A Voltage-Mode Controlled High-Input-Power-Factor AC Line Conditioner with Minimized Output Voltage Harmonics

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Abstract — A new AC line conditioner is presented. This system consists of a high-input-power-factor half-bridge switch-mode rectifier, DC-link capacitors and a low-output-harmonic half-bridge inverter. The rectifier is operated under voltage-mode feedback. No current sensing is required. The inverter is controlled by an EPROM-based modulator. Optimized SPWM patterns compensating DC-link voltage ripple and dead time effects are programmed. Synchronizing the SMR and the inverter can eliminate the voltage imbalance between DC-link capacitors. The presented system is feasible for single- and three-phase applications.

INTRODUCTION

More and more modern industrial applications require stable single- or three-phase AC power sources while the utility systems are contaminated by increasingly proliferated power electronics equipment. The demands for AC line conditioners providing high-quality AC output power are rapidly growing. The line conditioners usually consist of AC/DC diode-bridge rectifiers, DC-link capacitors and DC/AC PWM inverters. Diode rectifiers usually produce serious line current distortion. Bulky filter capacitors are utilized to remove the low-order line-frequency voltage harmonics on inverter's output. System dynamic response is inevitably slow. Switch-mode rectifiers (SMR) have been proposed recently[1-3]. Their advantages include unity power factor, lower current harmonics, smaller size of capacitors and regulated DC-link voltage. For SMR control, current-mode feedback is more popular than conventional voltage-mode. However, the current-mode technique has the following disadvantages[4]:

- 1) Current sensing adds complexity and cost to circuit implementation.
- 2) Slope compensation is necessary for peak inductor current control and leads to decreased current control loop gain.
- 3) Varying switching frequency in hysteresis control causes serious EMI problems and dead zones in the zero crossings of the current waveforms.

In this paper, a new AC line conditioner is presented as shown in Fig. 1. The three-phase line conditioner for high-power applications is built by simply paralleling

single-phase systems. This line conditioner consists of a high-input-power-factor half-bridge switch-mode rectifier, DC-link capacitors and a low-output-harmonic half-bridge inverter. The rectifier is operated under voltage-mode feedback[5]. There are only voltage control loops for input power factor correction and DC-link voltage regulation. No current sensing is required. The inverter is controlled by an fully-digitized EPROM-based PWM modulator. Optimized SPWM patterns compensating DC-link voltage ripple and dead time effects[6,7] are programmed. This approach featuring low cost, simple implementation and control is especially suitable in fixed-frequency applications. Synchronizing the SMR and the inverter can eliminate the voltage imbalance between DC-link capacitors.

Fig.1 Half-bridge AC line conditioner (a) 1φ (b) 3φ

VOLTAGE-MODE CONTROL OF HALF-BRIDGE SMR

A general circuit for 1φ SMR is proposed in [3], as shown in Fig. 2. v_u is the pulse voltage which could be E or $-E$ depending on the switching states. E is the dc voltage of the output capacitor and $v_{u,1}$ is the fundamental component of v_u .

Fig. 2 General circuit and phasor diagram for 1φ SMR

From the phasor diagram, the principles of the voltage-mode control are deduced as follows :

Let $v_s(t) = \sqrt{2} V_s \sin \omega t$, $i_s(t) = \sqrt{2} I_s \sin \omega t$ under steady state, $v_L(t) = L \cdot di_s(t)/dt = \sqrt{2} \omega L I_s \cos \omega t$, then $v_{u,1}(t) = \sqrt{2} (V_s \sin \omega t - \omega L I_s \cos \omega t)$. From the transfer characteristics of the SPWM inverter and the theorem of superposition, to produce an ac voltage $v_{u,1}$, the reference signal or the control voltage has to be :

$$v_{\text{control}} = v_{u,1} / E = \frac{\sqrt{2} V_s}{E} \sin \omega t - \frac{\sqrt{2} \omega L I_s}{E} \cos \omega t \\ = k_1 \sin \omega t - k_2 \cos \omega t$$

That is, the control voltage can be generated by summing an in-phase component and an in-quadrant component. k_1 is the parameter concerning the feedforward input voltage v_s and k_2 concerning the feedback output voltage E .

The discussions above are based on the ideal sinusoidal source voltage. If the circumstances are under a distorted source voltage, the deduction is the same except for replacement of the pure sine-term. That is, $v_{\text{control}} = k_1' v_s(t) - k_2 \cos \omega t$. Since the information about the source voltage is revealed in the first term of v_{control} , a sinusoidal input current can be obtained even with the distorted source present. The block diagram depicting the voltage-mode control for SMR is shown in Fig. 3.

Fig. 3 Block diagram of voltage-mode control

SELECTIONS OF L AND C

The input inductor can be designed by analyzing Fig. 4(a). Because of the sinusoidal variation of the input current, its switching ripple varies period by period. The current ripple can be defined as the average current variations in a switching period : $\Delta I = [(di/dt)_{\text{inc}} \cdot T_{\text{son}} + (di/dt)_{\text{dec}} \cdot T_{\text{soff}}] / 2$. $(di/dt)_{\text{inc}}$ and $(di/dt)_{\text{dec}}$ are the increasing and decreasing rates of input current. T_{son} and T_{soff} are the on-time and off-time of the switch, respectively. By specifying the

maximum current ripple $r_i\%$ and differentiating ΔI to θ , we can solve for ωL (see appendix 1) :

$$\omega L = \frac{A \cdot B + \sqrt{(AB)^2 - [A^2 - (2P/V_s)^2](B^2 - V_s^2)}}{A^2 - (2P/V_s)^2},$$

P is output power, $A = (P \cdot r_i \cdot m_f \cdot E) / (25 \cdot \pi \cdot V_s^2)$,

$$B = V_s + 2 \cdot E^2 / V_s, \quad mf = \frac{\text{switching frequency}}{\text{fundamental frequency}}$$

The design of output capacitor is according to the output switching current, as shown in Fig. 4(b). Assuming that a dc current is fed to the load, then I_o' contains only harmonics of I_o . It is known that the 2nd-order line-frequency harmonic, $I_{o,2}$, is the dominant one. The capacitance can then be determined as : $\Delta V_o = I_{o,2} / \omega C = V_o \cdot r_v\%$,

$$\omega C = 100 \cdot I_{o,2} / (V_o \cdot r_v).$$

V_o is the output voltage, which is $2 \cdot E$ in the half-bridge topology.

Fig. 4 Selections of (a) input inductor and (b) output capacitor

COMPENSATION TECHNIQUES FOR INVERTERS

With low-order line-frequency ripples on the DC-link voltage, typically-utilized sinusoidal PWM (SPWM) techniques will produce unwanted low-frequency harmonics in the inverter's output. Assuming that the DC-link voltage, V_{dc} , is ripple-free. Then the output of the inverter is an AC voltage modulated by a modulation function with frequency spectrum as depicted in Fig. 5. Harmonics are centered around the multiples of the switching frequency. This AC output can be written as :

$$V_{ac} = V_{dc} \cdot [m_a \sin \omega t + \sum_{n+k=\text{odd}} l(m_f \cdot n \pm k) \sin((m_f \cdot n \pm k)\omega t)],$$

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

Fig. 5 Spectrum of a SPWM modulation function

Let's now consider the case in which low-order harmonic $k \cdot h(\omega t)$ is superposed on the DC-link voltage, $V_{dc}' = V_{dc} \cdot (1 + k \cdot h(\omega t))$. The modulated AC output voltage then becomes :

$$V_{ac}' = V_{dc} (1 + k \cdot h(\omega t)) [m_a \sin \omega t + \sum_{n+k=\text{odd}} (m_f \cdot n \pm k) \sin((m_f \cdot n \pm k) \omega t)]$$

$$= V_{ac} + V_{dc} \cdot k \cdot h(\omega t) \cdot m_a \sin \omega t + \dots$$

Low-order output harmonics are produced due to the multiplication of two low-order terms, $k \cdot h(\omega t)$ and $m_a \sin \omega t$. By modifying the original reference function to $\frac{m_a \sin \omega t}{1 + k \cdot h(\omega t)}$, the effects of low-order ripples 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 the "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. A load-current-dependent compensation method[7] can eliminate the output voltage harmonic distortion due to dead time effects. The principles of the compensation techniques are summarized 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. This can be achieved by turning off the lower switch in advance and turning it on after a delay time. 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 with the original SPWM control patterns. An example of switching waveforms for dead time compensation is shown in Fig. 6.

Fig. 6 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 = resulted AC output voltages

The above method is impossible to implement on the exiting system because it requires primary generation of control patterns for switches in advance. Here a scheme adopting reference sinewaves with offsets is proposed. As shown in Fig. 7, while comparison between V_{tri} and V_{ro} produces the original SPWM signals, comparing V_{tri} with V_{r+} and V_{r-} can give control signals with dead time compensation. The added dc-offsets can be related to the minimum required dead time, d_{min} , as :

$|\Delta V| = V_{tri} \cdot d_{min} (4f_s + \omega_0 m_a)$, where f_s is the switching frequency and m_a is the amplitude modulation index (see appendix 2).

Fig. 7 SPWM with offsetted reference signals

EPROM-BASED SPWM MODULATOR

Conventionally, control signals fed to the PWM inverters are produced by electronic hardware or microprocessor. The cost and complexity of hardware implementation are considerable. On-line computation in the microprocessor approach is laborious and time-consuming. These two approaches seem impractical when applied to the proposed compensation techniques. An EPROM-based SPWM modulator[8] can solve the problems mentioned above. The SPWM control patterns are pre-calculated by computing the intersections of a reference wave and a triangular carrier wave with numerical method. To program the analogue SPWM patterns into the EPROM, each pulse is composed of several consecutive digital '1's to represent its width. The duration between pulses are represented by consecutive '0's. This digitized SPWM pattern can then be retrieved sequentially by incrementing the address counter with a clock. This modulator features the advantages of low cost, fast response, simple control and implementation.

In the proposed EPROM-based modulator for fixed-frequency inverters, an optimized SPWM switching function compensating DC-link voltage ripple and dead time effects can be off-line calculated and pre-programmed. A fixed-frequency oscillator with counters retrieves the PWM patterns by counting through the addresses of the EPROM. A reset signal is sent at the end of a programmed period from the EPROM to ensure a complete cycle. The block diagram of the proposed modulator is illustrated in Fig. 8. The fixed-frequency oscillator can be replaced by a tracking PLL circuit when the line frequency varies.

Fig.8 Block diagram of the proposed high-performance EPROM-based SPWM modulator

RESULTS AND DISCUSSIONS

Simulations have been performed on the 1 ϕ AC line conditioner with the following operating conditions :

- Source voltage $V_s = 110\text{V(rms)}$, 60Hz.
- AC-side impedance $\omega L = 5\Omega$
- DC-link capacitance $\omega C = 0.8 \text{ 1}/\Omega$
- Output power $P = 1\text{kW}$
- Regulated DC-link voltage = 400V
- Switching frequency = 2kHz
- Amplitude modulation index $m_a = 0.78$

Fig. 9(a) shows the input current of 1 ϕ SMR. Fig. 9(b) is the output voltage and Fig. 9(c) is the spectrum of input current (THD=8.32%). The switching frequency is set to 2kHz for visual clarity. If a realizable higher switching frequency, say, 20kHz, is adopted, the THD is 1.65% which is well within the maximum allowable limit of 8% in IEEE-519.

Fig. 9(a) Input current (b) Output voltage (c) Spectrum of (a)

Fig. 10 illustrates the ripple compensation of a modified modulation function, where 120Hz ripple amplitude is 20%. Fig. 10(a) is the normalized inverter's output voltage and (b) is the spectrum without compensation. Fig. 10(c) and (d) show those with compensation. The differences between Fig. 10(a) and Fig. 10(c) are not obvious since the compensation is revealed in the high-frequency modulation function. However, in Fig. 10(d) the 3rd harmonic is reduced greatly compared to that in Fig. 10(b). By filtering out the harmonics which are multiples of the switching frequency, qualified AC outputs can be supplied to the loads.

Fig. 10(a) Normalized output voltage without compensation
(b) Spectrum of (a)
(c) Normalized output voltage with compensation
(d) Spectrum of (c)

Fig. 11 shows the input currents of a 3 ϕ SMR.

Fig. 11 Input currents of 3 ϕ SMR

In the laboratory experiments, a 1 ϕ half-bridge SMR is built and the results are recorded. Illustrated in Fig. 12 is the input current and input voltage under voltage-mode control. The source voltage (50Vrms, 60Hz) is supplied by an AC power supply. Input inductance and output capacitance are 2.6mH and 1150 μ F, respectively. Switching frequency is 50kHz and a 667 Ω resistive load is used. Output DC voltage is 200V.

Fig. 12 Input voltage and input current (20V/div., 1A/div.)

Shown in Fig. 13(a) is the SPWM control signals fed to switches when the load current is positive. The upper part is the same as the original pattern. The lower part advances turning-offs and delays turning-ons to implement dead-time compensation. Fig. 13(b) depicts the SPWM control signals when load current is negative. The upper part advances turning-offs and delays turning-ons and the lower part is the original pattern.

In the half-bridge SMR, C1 charges and C2 discharges in the positive half-period, and vice versa. This causes the voltage imbalance between C1 and C2. In the proposed AC line conditioner, SMR charges/discharges and inverter discharges/charges the same capacitor. In this way, the voltage imbalance problem can be solved by synchronizing the SMR and inverter stages.

(a)

(b)

Fig. 13 SPWM control signals for inverter switches when (a) load current > 0 (b) load current < 0

CONCLUSION

A novel control scheme for a new AC line conditioner is presented. In this scheme, a half-bridge SMR with voltage-mode control serves as the AC/DC converter. No inner current loop is required. Based on the principles of the well-developed SPWM, relative circuit simplicity, low cost and regulated DC-link voltage can be achieved. The DC/AC inverter is also a half-bridge type. A fully-digitized SPWM modulator with compensation of DC-link voltage ripple and dead time effects is utilized. Synchronizing the SMR and the inverter can eliminate the voltage imbalance between DC-link capacitors. High power capability or

multi-phase applications can be implemented by simply paralleling 1 ϕ modules.

The applications of interfacing regulated DC or qualified AC sources with utility systems proliferate extensively. The proposed scheme provides a potentially effective and economic way to improve the performance of line conditioners.

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APPENDIX 1

From Fig. 4(a), the maximum current ripple under voltage-mode control can be calculated as follows :

$$T = \frac{2\pi}{\omega_s} = \frac{1}{f_s \cdot m_f}; \text{ slope of } V_{ri} = \pm \frac{1}{T/4} = \pm \frac{4}{T};$$

$$\text{So } T_{sON} = 2 \cdot \left(\frac{V_{con}}{\text{slope}} + \frac{T}{4} \right) = 2 \cdot \left(\frac{m_a \cdot \sin(\theta - \phi)}{\text{slope}} + \frac{T}{4} \right) \\ = \frac{T}{2} (1 + m_a \cdot \sin(\theta - \phi));$$

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

$$\text{where } V_{con} = V_s \sin \theta - \omega L I_s \cos \theta = m_a \cdot \sin(\theta - \phi), \\ m_a = \sqrt{V_s^2 + (\omega L I_s)^2} / E, \text{ and } \phi = \tan^{-1} \frac{\omega L I_s}{V_s}$$

According to the definition of current ripples :

$$\Delta I = [(dI/dt)_{inc} \cdot T_{sON} + (dI/dt)_{dec} \cdot T_{sOFF}] / 2 \\ = \frac{1}{2} \left[\frac{V_s \sin \theta + E}{L} \cdot \frac{T}{2} (1 + m_a \cdot \sin(\theta - \phi)) + \right. \\ \left. \frac{E - V_s \sin \theta}{L} \cdot \frac{T}{2} (1 - m_a \cdot \sin(\theta - \phi)) \right] \\ = \frac{T}{2L} [V_s \cdot m_a \cdot \sin \theta \sin(\theta - \phi) + E] \\ = \frac{T}{4L} [V_s \cdot m_a \cdot (\cos \phi - \cos(2\theta - \phi)) + 2E]$$

By maximizing ΔI with respect to θ , a maximum current ripple can be obtained at $\theta = \frac{180^\circ + \phi}{2}$. The minimum ωL thus can be calculated :

$$\Delta I_{max} = \frac{T}{4L} [V_s \cdot m_a \cdot (\cos \phi + 1) + 2E] \leq I_s \cdot r \% \\ V_s \cdot \frac{\sqrt{V_s^2 + (\omega L I_s)^2}}{E} \cdot \left(\frac{V_s}{\sqrt{V_s^2 + (\omega L I_s)^2}} + 1 \right) + 2E \leq \frac{I_s \cdot r \cdot \omega L \cdot m_f}{50\pi} \\ \sqrt{(\omega L I_s)^2 + V_s^2} \leq \frac{I_s \cdot r \cdot m_f \cdot E}{50\pi \cdot V_s} \omega L - (V_s + 2E^2 / V_s) \\ \leq A \cdot \omega L - B \\ (A^2 - I_s^2)(\omega L)^2 - 2BA \cdot \omega L + B^2 - V_s^2 \geq 0 \\ \text{Let } I_s = 2P/V_s, \text{ then} \\ \omega L = \frac{A \cdot B + \sqrt{(AB)^2 - [A^2 - (2P/V_s)^2](B^2 - V_s^2)}}{A^2 - (2P/V_s)^2}$$

APPENDIX 2

According to Fig. 7, the minimum offset voltage required for dead time compensation can be deduced as :

The slope of triangular wave is $2V_{ri}/(T/2) = 4f_s \cdot V_{ri}$.

At some t_0 , the slope of V_{ri} is $\omega_0 \cdot m_a \cdot V_{ri} \cdot \cos(\omega_0 t_0)$.

The intersection can then be calculated by solving the following two equations :

$$y = 4f_s \cdot V_{ri}(t - t_0) \\ y - m_a \cdot V_{ri} \cdot \sin(\omega_0 t_0) = \omega_0 \cdot m_a \cdot V_{ri} \cdot \cos(\omega_0 t_0)(t - t_0)$$

$$\Rightarrow \Delta t = t - t_0 = \frac{m_a \cdot \sin \omega_0 t_0}{4f_s - \omega_0 \cdot m_a \cdot \cos(\omega_0 t_0)}$$

If a dc offset ΔV is added in the reference wave, the resulted drift in t -axis is :

$$d = \frac{\Delta V / V_{tri}}{4f_s - \omega_0 \cdot m_a \cdot \cos(\omega_0 t_0)}$$

For a specified minimum dead time d_{min} , the required ΔV is :

$$\Delta V = V_{ri} \cdot d_{min} (4f_s - \omega_0 \cdot m_a \cdot \cos(\omega_0 t_0))_{min} \\ = V_{ri} \cdot d_{min} (4f_s + \omega_0 \cdot m_a)$$