Three-phase four wire voltage controlled AC line conditioner with unity input power factor and minimised output voltage harmonics

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Abstract: The paper presents a new high performance AC line conditioner isolating any configuration of nonlinear loads for three-phase four wire applications. The proposed conditioner is composed of a switch mode rectifier (SMR) and a ROM based sinusoidal pulse width modulation (SPWM) inverter. The three-phase half bridge type SMR operates under voltage mode control. No current sensing is required. The inverter is also a half bridge type controlled by a ROM based SPWM modulator. Preoptimised SPWM control patterns rejecting effects of DC link voltage ripples and dead times are programmed. Owing to the nature of the three-phase scheme, the voltage imbalance problem between the DC link capacitors can be lessened. High quality three-phase AC outputs can be obtained while nearly sinusoidal input currents are drawn.

1 Introduction

More and more industrial applications require the supply of three-phase power source to nonlinear loads. These nonlinear loads cause problems in the power distribution system, such as low power factor, high input current harmonic contents and distortion in line input voltage waveform. This will result in the degradation of the power quality supplied to other loads. Standards and recommended guidance such as IEEE-519 and IEC-555 have put limits on specified harmonics in line current owing to load nonlinearities.

Several approaches for minimising line current harmonics in three-phase, three wire systems have been proposed [1–3]. Nevertheless, in some industrial and commercial applications, power is distributed through three-phase four-wire systems. Such systems may suffer from excessive neutral currents caused by nonlinear or unbalanced loads. A four wire, current controlled active filter proposed [4] can provide harmonic neutralisation. However, by using a variable hysteresis current mode controller, undesirable harmonics are present. Sensing large load currents in a high power system is ineffective

due to the impractical scaling down. Also the three-phase AC voltages are not guaranteed to be pure sinewaves.

In this paper, a three-phase four wire AC line conditioner is presented. The block diagram of the proposed system is shown in Fig. 1. The three-phase half bridge

Fig. 1 Block diagram of proposed line conditioner

SMR is built by paralleling three single phase modules [5] which achieve unity input power factor with a minimum of switches and low conduction losses. The SMR is controlled under voltage mode feedback [6]. No current sensing is required. Because the current waveform reference contained in the control signal is irrelevant with the AC source voltage, input current is still a pure sine-wave despite of distortion in source voltage. Two series connected capacitors provide a filtered DC voltage and energy storage. A ROM based SPWM inverter [7] converts the DC link voltage into fixed-frequency (50 or 60 Hz) three-phase AC sources. Preoptimised SPWM control patterns compensating DC link voltage ripple [8, 9] and dead time degradation [10, 11] are stored in the proposed modulator. The possibility of excessive load current in the neutral, due to nonlinear or unbalanced loads, is prevented by the DC link capacitor tank. The neutral current at the source end is the negative sum (ideally zero) of the three-phase input currents which are forced to follow a balanced situation. Both the neutral conductor and the transformer can be protected from damage. In the following Sections, control principles, the performance simulations and laboratory experiments of the new AC line conditioner will be discussed in detail.

2 Voltage mode control of SMR

A 1 ϕ half-bridge SMR is shown in Fig. 2A. In the positive half cycle of the AC voltage, input current increases when S2 is on and decreases when S2 is off and D1 is forced to conduct. In the negative half cycle the corresponding switches are S1 and D2. By turning on and off the respective switch for a proper duration, the input current can be controlled to follow a sinusoidal waveform. An equivalent circuit for a 1 ϕ SMR, derived by

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looking into the switching rectifier, is shown in Fig. 2B. Depending on the switching state, v_u could be E or $-E$, where E is the DC voltage of the output capacitor.

Fig. 2A 1φ SMR

Fig. 2B Equivalent circuit

From Fig. 2B the control principles are deduced as:
 $v_s(t) = v_{Ls}(t) + v_{u,1}(t)$ from Kirchhoff's voltage law, where $v_{u,1}(t)$ is the fundamental component of v_u . Let $i_s(t) = I_s \sin \omega t$ under steady state, then $v_{Ls}(t) = L_s di_s(t)/dt = \omega L_s I_s \cos \omega t$ and $v_{u,1}(t) = v_s(t) - \omega L_s I_s \cos \omega t$. From the transfer characteristics of the SPWM inverter and the theorem of superposition, to produce an AC voltage $v_{u,1}(t)$, the control signal has to be

$$v_{con} = v_{u,1}/E = \frac{v_s(t)}{E} - \frac{\omega L_s I_s}{E} \cos \omega t \quad (1)$$

I_s can be related to output voltage through a PI controller. That is $v_{con} = k_1 v_s(t) - k_2 \cos \omega t$, where k_1 is the parameter concerning the feedforward input voltage and k_2 concerning the feedback output voltage. Even when $v_s(t)$ is distorted, the input current still can be an ideal sine wave. The reason is that the waveform reference of input current subjects only to the cosine-term in control signal and is not affected by the source voltage. The block diagram is shown in Fig. 3.

Fig. 3 Block diagram of SMR voltage-mode control

A three-phase half bridge SMR composed of three single-phase modules is shown in Fig. 4. Three sets of control signals for each phase are generated through the same mechanism except for 120° phase lead or delay.

Fig. 4 Three-phase half-bridge SMR circuit

That is,

$$v_{con,a} = k_1 v_{sa}(t) - k_2 \cos \omega t$$

$$v_{con,b} = k_1 v_{sb}(t) - k_2 \cos(\omega t - 120^\circ)$$

$$v_{con,c} = k_1 v_{sc}(t) - k_2 \cos(\omega t + 120^\circ)$$

If the three-phase sources are purely sinusoidal, $v_{sa}(t) = V_s \sin \omega t$, $v_{sb}(t) = V_s \sin(\omega t - 120^\circ)$ and $v_{sc}(t) = V_s \sin(\omega t + 120^\circ)$, the cosine part in control signal of each phase can be obtained by summing the other two-phase sources. For example, $\cos \omega t = [1/\sqrt{3}][\sin(\omega t - 120^\circ) + \sin(\omega t + 120^\circ)]$. The block diagram of generating phase A control signal is illustrated in Fig. 5.

Fig. 5 Generation of phase-A control signal

3 ROM based SPWM inverter

Conventionally, control signals fed to the SPWM inverters are produced by electronic hardware or microprocessor. The cost and complexity of hardware implementation and online computation in the microprocessor approach are considerable. A ROM based SPWM modulator can solve the problems mentioned above. Unlike the previous works, in which the programmed contents are either digital representations of sampled sine wave heights or harmonic elimination PWM patterns with fewer notches, the proposed modulator stores the SPWM patterns computed offline in the EPROM. To program the analogue SPWM patterns into digital forms, each pulse is composed of several consecutive 1s to represent its width. The duration between pulses are represented by consecutive 0s. This digitised SPWM pattern can then be retrieved sequentially by incrementing the address counter. This modulator features the advantages of low cost, fast response, simple implementation and control. In this paper, the SPWM control signals are precalculated by computing the intersections of a sinusoidal reference wave and a triangular carrier wave with numerical method. A fixed frequency clock counts through the addresses to retrieve the stored SPWM control patterns.

With low order line frequency ripple on the DC link voltage, typically utilised SPWM techniques will produce unwanted low frequency harmonics in the inverter's output. They can be eliminated by modifying the modulation function [8, 9]. Assuming that the DC link voltage, V_{DC} , is ripple free, the output of the inverter is then an AC voltage modulated by a switching function with frequency spectrum as depicted in Fig. 6. Harmonics

Fig. 6 Spectrum of SPWM modulation function

are centred around the multiples of the switching frequency. This AC output can be written as

$$V_{AC} = V_{DC} \left[m_a \sin \omega t + \sum_{n+k=\text{odd}} l(m_f n \pm k) \sin((m_f n \pm k)\omega t) \right] \quad (2)$$

where m_a is the amplitude modulation index and m_f is the frequency modulation index. $l(m_f n \pm k)$ is the amplitude of the corresponding $(m_f n \pm k)$ th order harmonic. By filtering out the high-order harmonics, a fundamental AC voltage $V_{DC} m_a \sin \omega t$ can be obtained.

Let us now consider the case in which low-order harmonic $kh(\omega t)$ is superimposed on the DC link voltage, $V'_{DC} = V_{DC}(1 + kh(\omega t))$. The modulated AC output voltage then becomes

$$\begin{aligned} V'_{AC} &= V_{DC}(1 + kh(\omega t)) \left[m_a \sin \omega t + \sum_{n+k=\text{odd}} l(m_f n \pm k) \sin((m_f n \pm k)\omega t) \right] \\ &= V_{AC} + V_{DC} kh(\omega t) m_a \sin \omega t + \dots \end{aligned} \quad (3)$$

Undesirable low-order harmonics are produced owing to the multiplication of two low-order terms $kh(\omega t)$ and $m_a \sin \omega t$. By modifying the original modulation function to $m_a \sin \omega t / (1 + kh(\omega t))$, the effects of low-order ripple can be removed.

For a bridge type inverter, a short period of time is always included in control signals between turning one switch in an inverter leg off and turning the other one on. This is referred to as 'dead time' which prevents two switches of an inverter leg from simultaneous conduction. It has been found that the dead time causes a reduction in the fundamental component of the output voltage and introduces low-order harmonics which should not be present [10]. A load current dependent compensation method [11] can eliminate the output voltage harmonic distortion due to dead time effects. The principles of the compensation techniques are summarised here. When the load current is positive, the timing of the triggering signals to the upper switch in an inverter leg coincides with the ideal control patterns. Conversely, when the load current is negative, the firing signals to the lower switch coincide with the ideal requirements. In this way the output AC voltage waveform is almost the same as the original SPWM control patterns. An example of

Fig. 7 Switching waveforms for dead time compensation

a = ideal triggering signal for upper switch, b = ideal triggering signal for lower switch, c = control signal for upper switch when current > 0, d = control signal for lower switch when current > 0, e = control signal for upper switch when current < 0, f = control signal for lower switch when current < 0, V_o = resulting AC output voltage

switching waveforms for dead time compensation is shown in Fig. 7. The proposed method is impossible to implement on the existing system because it requires primary generation of control patterns for switches. However, the real time compensation can be performed on preprogrammable fixed frequency SPWM patterns.

In the proposed ROM based modulator, an optimised SPWM switching function compensating DC link voltage ripple and dead time effects can be offline calculated and preprogrammed. A fixed frequency oscillator with counters retrieves the SPWM patterns by counting through the addresses of the ROM. A reset signal is sent from the ROM to ensure a complete cycle. The schematic version of a three-phase inverter is shown in Fig. 8 and the block diagram of the proposed modulator is illustrated in Fig. 9.

Fig. 8 Three-phase inverter

Fig. 9 Proposed ROM-based SPWM modulator

4 Design considerations

For the three-phase SMR, the designed issues are the input current ripple and the switching frequency. The former decides the value of the THD, and the latter concerns the frequency distribution of the harmonics. The input current ripple is a function of switching frequency, input inductance, input voltage and output voltage. Since all other three factors are set, the only controllable is the switching frequency. The switching frequency is recommended to be high above audible noise, which is about 20 kHz. Another advantage is to push the harmonics far away from source frequency. Thus the switching frequency is high and constant under the voltage feedback control. These contribute to easy filtering of switching harmonics. The allowable switching speed and switching loss of a practical switch should be taken into account. Based on the considerations above, the switching frequency can be designed by analysing Fig. 10.

Fig. 10 Production of SPWM control signals

Since the input current varies sinusoidally, its switching ripple varies period by period. The current ripple can be defined as the average current variations in a switching period:

$$2\Delta I = [(dI/dt)_{inc} T_{sON} + (dI/dt)_{dec} T_{sOFF}] \quad (4)$$

$(dI/dt)_{inc}$ and $(dI/dt)_{dec}$ are the increasing and decreasing rates of input current. T_{sON} and T_{sOFF} are the on time and off time of the switches, respectively. T is a switching period. By specifying the maximum current ripple $r_i\%$ and differentiating ΔI to θ , we can solve for m_f :

$$m_f \geq \frac{50\pi}{I_s r_i \omega L} \left(\frac{V_s^2}{E} + \frac{V_s}{E} \sqrt{[(\omega L I_s)^2 + V_s^2] + 2E} \right) \quad (5)$$

where f_1 is the fundamental frequency. The deduction is as follows:

$$T = \frac{2\pi}{\omega_s} = \frac{1}{f_1 m_f} \quad \text{slope of } V_{tri} = \pm \frac{1}{T/4} = \pm \frac{4}{T}$$

So

$$\begin{aligned} T_{sON} &= 2 \left(\frac{V_{con}}{\text{slope}} + \frac{T}{4} \right) \\ &= 2 \left(\frac{m_a \sin(\theta - \phi)}{\text{slope}} + \frac{T}{4} \right) \\ &= \frac{T}{2} (1 + m_a \sin(\theta - \phi)) \end{aligned} \quad (6)$$

$$T_{sOFF} = T - T_{sON} = \frac{T}{2} (1 - m_a \sin(\theta - \phi)) \quad (7)$$

where $V_{con} = V_s \sin \theta - \omega L I_s \cos \theta = m_a \sin(\theta - \phi)$, $m_a = \sqrt{[V_s^2 + (\omega L I_s)^2]/E}$, and $\phi = \tan^{-1}(\omega L I_s/V_s)$. Substituting eqns. 6 and 7 into eqn. 4:

$$\begin{aligned} \Delta I &= \frac{1}{2} \left[\frac{V_s \sin \theta + E}{L} \frac{T}{2} (1 + m_a \sin(\theta - \phi)) \right. \\ &\quad \left. + \frac{E - V_s \sin \theta}{L} \frac{T}{2} (1 - m_a \sin(\theta - \phi)) \right] \\ &= \frac{T}{4L} [V_s m_a (\cos \phi - \cos(2\theta - \phi)) + 2E] \end{aligned} \quad (8)$$

By maximising ΔI wrt θ , a maximum current ripple can be obtained at $\theta = 180^\circ + \phi/2$. The minimum m_f thus can be calculated:

$$\Delta I_{max} = \frac{T}{4L} [V_s m_a (\cos \phi + 1) + 2E] \leq I_s r_i\%$$

$$\begin{aligned} V_s \frac{\sqrt{[V_s^2 + (\omega L I_s)^2]}}{E} \left(\frac{V_s}{\sqrt{[V_s^2 + (\omega L I_s)^2]} + 1} \right) \\ + 2E \leq \frac{I_s r_i f_1 m_f}{25} \end{aligned}$$

$$m_f \geq \frac{50\pi}{I_s r_i \omega L} \left(\frac{V_s^2}{E} + \frac{V_s}{E} \sqrt{[V_s^2 + (\omega L I_s)^2] + 2E} \right)$$

If m_f is a value smaller than 333, which means a switching frequency of 20 kHz, we can set m_f to be 333 or higher.

For the three-phase inverter, switching frequency is set to be the same as in SMR stage. An important factor to be considered is the amplitude modulation index m_a , which is the amplitude ratio of V_{con} to the V_{tri} . According

to analysis [12], m_a (less than 1 in linear modulation) is also the amplitude ratio of fundamental output component to AC source voltage. To prevent too short turn on or off times near the odd multiples of $\pi/2$ in V_{con} , m_a should not be too small. In this paper, m_a is set to be the value V_s/E for AC output to obtain the same amplitude with the source. The counting rate of the ROM based SPWM modulator should be fast enough to minimise the quantisation error. Its upper limit is the fetching speed of the ROM. It also should be noted that the required memory size is proportional to the counting rate of the oscillator. Three-phase SPWM control patterns can be programmed by delaying phase A signals $2\pi/3$ and $4\pi/3$, respectively, to get patterns of phase B and phase C.

Usually the DC link capacitors are determined by the specified voltage ripple. In the single-phase SMR, as shown in Fig. 2a, I_1 and I_2 are sinewaves modulated by SPWM pulses.

$$\begin{aligned} I_1 &= I_s \sin \omega t \\ &\times \frac{1}{2} [1 + m_a \sin(\omega t - \phi) + \text{high frequency terms}] \\ &= \frac{1}{4} I_s m_a \cos \phi + \frac{1}{2} I_s \sin \omega t \\ &\quad + \frac{1}{4} I_s m_a \cos(2\omega t - \phi) \\ &\quad + \text{high frequency terms} \end{aligned} \quad (9)$$

$$\begin{aligned} I_2 &= I_s \sin \omega t \\ &\times \frac{1}{2} [1 - m_a \sin(\omega t - \phi) + \text{high frequency terms}] \\ &= -\frac{1}{4} I_s m_a \cos \phi + \frac{1}{2} I_s \sin \omega t \\ &\quad - \frac{1}{4} I_s m_a \cos(2\omega t - \phi) \\ &\quad + \text{high frequency terms} \end{aligned} \quad (10)$$

In the above equations of I_1 and I_2 , DC components supply the load. The source current splits into two equal fundamental components. The second-order harmonic results in the voltage ripple superimposed on the DC link voltage. According to the V - I characteristics of the capacitors, the required capacitance for a specified voltage ripple can be determined:

$$\begin{aligned} \Delta V &= (\text{second order harmonic of } I_1) \left(\frac{1}{2\omega C} + \frac{1}{2\omega C} \right) \\ &= \frac{\frac{1}{4} I_s m_a}{\omega C} \quad C = \frac{I_s m_a}{4\omega \Delta V} \end{aligned} \quad (11)$$

Bulky DC link capacitors are needed to provide filtering and energy storage. The cost, weight and size are considerable, and dynamic response slows down. In the proposed line conditioner scheme, the effect of DC link voltage ripple is compensated in the programmed SPWM control patterns. Also by taking advantage of the nature of three-phase system in which balanced three-phase vectors sum up zero, the energy storage function of the capacitors becomes not so stringent. In this way, the voltage imbalance problem can be lessened.

5 Simulations and experiments

Simulations have been tested on a system with the following operating conditions:

Line-to-neutral source voltage $V_s = 110$ V (RMS), 60 Hz.

AC side impedance $\omega L = 5 \Omega$; input current ripple $r_i\% = 10\%$

Output power $P = 3$ kW; amplitude modulation index $m_a = 0.78$

Regulated DC link voltage = 400 V; voltage ripple = 3 V

Substituting these parameters into eqns. 5 and 11, we can obtain:

$m_f = 67$ or switching frequency = 4 kHz and DC link capacitance $C = 2200 \mu\text{F}$ (voltage rating > 400 V).

Shown in Fig. 11A are the three-phase input currents and the current in the source neutral line. By inspecting Fig. 11A we can see that three-phase currents are fundamental sinewaves superimposed with high order harmonics.

Fig. 11A Three-phase source and neutral currents

The neutral currents are compressed to a limited level and only with switching frequency ripple. The harmonics of the three-phase input currents are gathered around the multiple of switching frequency and the THD is 3.34% well within the standards. Compared to a typical nonlinear load composed of three single phase diode rectifiers, which bears significant harmonics in both source and neutral currents (principally third harmonic), much lower THD is obtained. Input current of the phase A diode bridge is shown in Fig. 11B and the neutral current of

Fig. 11B Phase-A load current

loads is shown in Fig. 11C. Here the load resistance is 20Ω per phase. By studying the harmonic contents in Figs. 11D and E, it can be found that the utility only supplies fundamental currents. Load peak current is about 2.5 times utility peak current. Also the level and harmonic contents of the utility neutral current has been

Fig. 11C Neutral currents of load

considerably reduced. The regulated DC link voltage (400 V) is illustrated in Fig. 11F. The small 60 Hz ripple is the result of periodically changing the PWM control patterns with fundamental frequency.

Fig. 11D Comparison of harmonic content of phase-A load current and phase-A source current

■ load
□ source

Fig. 11E Comparison of harmonic content of load neutral current and source neutral current

■ load
□ source

Fig. 11F Regulated DC-link voltage

If a switching frequency of 20 kHz ($m_f = 333$) is adopted and the AC side impedance is a more typical value of 1Ω , the THD of the source currents reduces further to 0.52%. The harmonics are pushed to be above audible noises. This is illustrated in Fig. 12A. DC link

Fig. 12A Source and neutral currents

$m_f = 333$

capacitors of smaller values reduce the size, weight and cost of the elements. In Fig. 12B, the capacitances are reduced to $560 \mu\text{F}$. Though the ripple amplitude increases, it is still a very small portion (about 3 V peak-to-peak) of the DC link voltages. We can see that the

voltages of the two capacitors are almost identical. Thus the voltage imbalance problem encountered in half

Fig. 12B DC link voltages of upper (V_1) and lower (V_2) capacitors

bridge schemes can be eliminated. Depicted in Fig. 13 is the phase A source current with 10% third harmonic distortion in source voltage. Even under the distorted source voltage, input current still remains nearly sinusoidal.

Fig. 13 Source current under distorted source voltage

As for the inverter, by programming the three-phase PWM control patterns with compensation on voltage ripple and dead times as discussed earlier, AC outputs without distortion can be achieved. Simulations have been performed on the system with 10% 120 Hz ripple superimposed on the DC link voltages and $m_a = 0.78$. Shown in Figs. 14A and B are the normalised illustrations of unfiltered phase A AC output and its spectrum. From Fig. 14B, a low-order (third) harmonic is present.

Fig. 14A Normalised AC output without compensation

This will make the output filtering difficult and impractical. As a comparison, the normalised phase A AC output and spectrum with compensation are depicted in Figs. 14C and D. The differences between Figs. 14A and Fig. 14C are not obvious since the compensation is revealed in the high frequency modulation function. However, in Fig. 14D the third-order harmonic is reduced greatly. By

filtering out the harmonics which are multiples of the switching frequency, qualified AC outputs can be supplied to the loads.

Fig. 14B Normalised spectrum of Fig. 14A

Fig. 14C Normalised AC output with compensation

Fig. 14D Normalised spectrum of Fig. 14C

Illustrated in Fig. 15 is the experimental SMR phase A source current under voltage mode control. The source voltage (50 V RMS, 60 Hz) is supplied by an AC power

Fig. 15 Phase-A voltage and current
20 V/div., 1 A/div.

supply. Input inductance and output capacitance are 2.6 mH and 1150 μ F, respectively. Switching frequency is 50 kHz and 667 Ω /phase loads are used. Output DC voltage is 400 V. Shown in Fig. 16 are the SPWM control

Fig. 16 Control signals for inverter switches with dead time compensation

signals fed to the inverter switches. (1) and (2) are for the lower and upper switches, respectively, when load current is positive. (3) and (4) are for the lower and upper switches, respectively, when load current is negative. We can see that (2) and (3) are the same as the original control signals, while (1) and (4) advance their turn offs and delay their turn ons to compensate the dead time degradation.

6 Conclusions

A new AC line conditioner has been presented in this paper for three-phase four wire applications. The aim of sinusoidal input currents and reduced level of neutral line current are achieved. Also a high quality AC output voltage is obtained regardless of DC link voltage ripples and dead times. The feasibility of voltage mode control for SMRs is proven. The degradation effects of nonlinear

or unmatched loads can be separated from the power distribution system. Simulations show satisfactory performances of the proposed line conditioner.

With the expeditious progresses in semiconductor, high power rating and fast speed power switches such as GTOs and IGBTs, are now available. The proposed AC line conditioner can be implemented by utilising these switches and applied to any configuration of loads connected to three-phase four wire distribution systems.

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