha_s \text{sgn}(\varepsilon_{dq} i_{dq})
\]

\[
\hat{b}_z = \text{sgn}(\varepsilon_{dq} v_{f dq}^e) + b_z(\text{nom})
\]

Since the following restrictions are satisfied \( \alpha_s > |b_z| \) and \( \tilde{\alpha}_s > |\frac{b_z}{b_z(\text{nom})}| \) with \( b_z(\text{nom}) \) being the nominal value of \( b_z \). This guarantees that \( \varepsilon_{dq} = 0 \) and those are the

\[
\varepsilon_{dq} = 2(i_{dq} \frac{di_{dq}}{dt})_{\text{max}} + \psi_2
\]
globally asymptotically stable equilibrium points. The pole placements and the tracking objectives of the proposed VS-APPC are achieved if the following control law is employed

$$Q_m(s) L(s) \frac{V_{f_{dq}}(s)}{I_{f_{dq}}(s)} = -P(s) \left( I_{f_{dq}}^2(s) - I_{f_{dq}}(s) \right)$$  \hspace{1cm} (36)

This addresses to the implementation of controllers transfer functions

$$T_{f_{dq}}(s) = \frac{P(s)}{Q_m(s) L(s)}$$  \hspace{1cm} (37)

Where $Q_m(s)$ is the internal model (IMP) of reference currents $I_{f_{dq}}^2; P(s)$ and $L(s)$ are the polynomials (with $L(s)$ monic). $Q_m(s)$ is chosen to satisfy $Q_m(s) I_{f_{dq}}^2 (s) = 0$. For the first order SAPF current–voltage control plant [see (15)] and considering that the VS-APPC control algorithm is implemented on the stationary reference frame, which results in sinusoidal reference currents, a suitable choice for the controller polynomials are $Q_m(s) = s^2 + \omega_c^2$, $L(s) = 1$, and $P(s) = \beta_2 s^2 + \beta_1 s + \beta_0$, where $\omega_c$ is the angular frequency of the voltage vector of power grid. This choice results in a current controller with the following transfer function

$$T_{f_{dq}}(s) = \frac{\beta_2 s^2 + \beta_1 s + \beta_0}{s^2 + \omega_c^2}$$

By solving the Diophantine equation for desired Hurwitz polynomials $A_2^*(s)$, the coefficients $\beta_2, \beta_1$ and $\beta_0$ can be determined by

$$\beta_2 = \frac{a_2^2 - a_3}{b_3}$$  \hspace{1cm} (38)

$$\beta_1 = \frac{a_1^2 - \omega_c^2}{b_3}$$  \hspace{1cm} (39)

$$\beta_0 = \frac{a_0^2 - \omega_c^2 a_3}{b_3}$$  \hspace{1cm} (40)

The control signals $v_{f_{dq}}$ generated at the outputs of VS-APPC controllers can be determined by using (37) as

$$\frac{dx_1^2}{dt} = x_1^2 + \beta_1 e_{f_{dq}}$$  \hspace{1cm} (41)

$$\frac{dx_2^2}{dt} = \omega_c^2 x_1^2 + (\beta_0 - \omega_c \beta_2) e_{f_{dq}}$$  \hspace{1cm} (42)

$$v_{f_{dq}} = x_1^2 + \beta_2 e_{f_{dq}}$$  \hspace{1cm} (43)

Where $e_{f_{dq}} = i_{f_{dq}}^2 - i_{f_{dq}}^2$

**2. Design of the Current Controllers**

For designing the proposed VS-APPC current controllers, the following steps are necessary.

1) **Step 1.** Identify the system impedances $Z_x, Z_f$ and $Z_1$ in order to calculate the parameters $a_2$ and $b_2$. The system model is obtained substituting $a_2$ and $b_2$ values in (15).

2) **Step 2.** Choose the current controller bandwidth $\omega_c$ and calculate the coefficients $a_0^*, a_1^*$ and $a_2^*$ from (28)–(30). Substitute $a_2^*, a_1^*$ and $a_0^*$ in (26) to find the Hurwitz polynomial $A_2^*(s)$.

3) **Step 3.** Choose a positive constant $\alpha_m$. A suitable choice is $\alpha_m > \omega_c$.

4) **Step 4.** Choose parameter $\alpha_x$ to be used in the switching law (34). The $\alpha_x$ value determines the
estimation range. A suitable choice is \( b_2 = 2 \alpha_2 \), where \( \alpha_2 \) is the value calculated in Step 1.

5) Step 5. Choose parameters \( b_2 \) and \( b_2 \) (nom) to be used in the switching law (35). These values determine the \( b_2 \) estimation range. A suitable choice is \( b_2 \) (nom) = \( b_2 \) and \( b_2 = 0.75 b_2 \), where \( b_2 \) is the value calculated in Step 1.

6) Step 6. Calculate controller parameters \( \beta_2 \), \( \beta_1 \) and \( \beta_0 \) from (38)–(40).

Based on the simulation and in the theoretical studies, it can be observed that the magnitudes of the switching laws \( \alpha_2 \) and \( b_2 \) determine how fast the VS-APPC controllers converge to their respective references. However, high values result in high amplitudes of control signals \( v_f \), which can address to the nonlinear behaviour of the SAPF control system.

3. DC-Link Voltage Controller

Fig. 5 shows the block diagram of the dc-link voltage control loop. Block \( R_p(s) \) refers to the standard \( PI \) controller which transfer function is

\[
R_p(s) = \frac{K(1+sT_i)}{sT_i} \tag{45}
\]

Where \( K \) is the gain and \( T_i \) is the integration time constant of the \( PI \) controller used in the dc-link voltage regulation. The algorithm of the \( PI \) dc-link voltage controller is implemented with an antiwind up scheme. The delay introduced by the current control loop is neglected and its representation on this diagram is omitted (see Fig. 5).

\[
\text{Fig. 5. Block diagram of the dc-link voltage control loop}
\]

To obtain a smooth current command \( i_{d2}^{*} \) at the output of the dc-link voltage regulator, a first-order low-pass filter is introduced in the dc-link voltage measurement, which is represented by the block \( G_v(s) \) given by

\[
G_v = \frac{1}{1 + T_p s} \tag{46}
\]

Where \( T_p \) is low-pass filter time constant used in the dc-link voltage measurement. The parcel \( i_{d2}^{*} \) related to the harmonic compensation (see Fig. 5.2) is considered as a disturbance to be compensated by the dc-link voltage controller. Therefore, the dynamic model of the dc-link together with the low-pass filter can be given by

\[
G_v = \frac{1}{C_2(1 + T_p s)} \tag{47}
\]

Thus, if the controller is included in the transfer function before, the dc-link voltage open loop is

\[
G_{tr} = \frac{K(1+sT_i)}{CT_p s^2(1 + T_p s)} \tag{48}
\]

4. Design Criteria for the DC-Link Voltage Controller

The design of the dc-link controller is determined by using the symmetrical optimum tuning optimization (SOTO) [29]. This method is based on the idea of finding a controller that makes the frequency response from the set point to the plant output as close as possible to one (0 dB) for low frequencies. The design method based on SOTO consists of two steps [29]. In the first step, the transfer function given by (47) is simplified to the following form

\[
G_{cp}(s) = \frac{K_\omega}{s(1 + sT)} \tag{49}
\]

Where \( K_\omega = 1/C \) and \( T = T_p \). The poles of dc-link voltage transfer function may be cancelled by controller zeros to obtain the loop transfer function. If the dc-link is controlled with a \( PI \) controller, the loop transfer function becomes

\[
G_{opt}(s) = \frac{K_\omega(s + T_p)}{s^2 T_p(1 + s T)} \tag{50}
\]

The SOTO transfer function suitable for the a 2-DOF controller is [29]

\[
G_{soto}(s) = \frac{\omega_0^2 (2s + \omega_0)}{s^2 + 2\omega_0 s} \tag{51}
\]

Where \( \omega_0 \) is the frequency response of \( G_{cp}(s) \). Notice that the Bode diagram of this transfer function is symmetrical around the frequency \( \omega = \omega_0 \). Therefore, to make the transfer function \( G_{soto}(s) \) [see (50)] to be identical to the symmetrical optimum \( G_{soto}(s) \) [see (51)], it is required that

\[
\omega_0 = \frac{1}{2T_p} \tag{52}
\]

With the controller gains given by

\[
K = \frac{C}{2T_p} \tag{53}
\]

\[
T_p = 4T_v \tag{54}
\]

This design procedure addresses to a theoretical step response of the closed-loop transfer function with an overshoot of approximately 8.1% and a settling time of 2% at the steady-state value \( i_{d2} = 9.4/\omega_0 \). Such performance is adequate for the dc-link control loop transient requirements.

**FUZZY LOGIC CONTROL**

Fuzzy logic becomes more popular due to dealing with problems that have uncertainty, vagueness, parameter variation and especially where system model is complex or not accurately defined in mathematical terms for the designed control action.
The conception of the fuzzy logic introduced by Zadeh [16] is a combination of fuzzy set theory and fuzzy inference system (FIS). Elements of a fuzzy set belong to it with a certain degree, called degree of membership. The degree of membership is a result of Mapping the input to certain rules using a membership function (MF). The progression which maps the specified input data to the output using fuzzy logic is known as fuzzy inference. A fuzzy inference system can be classified as

(a) Fuzzification: This is the process of converting any crisp value to analogous linguistic variable based on certain MF.

(b) Inference engine: simulates human decision,

(c) Knowledge base: consists of MF definitions and necessary rules like IF-THEN or it is combination of condition part with their associated rules

(d) Defuzzification: is the progression of transforming the fuzzy output into a crisp numerical value. In this paper main control input variable is the DC-link voltage error and output of FLC is the peak value of the reference source current. The range of operating current, normalization and de-normalization is one of the important design factors of fuzzy controller.

V. DESIGNING OF CONTROL RULES

Computational methods determine the computational efficiency, processor memory requirement and processing time. The fuzzy control rules based on membership function defining or relate input variables to output variables. The number and type of MF determines the computational efficiency of fuzzy control technique. The determination of MFs depends on the designer's experience and knowledge. The shape decision of MFs affects how well a fuzzy system rules approximate a function. Triangles or triangular membership function (TMF) have been frequently used in several applications of FLC [22], [23]. TMF are preferred due to simplicity, easy implementation, symmetrical along the axis. Fig. 7 shows the MFs relating input and output linguistic variables. The number of linguistic variables is directly related to the accuracy of approximating function and plays an important role for input-output mapping [24]. However, some limits have to consider while designing number of linguistic variables in view of accuracy and complexity of FLC. Triangles or triangular membership function (TMF) have been frequently used in several applications of FLC [22], [23]. TMF are preferred due to simplicity, easy implementation, symmetrical along the axis. Fig. 7 shows the MFs relating input and output linguistic variables. The number of linguistic variables is directly related to the accuracy of approximating function and plays an important role for input-output mapping [24]. However, some limits have to consider while designing number of linguistic variables in view of accuracy and complexity of FLC.

The error $e$ and change of error $ce$ at $n^{th}$ sampling instant can be written as

$$e = V_{dc,ref} - V_{dc}$$

$$ce(n) = e(n) - e(n-1)$$

The output of FLC with limiter is considered as amplitude of derived reference current ($I_{ref}$). In this paper Seven triangular membership functions have been chosen for representing numerical variables into linguistic variables, viz., NL (negative large), NM (negative medium), NS (negative small), ZE (zero), PS (positive small), PM (positive medium), PL (positive large). The spacing between MFs may be equal or unequal; it is set here for cover a band of load current with good accuracy. After this rules formation as knowledge base, different inference mechanisms have been developed for defuzzify fuzzy rules. In this paper, authors apply Mamdani's max-min inference method to get an implied fuzzy set of tuning rules. Finally centre of mass method is used defuzzify the implied control variables.

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<td>NS</td>
<td>Z</td>
<td>PS</td>
<td>PM</td>
<td>PB</td>
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<tr>
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<td>PB</td>
<td>PVB</td>
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<tr>
<td>PM</td>
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<td>PM</td>
<td>PB</td>
<td>PVB</td>
<td>PVB</td>
<td>PVB</td>
</tr>
</tbody>
</table>

Table 1. Fuzzy Control Rule Base

![Membership Curve](image_url)

(a) Error 'e' and change in error 'ce'
VI. SIMULATION RESULTS

RAC for Non-linear Load without compensation

If the compensation is not provided for balanced non-linear load without compensation source voltage, load current and source current have the harmonics. It is shown in above fig.

With Compensation

From the above fig, we can conclude that harmonics reduced in source current after the compensation.

CASE-2 FLC for Non-linear Load without compensation
HARMONIC ANALYSIS

CASE-1

<table>
<thead>
<tr>
<th>Order of Harmonics</th>
<th>Before Compensation</th>
<th>After Compensation</th>
</tr>
</thead>
<tbody>
<tr>
<td>3rd</td>
<td>2</td>
<td>0.2</td>
</tr>
<tr>
<td>5th</td>
<td>25</td>
<td>5</td>
</tr>
<tr>
<td>7th</td>
<td>12</td>
<td>2.5</td>
</tr>
<tr>
<td>9th</td>
<td>2</td>
<td>0.2</td>
</tr>
<tr>
<td>%THD</td>
<td>30.04</td>
<td>6.84</td>
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</tbody>
</table>

CASE-2

<table>
<thead>
<tr>
<th>Order of Harmonics</th>
<th>Before Compensation</th>
<th>After Compensation</th>
</tr>
</thead>
<tbody>
<tr>
<td>3rd</td>
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<td>0.4</td>
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<tr>
<td>5th</td>
<td>18</td>
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<tr>
<td>%THD</td>
<td>22.73</td>
<td>2.59</td>
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VII. CONCLUSION

This paper has presented a robust adaptive control and fuzzy logic control strategies of SAPF for power-factor correction, harmonic compensation, and balancing of nonlinear loads. The proposed control schemes are implemented without load harmonic detection and the compensation constraints are obtained by regulating indirectly the currents of the power mains. Due to the fact that the harmonic detection scheme is not used, the number of current sensors employed in the SAPF is reduced, which decreases the effective cost of SAPF implementation. The proposed current control strategy (VS-APPC) was demonstrated to be robust to the system parameters' variation, unbalanced conditions of mains, and the load.

REFERENCES


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