Letters to the Editor

A New Circuit for Measuring Power Factor in Nonsinusoidal Load Current

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Abstract—This letter presents a new method for measuring the power factor in nonsinusoidal load current. The proposed method is based on the technique of sampling and holding the amplitude of the real part of the load current. Because no Fourier or fast Fourier transform calculation is factor in nonsinusoidal load current. The proposed method is based on the

I. INTRODUCTION

Power factor (PF) is an important index to evaluate the power efficiency of electric equipment or electric utility. The design of conventional PF measurement equipment is based on the assumption that the measured voltage and current signals are sinusoidal. Hence, the measured results of conventional PF measurement methods are not accurate under the distorted load current [1]–[4]. Unfortunately, nonlinear load, where current waveform is nonsinusoidal, is widespread due to the wide use of power-electronic-related equipment [5]–[7]. The commercial PF measurement equipment, which can measure the PF under the nonsinusoidal load current, is using the fast Fourier transform (FFT) analysis in the calculation process to calculate the real power, root mean value of voltage, and root mean value of current [8]. An electronic circuit for measuring the PF under the nonsinusoidal load current was propose in [9]. However, three power components \( I_1 \), \( \theta_1 \), and \( I_{\text{rms}} \) must be measured individually in that paper. In this letter, a new PF measurement method, based on the sample-and-hold technique, is proposed. The amplitude of the real part of load current can be measured directly. Hence, the proposed method is simpler.

II. POWER FACTOR DEFINITION

The voltage supplied from the power system can be assumed to be a sinewave. It is represented as

\[
v(t) = \sqrt{2} V_{\text{rms}} \sin(\omega t).\]

(1)

If the nonlinear load current contains no dc component in the steady state, it can be represented as [10]

\[
i_L(t) = i_{L,1}(t) + \sum_{n=2}^{\infty} i_{L,n}(t)
\]

(2)

where \( i_{L,1}(t) \) is the fundamental component and \( i_{L,n}(t) \) is the component at the \( n \)-th harmonic frequency. The load current components in (2) can also be expressed as

\[
i_{L}(t) = \sqrt{2} I_1 \sin(\omega t - \theta_1) + \sqrt{2} \sum_{n=2}^{\infty} I_n \sin(n \omega t - \theta_n)
\]

(3)

where \( I_n \) and \( \theta_n \) are the rms value and phase of the load current at the \( n \)-th order harmonic. The average power \( P \) is defined as

\[
P = \frac{1}{T} \int_0^T p(t)dt = \frac{1}{T} \int_0^T v(t)i_L(t)dt.
\]

(4)

Using \( v(t) \) from (1) and \( i_L(t) \) from (3), then

\[
P = \frac{1}{T} \int_0^T \left(\sqrt{2} I_1 \sin(\omega t - \theta_1) + \sqrt{2} \sum_{n=2}^{\infty} I_n \sin(n \omega t - \theta_n)\right) dt
\]

\[
= \sqrt{2} I_1 V_{\text{rms}} \cos(\theta_1).
\]

(5)

The apparent power \( S \) is the product of the rms voltage \( V_{\text{rms}} \) and the rms current \( I_{\text{rms}} \). It is represented as

\[
S = V_{\text{rms}} I_{\text{rms}}.
\]

(6)

The PF is defined as

\[
\text{PF} = \frac{P}{S}.
\]

(7)

Using (5)–(7), the PF can be derived as [10]

\[
\text{PF} = \frac{P}{S} = \frac{V_{\text{rms}} I_1 \cos(\theta_1)}{V_{\text{rms}} I_{\text{rms}}} = \frac{I_1}{I_{\text{rms}}} \cos(\theta_1)
\]

(8)

where \( \cos(\theta_1) \) is defined as the displacement PF (DPF). DPF can be represented as

\[
\text{DPF} = \cos(\theta_1).
\]

(9)

Therefore, the PF in nonsinusoidal current is

\[
\text{PF} = \frac{I_1}{I_{\text{rms}}} \text{DPF}.
\]

(10)

The PF definition shown in (10) [10] is also named the “true PF” [11].

If the load is linear, the load current is a pure sinusoidal waveform. It is represented as

\[
i_L(t) = \sqrt{2} I_1 \sin(\omega t - \theta_1).
\]

(11)

In this condition, \( I_1 \) is equal to \( I_{\text{rms}} \). Then, the PF is equal to the DPF. It is written as

\[
\text{PF} = \cos(\theta_1) = \text{DPF}.
\]

(12)

Nevertheless, the PF definition written as (12) is accurate only under the condition that the waveform of load current is sinusoidal.
III. PROPOSED METHOD

A. PF Measurement for Sinusoidal Current

If the load is linear, the load current is a pure sinusoidal signal and $I_1$ is equal to $I_{\text{rms}}$. Then, the PF $\cos(\theta_1)$ is equal to the DPF. A pure sinusoidal reference signal with the phase of 90° lagging with the load current signal can be written as

$$v_{r1}(t) = \sqrt{2} I_1 \cos(wt - \theta_1).$$

The absolute value of $v_{r1}(t)$ can be written as

$$|v_{r1}(t)| = \sqrt{2} I_1 |\cos(wt - \theta_1)|.$$

The phase relationships between the above signals are shown in Fig. 1(a). The points $P_1$ and $P_2$ are at the time of zero-crossing point of the input voltage. The value of $v_{r2}(t)$ at $P_1$ or $P_2$ is just $\sqrt{2} I_1 \cos(\theta_1)$. By multiplying a factor $1/\sqrt{2}$, $I_1 \cos(\theta_1)$ is obtained. If the load current is sinusoidal, $I_1$ is equal to $I_{\text{rms}}$. PF can be obtained by dividing $I_1 \cos(\theta_1)$ to $I_{\text{rms}}$.

B. PF Measurement for Nonsinusoidal Current

In a practical industrial power system, the load is frequently nonlinear. In this condition, the conventional PF definition is not suitable. For measuring the true PF, the measurement method must be modified.
Fig. 2. Test results of the proposed detector. (a) Input voltage $V_i(t)$. (b) Load current $i_l(t)$. (c) Absolute value of reference signal $V_r(t)$. (d) Detected output voltage $V_o(t)$ (channel (a): 5 V/div, 5 ms/div; channel (b): 10 V/div, 5 ms/div; channel (c): 5 V/div, 5 ms/div; channel (d): 0.5 V/div, 5 ms/div).

To obtain the fundamental component of load current, the load current is fed to a bandpass filter to eliminate the harmonic component. Then, the fundamental component of load current, $\sqrt{2} I_1 \sin(\omega t - \theta_1)$, is gotten. Since the fundamental component of load current is obtained by using the bandpass filter, the following measurement process is similar to the condition of Section III-A.

IV. HARDWARE IMPLEMENTATION

Fig. 1(b) and (c) shows the block and schematic diagrams for implementing this method. The voltage signal is fed to a precision full-wave rectifier and a trigger pulse generator to generate a pulse every half cycle.

The load current shown as (2) is divided into two paths. One is fed to an rms measurement circuit for measuring the rms value of the load current. The other is fed to a bandpass filter to extract its fundamental component. The output of the bandpass filter is fed to a 90° phase-shift circuit to generate the signal shown as (13). The output signal of the phase-shift circuit is fed to a precision full-wave rectifier. Then, the signal shown as (14) is obtained. The generated pulse and the signal shown as (14) are fed to a sample-and-hold circuit, then, $\sqrt{2} I_1 \cos(\theta_1)$ is obtained. For getting $I_1 \cos(\theta_1)$, the output of the sample-and-hold circuit is multiplying a factor $1/\sqrt{2}$. The rms of load current, $I_{rms}$, and $I_1 \cos(\theta_1)$ are fed to an analog divider to operate as a denominator and numerator, respectively. Then, the true PF shown as (10) is obtained. The value of PF is in the range from 0 to 1 in practice. For representing the PF as a dc voltage, the correction factor in this letter is designed as 1. For example, if the dc output voltage is 1 V, it represents that PF is unity.

V. TEST RESULTS

The computer-based signal generator is used in the following test to generate the input voltage and the load current. The phase difference between two signals is defined as the same frequency in the normal condition. However, the frequency of the two input signals in Fig. 2 is different to intensify the transient performance of the proposed method clearly [9]. Fig. 2 shows the test results of this method under the pure sinewave. It shows that the proposed method can measure the PF as a dc voltage.

To verify the performance of the proposed circuit under the nonsinusoidal load current, the nonlinear load current which contains a fixed fundamental component and the varied third-order harmonic is used. The nonsinusoidal load current used in the following test can be represented as

$$i_L(t) = \sin(\omega t) + A \sin(3\omega t + \phi)$$  \hspace{1cm} (15)

where $A$ is the amplitude of harmonic and $\phi$ is the phase of the
A Reduced Hysteresis Controller for a Four-Switch Three-Phase Bidirectional Power Electronics Interface

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Abstract—A reduced hysteresis controller for a four-switch three-phase bidirectional power electronics interface is proposed in this letter. Polarity detection of the input voltages to determine the switching states is not required. The circuit is simple. Experimental results show unity input power factor operation and bidirectional power transfer capability.

Index Terms—Hysteresis controller, three-phase bidirectional interface, unity power factor.

I. INTRODUCTION

Many types of modern power electronics equipment utilize ac/dc rectifiers as front stages. Recently, switch-mode rectifiers have been proposed as substitutes for diode rectifiers for lower current harmonics contents and high power factors in single- or three-phase applications [1]–[4]. One novel three-phase bidirectional power electronics interface was presented in [4], which claims a high output voltage and the use of four active switches only. The hysteresis current control scheme was adopted for the novel three-phase topology. Besides two hysteresis comparators, there are, in addition, four comparators for detecting the polarities of the phase voltages. In this letter, a reduced hysteresis controller is proposed to reduce the cost of the circuit. Only two hysteresis comparators are required. Experimental results demonstrate the unity input power factor and bidirectional capabilities of the circuit.

II. CIRCUIT DESCRIPTION

A novel four-switch three-phase bidirectional interface is shown in Fig. 1. The phase-C source voltage is directly connected to the midpoint of two output capacitors through the filter inductor. In a balanced three-phase system, the current sum is zero. Thus, by controlling the input currents of phases A and B with the four switches, the phase-C current can be forced to remain within a hysteresis band. Under steady state, it can be deduced that

$$\Delta i_k = \frac{v_s - V_s'}{L} \Delta T \sum_{k=a,b,c} \Delta i_k = 0, \quad k = a, b, c$$

where \(v_s\) is the source voltage, \(V_s'\) is the voltage at the other side of the inductor, \(i_k\) is the phase current, \(\Delta T\) is the duration of a switching state, and \(\Delta i_k\) is the variation of each phase current in \(\Delta T\). According to the polarities of \(v_s\) and \(v_s'\), one can divide a \(2\pi\) period into four sections. There are four possible conducting paths in each section with different state combinations of the switches. \(V_s'\) and \(\Delta i_k\) can then be determined, as summarized in Table I. Whether a switch or its paralleled reverse diode is conducting depends on the direction of the phase current and, therefore, the polarity of the source voltage under unity power factor operation. In conducting path 1, \(S_1\) (or \(S_2\)) and \(S_3\) (or \(S_4\)) are carrying the currents. From (1), \(i_a\) increases, thus, \(i_3\) must decrease. The maximum ripple of \(i_3\) could be

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