17

Power Quality and Utility Interface Issues

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17.1 Overview

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In the traditional sense, when one thinks of power quality, visions of classical waveforms containing 3rd, 5th, and 7th, etc. harmonics appear. It is from this perspective that the IEEE in cooperation with utilities, industry, and academia began to attack the early problems deemed "power quality" in the 1980s. However, in the last 5 to 10 years, the term *power quality* has come to mean much more than simple power system harmonics. Because of the prolific growth of industries that operate with sensitive electronic equipment (e.g., the semiconductor industry), the term power quality has come to encompass a whole realm of anomalies that occur on a power system. Without much effort, one can find a working group or a standards committee at various IEEE meetings in which there is much lively debate on the issue of what power quality entails and how to define various power quality events. The IEEE Emerald Book (IEEE Standard 1100-1999, *IEEE Recommended Practice for Power and Grounding Electronic Equipment*) defines power quality as:

> The concept of powering and grounding electronic equipment in a manner that is suitable to the operation of that equipment and compatible with the premise wiring system and other connected equipment.

The increased use of power electronic devices at all levels of energy consumption has forced the issue of power quality out of abstract discussions of definitions down to the user level. For example, a voltage sag of only a few cycles could cause the variable frequency drives or programmable logic controllers (PLC) on a large rolling mill to "trip out" on low voltage, thereby causing lost production and costing the company money. The question that is of concern to everyone is, "What initiated the voltage sag?" It could have been a fault internal to the facility or one external to the facility (either within a neighboring facility or on the utility system). In the latter two cases, the owner of the rolling mill will place blame on the utility and, in some cases, seek recompense for lost product.

Another example is the semiconductor industry. This segment of industry has been increasingly active in the investigation of power quality issues, for it is in this volatile industry that millions of dollars can possibly be lost due to a simple voltage fluctuation that may only last two to four cycles. Issues of power quality have become such a concern in this industry that Semiconductor Equipment and Materials International (SEMI), the worldwide trade association for the semiconductor industry, has been working to produce power quality standards explicitly related to the manufacturing of equipment used within that industry.

Two excellent references on the definitions, causes, and potential corrections of power quality issues are Refs. 1 and 2. The purpose of this section, however, is to address some of the main concerns regarding the power quality related to interfacing with a utility. These issues are most directly addressed in IEEE 519-1992 [3] and IEEE/ANSI C62.41 [4].

Harmonics and IEEE 519

Harmonic generation is attributed to the application of nonlinear loads (i.e., loads that when supplied a sinusoidal voltage do not draw a sinusoidal current). These nonlinear loads not only have the potential to create problems within the facility that contains the nonlinear loads but also can (depending on the stiffness of the utility system supplying energy to the facility) adversely affect neighboring facilities. IEEE 519-1992 [3] specifically addresses the issues of steady-state limits on harmonics as seen at the *point of common coupling* (PCC). It should be noted that this standard is currently under revision and more information on available drafts can be found at http://standards.ieee.org.

The whole of IEEE 519 can essentially be summarized in several of its own tables. Namely, Tables 10.3 through 10.5 in Ref. 3 summarize the allowable harmonic current distortion for systems of 120 V to 69 kV, 69.001 kV to 161 kV, and greater than 161 kV, respectively. The allowable current distortion (defined in terms of the total harmonic distortion, THD) is a function of the stiffness of the system at the PCC, where the stiffness of the system at the PCC is defined by the ratio of the maximum short-circuit current at the PCC to the maximum demand load current (at fundamental frequency) at the PCC. Table 11.1 in Ref. 3 provides recommended harmonic voltage limits (again in terms of THD). Tables 10.3 and 11.1 are of primary interest to most facilities in the application of IEEE 519. The total harmonic distortion (for either voltage or current) is defined as the ratio of the rms of the harmonic content to the rms value of the fundamental quantity expressed in percent of the fundamental quantity. In general, IEEE 519 refers to this as the distortion factor (DF) and calculates it as the ratio of the square root of the sum of the squares of the rms amplitudes of all harmonics divided by the rms amplitude of the fundamental all times 100%. The PCC is, essentially, the point at which the utility ceases ownership of the equipment and the facility begins electrical maintenance (e.g., the secondary of a service entrance transformer for a small industrial customer or the meter base for a residential customer).

Surge Voltages and C62.41

Reference 4 is a *guide* (lowest level of standard) for characterizing the ability of equipment on low-voltage systems (<1000 V) to withstand voltage surges. This guide provides some practical basis for selecting appropriate test waveforms on equipment. The primary application is for residential, commercial, and industrial systems that are subject to lightning strikes because of their close proximity (electrically speaking) to unshielded overhead distribution lines. Certain network switching operations may also result in similar voltage transients being experienced.

Other Standards Addressing Utility Interface Issues

Many power quality standards are at present in existence and are under constant revision. The following standards either directly or indirectly address issues with the utility interface and can be applied accordingly: IEEE 1159 for the monitoring of power quality events, IEEE 1159.3 for the exchange of measured power quality data, IEEE P1433 for power quality definitions, IEEE P1531 for guidelines regarding harmonic filter design, IEEE P1564 for the development of sag indices, IEEE 493 (the *Gold Book*) for industrial and commercial power system reliability, IEEE 1346 for guidelines in evaluating component compatibility with power systems (this guideline is an attempt to better quantify the CBEMA and ITIC curves), C84.1 for voltage ratings of power systems and equipment, IEEE 446 (the *Orange Book*) for emergency and standby power systems (this standard contains the so-called power acceptability curves), IEEE 1100 (the *Emerald Book*), IEEE 1409 for development of guidelines for the application of power electronic devices/technologies for power quality improvement on distribution systems, and IEEE P1547 for the power quality issues associated with distributed generation resources.

As previously mentioned, the IEEE is not the only organization to continue investigation into the impacts of nonlinear loads on the utility system. Other organizations such as CIGRE, UL, NEMA, SEMI, IEC, and others all play a role in these investigations.

References

1. IEEE Standards Board, *IEEE Recommended Practice for Powering and Grounding Electronic Equipment,* IEEE Std. 1100-1999.
2. R. C. Dugan, M. F. McGranaghan, and H. W. Beaty, *Electrical Power Systems Quality,* McGraw-Hill, New York, 1996.
3. IEEE Standards Board, *IEEE Recommended Practices and Requirements for Harmonic Control in Electrical Power Systems,* IEEE Std. 519-1992.
4. IEEE Standards Board, *IEEE Recommended Practice on Surge Voltages in Low-Voltage AC Power Circuits,* IEEE Std. C62.41-1991.

17.2 Power Quality Considerations

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Harmonics

In the past, utilities had the responsibility to provide a single-frequency voltage waveform, and for the most part, customers' loads had little effect on the voltage waveform. Now, however, power electronics are used widely and create nonsinusoidal currents that contain many harmonic components. Harmonic currents cause problems in the power system and for other loads connected to the same portion of the power system. Because utility customers can now cause electrical problems for themselves and others, the Institute of Electrical and Electronic Engineers (IEEE) developed IEEE Standard 519, which places the responsibility of controlling harmonics on the user as well as the utility. This section describes harmonics, their cause, and their effects on the system voltage and components.

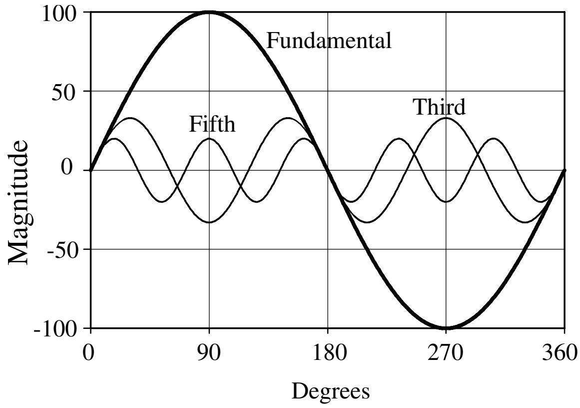

FIGURE 17.1 Fundamental, third, and fifth harmonics.

What Are Harmonics?

Ideally, the waveforms of all the voltages and currents in the power system would be single-frequency (60 Hz in North America) sine waves. The actual voltages and currents in the power system, however, are not purely sinusoidal, although in the steady state they do look the same from cycle to cycle; i.e., $f(t+T) = f(t)$, where T is the period of the waveform and t is any value of time. Such repeating functions can be viewed as a series of components, called harmonics, whose frequencies are integral multiples of the power system frequency. The second harmonic for a 60-Hz system is 120 Hz, the third harmonic is 180 Hz, etc. Typically, only odd harmonics are present in the power system.

Figure 17.1 shows one cycle of a sinusoid (labeled as the fundamental) with a peak value of 100. The fundamental is also know as the first harmonic, which would be the nominal frequency of the power system. Two other waveforms are shown on the figure—the third harmonic with a peak of 50 and the fifth harmonic with a peak of 20. Notice that the third harmonic completes three cycles during the one cycle of the fundamental and thus has a frequency three times that of the fundamental. Similarly, the fifth harmonic completes five cycles during one cycle of the fundamental and thus has a frequency five times that of the fundamental. Each of the harmonics shown in Fig. 17.1 can be expressed as a function of time:

$$V_1 = 100\sin(\omega t), \qquad V_3 = 50\sin(3\omega t), \qquad V_5 = 20\sin(5\omega t) \tag{17.1}$$

Equation 17.1 shows three harmonic components of voltage or current that could be added together in an infinite number of ways by varying the phase angles of the three components. Thus, an infinite number of waveforms could be produced from these three harmonic components. For example, suppose V_3 is shifted in time by 60° and then added to V_1 and V_5. In this case, all three waveforms have a positive peak at 90° and a negative peak at 270°. One half cycle of the resultant waveform is shown in Fig. 17.2, which is clearly beginning to look like a pulse. In this case, we have used the harmonic components to synthesize a waveform. Generally, we would have a nonsinusoidal voltage or current waveform and would like to know its harmonic content. The question, then, is how to find the harmonic components given a waveform that repeats itself every cycle.

Fourier, the mathematician, showed that it is possible to represent any periodic waveform by a series of harmonic components. Thus, any periodic current or voltage in the power system can be represented by a Fourier series. Furthermore, he showed that the series can be found, assuming the waveform can be expressed as a mathematical function. We will not go into the mathematics behind the solution of Fourier series here; however, we can use the results. In particular, if a waveform $f(t)$ is periodic, with period T, then it can be approximated as

$$f(t) = a_0 + a_1\sin(\omega t + \theta_1) + a_2\sin(2\omega t + \theta_2) + a_3\sin(3\omega t + \theta_3) + \cdots + a_n\sin(n\omega t + \theta_n) \tag{17.2}$$

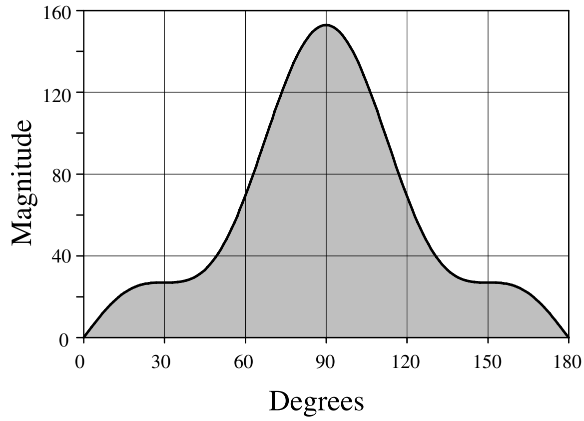

FIGURE 17.2 Pulse wave formed from the three harmonics in Eq. 17.1 with 60° shift for V_3.

where a_0 represents any DC (average) value of the waveform, a_1 through a_n are the Fourier amplitude coefficients, and θ_1 through θ_n are the Fourier phase coefficients. The amplitude coefficients are always zero or positive and the phase coefficients are all between 0 and 2π radians. As "n" gets larger, the approximation becomes more accurate.

For example, consider an alternating square wave of amplitude 100. The Fourier series can be shown to be

$$V_{\text{square}} = 100\sum_{n=1}^{\infty}\frac{1}{2n-1}\sin[(2n-1)\omega t] \tag{17.3}$$

Since the alternating waveform has zero average value, the coefficient a_0 is zero. Note also that only odd harmonics are included in the series given by Eq. (17.3), since $(2n-1)$ will always be an odd number, and all of the phase coefficients are zero. Expanding the first five terms of Eq. (17.3) yields:

$$V_{\text{square}} = 100\left[\sin(\omega t)+\frac{1}{3}\sin(3\omega t)+\frac{1}{5}\sin(5\omega t)+\frac{1}{7}\sin(7\omega t)+\frac{1}{9}\sin(9\omega t)\right] \tag{17.4}$$

Figure 17.3 shows one cycle for the waveform represented by the right-hand side of Eq. (17.3). Although only the first five terms of the Fourier series were used in Fig. 17.3, the resultant waveform already resembles a square wave. Harmonics have a number of effects on the power system as will be seen later, but for now we would like to have some way to indicate how large the harmonic content of a waveform is. One such figure of merit is the total harmonic distortion (THD).

Total harmonic distortion can be defined two ways. The first definition, in Eq. (17.5), shows the THD as a percentage of the fundamental component of the waveform, designated as THD_F. This is the IEEE definition of THD and is used widely in the United States.

$$\text{THD}_\text{F} = \frac{\sqrt{\sum_{h=2}^{\infty}V_{h\text{rms}}^2}}{V_{1\text{rms}}}\times 100\% \tag{17.5}$$

In Eq. (17.5), $V_{1\text{rms}}$ is the rms of the fundamental component and $V_{h\text{rms}}$ is the amplitude of the harmonic component of order "h" (i.e., the "hth" harmonic) . Although the symbol "V" is used in Eqs. (17.5) to (17.10), the equations apply to either current or voltage. The rms of a waveform composed of harmonics

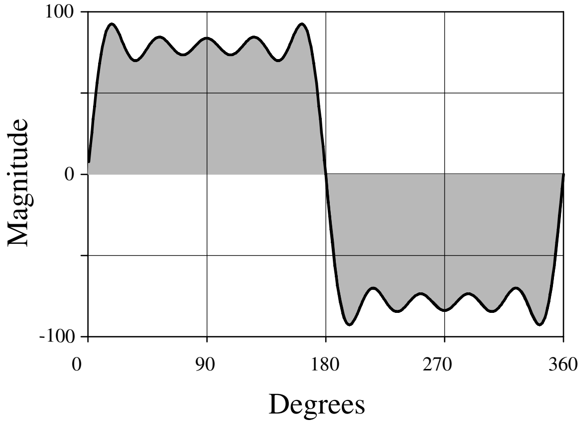

FIGURE 17.3 Approximation to a square wave using the first five terms of the Fourier series.

is independent of the phase angles of the Fourier series, and can be calculated from the rms values of all harmonics, including the fundamental:

$$V_{\text{rms}} = \sqrt{\sum_{h=1}^{\infty} V_{h\text{rms}}^2} \tag{17.6}$$

Because the series in Eq. (17.6) has only one more term (the rms of the fundamental) than the series in the numerator of Eq. (17.5), we can also find the total rms in terms of percent THD_F.

$$V_{\text{rms}} = V_{1\text{rms}} \sqrt{1 + \left(\frac{\%\text{THD}_\text{F}}{100}\right)^2} \tag{17.7}$$

In the opinion of some, Eq. (17.5) exaggerates the harmonic problem. Thus, another technique is also used to calculate THD. The alternate method, designated as THD_R, calculates THD as a percentage of the total rms instead of the rms of the fundamental. From Eq. (17.7), it is clear that the total rms will be larger than the rms of the fundamental, so such a calculation will yield a lower value for THD. This definition is used by the Canadian Standards Association and the IEC:

$$\text{THD}_\text{R} = \frac{\sqrt{\sum_{h=2}^{\infty} V_{h\text{rms}}^2}}{V_{\text{rms}}} \times 100\% \tag{17.8}$$

The value for THD_R can be obtained from THD_F by multiplying by $V_{1\text{rms}}$ and dividing by V_{rms}.

$$\text{THD}_\text{R} = \text{THD}_\text{F} \times \left(\frac{I_{1\text{rms}}}{I_{\text{rms}}}\right) \tag{17.9}$$

Substituting Eq. (17.7) into Eq. (17.9) yields another expression for THD_R in terms of THD_F:

$$\text{THD}_\text{R} = \frac{\text{THD}_\text{F}}{\sqrt{1 + \left(\frac{\%\text{THD}_\text{F}}{100}\right)^2}} \tag{17.10}$$

THD_R, as given by Eq. (17.8) and (17.10), will always be less than 100%. THD is very important because the IEEE Standard 519 specifies maximum values of THD for the utility voltage and the customer's current. Having considered what harmonics are, we can now look at some of their properties. The next section deals with the phase sequence of various harmonics.

Harmonic Sequence

In a three-phase system, the rotation of the phasors is assumed to have an A-B-C sequence as shown in Fig. 17.4a. As the phasors rotate, phase A passes the *x*-axis, followed by phase B and then phase C. An A-B-C sequence is called the *positive sequence*. However, phase A could be followed by phase C and then phase B, as shown in Fig. 17.4b. A set of phasors whose sequence is reversed is called the *negative sequence*. Finally, if the waveforms in all three phases were identical, their phasors would be in line with each other as shown in Fig. 17.4c. Because there are no phase angles between the three phases, this set of phasors is call the *zero sequence*.

When negative and zero sequence currents and voltages are present along with the positive sequence, they can have serious effects on power equipment. Not all harmonics have the same sequence; in fact, the sequence depends on the number of the harmonic, as shown in Fig. 17.5. Figure 17.5a, b, and c show the fundamental component of a three-phase set of waveforms (voltage or current) as well as their second harmonics. In each case, the phase-angle relationship has been chosen so both the fundamental and the second harmonic cross through zero in the ascending direction at the same time.

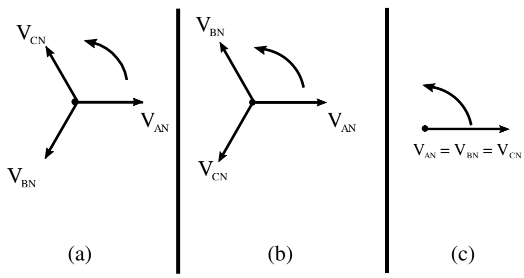

FIGURE 17.4 Positive (a), negative (b), and zero (c) sequences.

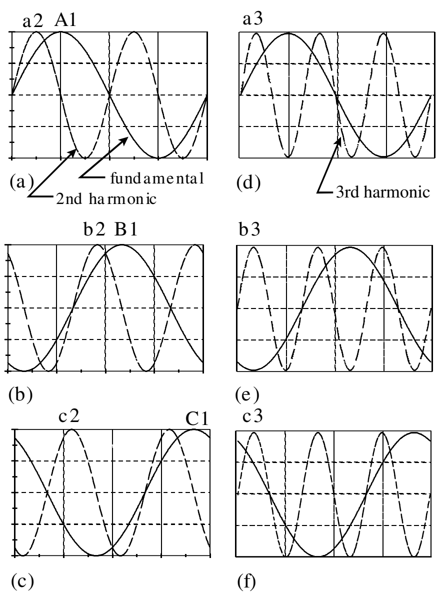

FIGURE 17.5 First, second, and third harmonics.

To establish the sequence of the fundamental components, label the positive peak values of the three phases A1, B1, and C1. Clearly, A1 occurs first, then B1, and finally C1. Thus, we can conclude that the fundamental component has an A-B-C, or positive, sequence. In fact, it was chosen to have a positive sequence. Given that the fundamental has a positive sequence, we can now look at other harmonics. In a similar manner, the first peak of each of the second harmonics are labeled a2, b2, and c2. In this case, a2 occurs first, but it is followed by c2 and then b2. The second harmonic thus has an A-C-B, or negative, sequence.

Now consider Fig. 17.5d, e, and f, which also show the same fundamental components, but instead of the second harmonic, the third harmonic is shown. Both the fundamental and third harmonics were chosen so they cross through zvoero together. When the peaks of the third harmonics are labeled as a3, b3, and c3, it is evident that all three occur at the same time. Since the third harmonics are concurrent, they have no phase order. Thus, they are said to have zero sequence. If the process in Fig. 17.5 was continued, the fourth harmonic would have a positive sequence, the fifth a negative sequence, the sixth a zero sequence, and so on.

All harmonics whose order is $3n$, where n is any positive integer, are zero sequence and are called *triplen* harmonics. Triplen harmonics cause serious problems in three-phase systems as discussed later in this section. First, however, consider what causes harmonics in the power system.

Where Do Harmonics Come From?

Electrical loads that have a nonlinear relationship between the applied voltages and their currents cause harmonic currents in the power system. Passive electric loads consisting of resistors, inductors, and capacitors are linear loads. If the voltage applied to them consists of a single-frequency sine wave, then the current through them will be a single-frequency sine wave as well. Power electronic equipment creates harmonic currents because of the switching elements that are inherent in their operation. For example, consider a simple switched-mode power supply used to provide DC power to devices such as desktop computers, televisions, and other single-phase electronic devices.

Figure 17.6 shows an elementary power supply in which a capacitor is fed from the power system through a full-wave, diode bridge rectifier. The instantaneous value of the AC source must be greater than the voltage across the capacitor for the diodes to conduct. When first energized, the capacitor charges to the peak of the AC waveform and, in the absence of a load, the capacitor remains charged and no further current is drawn from the source.

If there is a load, then the capacitor acts as a source for the load. After the capacitor is fully charged, the AC voltage waveform starts to decrease, and the diodes shut off. While the diode is off, the capacitor discharges current to the DC load, which causes its voltage, V_{dc}, to decrease. Thus, when the AC source becomes larger than V_{dc} during the next half-cycle, the capacitor draws a pulse of current to restore its charge.

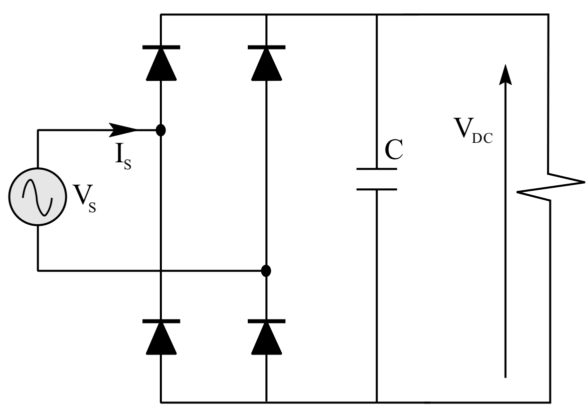

FIGURE 17.6 Simple single-phase switch-mode power supply.

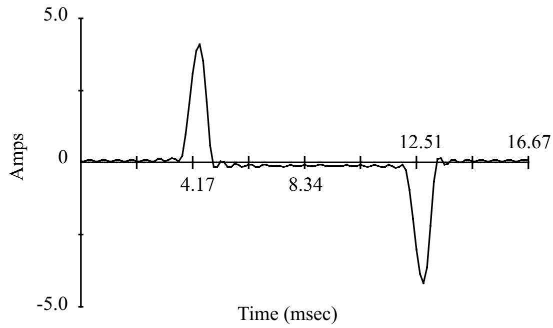

FIGURE 17.7 Input current to single-phase, full-wave rectifier.

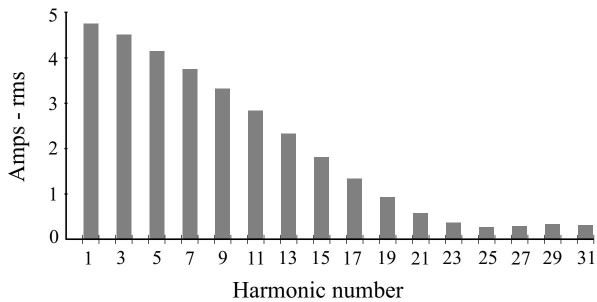

FIGURE 17.8 Harmonic spectrum of current for the circuit shown in Fig. 17.6.

Figure 17.7 shows the current of such a load (actually the input current to a variable-speed motor drive). Since the current has a repetitive waveform, it is composed of a series of harmonics. The harmonics can be found using a variety of test equipment with the capability to process a fast Fourier transform (FFT). This particular waveform has a large amount of harmonics, as shown by the harmonic spectrum (through the 31st harmonic) in Fig. 17.8. Note that the first several harmonics after the fundamental are almost as large as the fundamental. This waveform, as shown in Fig. 17.7, has a peak value of 4.25 A, but the rms of the waveform is only 1.03 A. This leads to another quantity that is an indicator of harmonic distortion. The *crest factor* (CF) is defined as the ratio of the peak value of the waveform divided by the rms value of the waveform:

$$\text{CF} = \frac{\text{peak of waveform}}{\text{rms of waveform}} \tag{17.11}$$

For the current shown in Fig. 17.7, the crest factor is 4.25 divided by 1.03, or 4.12. For a sinusoidal current or voltage, the crest factor would be the square root of 2 (1.414). Waveforms whose crest factor are substantially different from 1.414 will have harmonic content. Note that the crest factor can also be lower than 1.414. A square wave, for example, would have a CF of 1.

As shown in Fig. 17.8, the third harmonic of a single-phase bridge rectifier is very large. Putting such loads on the three phases of a three-phase, wye-connected system could cause problems because the third harmonics add on the neutral conductor. The best way to handle these problems is to eliminate the triplen harmonics.

Whereas single-phase rectifiers require a large amount of triplen current, three-phase bridge rectifiers do not. Figure 17.9 shows the input current and harmonic content for a three-phase bridge rectifier (again, the input current to a variable-frequency motor drive). In this case, the phase current contains two pulses in each half-cycle, which results in the elimination of all the triplen harmonics. Examination of

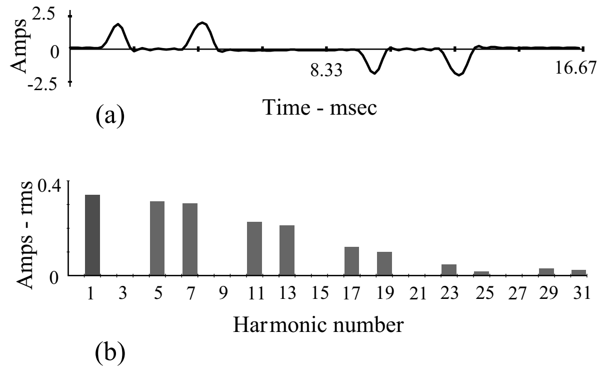

FIGURE 17.9 Line current and harmonic content for three-phase bridge rectifier.

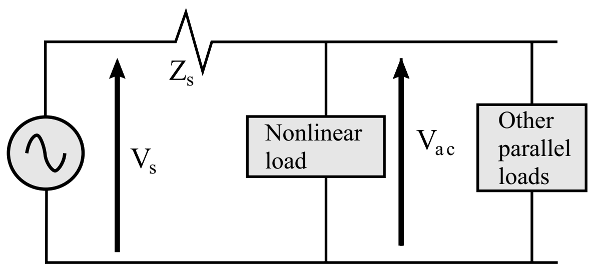

FIGURE 17.10 Simple single-phase power system.

the spectrum in Fig. 17.9 shows that the only harmonics that remain are those whose order numbers are of the form:

$$h = 6n \pm 1 \tag{17.12}$$

where n is any positive integer, beginning with 1. Setting $n = 1$, indicates the 5th and 7th harmonics will be present, $n = 2$ yields the 11th and 13th harmonics, and so on.

Harmonic currents have many impacts on the power system, both on the components of the system as well as the voltage. The next section considers some of these effects.

Effects of Harmonics on the System Voltage

A simple circuit representing a single-phase power system is shown in Fig. 17.10. In North America, the utility generates a 60-Hz sinusoidal voltage, indicated by the ideal source. However, the load current flows through transmission lines, transformers, and distribution feeders, which all have impedance. The impedance of the system is represented in Fig. 17.10 by Z_s. Finally, the load for this system is considered to be a nonlinear load in parallel with other loads.

Harmonic currents drawn from the power system by nonlinear loads create harmonic voltages ($RI + j\omega_h LI$) across the system impedance, and their effect can be significant for higher-order harmonics because inductive reactance increases with frequency. The load voltage is the difference between the source voltage and the voltage drop across the system impedance. Since the voltage drop across the system impedance contains harmonic components, the load voltage may become distorted if the nonlinear loads are a large fraction of the system capacity.

Referring back to Fig. 17.6, note the current pulse drawn by the rectifier occurs only when the AC source voltage is near its peak. This means the voltage drop across the source impedance will be large when

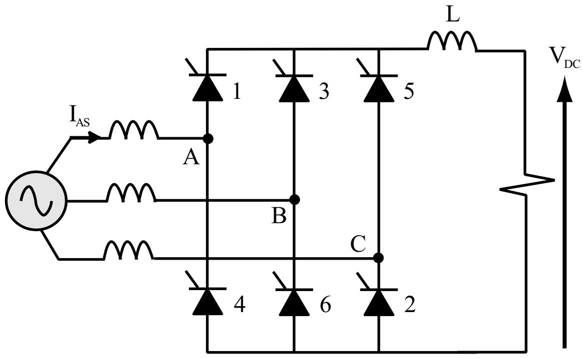

FIGURE 17.11 Three-phase bridge rectifier.

the source voltage is near its peak and essentially zero during the remainder of the half-cycle. Thus, the voltage delivered to the load will be "flattened" by the subtraction of the system impedance voltage drop. Unfortunately, some power electronic devices, such as the rectifier front-end of motor drives, are sensitive to the peak value of the AC voltage waveform, and may shut down or operate incorrectly when the incoming AC voltage is distorted. Voltage distortion affects the nonlinear load that created the harmonics and any other load that is connected in parallel with it. The interface between the loads and the power system is called the point of common coupling (PCC), and the PCC is where the harmonic content of system voltage and current must be controlled to comply with IEEE Standard 519. Although three-phase rectifiers do not cause triplen harmonic currents, they do cause another problem as a result of their operation.

Notching

A three-phase bridge rectifier is shown in Fig. 17.11, the details of which are described in Chapter 4 of this handbook. However, consider briefly how the diodes operate. Each diode in the top or bottom half conducts while one or two diodes in the oppositie half conduct. For example, diode 1 is connected to phase A and conducts during the period of time when diode 6 (phase B) and diode 2 (phase C) are conducting. Clearly, diode 1 should not conduct when diode 4 is conducting as that would constitute a short circuit. The inductor in series with the DC load tends to keep the current constant, so current must be passed from one diode to another. This transfer of the load current from one diode to another is called *commutation*. While diode 1 is conducting in the upper half of the bridge, the current in the lower half of the bridge will commutate from diode 6 to diode 2. Since the three-phase source has inductance as well, this transfer of current cannot occur instantaneously. Instead, the current in diode 2 must increase while the current in diode 6 decreases.

While it is conducting, a diode is essentially a short circuit, so during the commutation interval, two diodes in one side of the bridge are conducting. This results in two phases of the source being shorted together. For example, while the load current commutates from diode 6 to diode 2, points B and C are connected together, which means the voltage from B to ground and from C to ground is the same. The effect of commutation is to create a notch in the voltage waveform. Figure 17.12 shows the voltage from A to ground as calculated by a simulation of a three-phase bridge rectifier. The notching effect is evident, six times per cycle.

Notching is a repetitive event and the voltage waveform shown in Fig. 17.12 could be represented by a Fourier series. However, the order of the harmonics is extremely high, well above the range of many monitors normally used for making power quality measurements. Thus, notching is a special case somewhere between harmonics and transients. Devices connected in parallel with the bridge rectifier could be affected by notching, especially if the rectifier load is large relative to the size of the system from which it is fed.

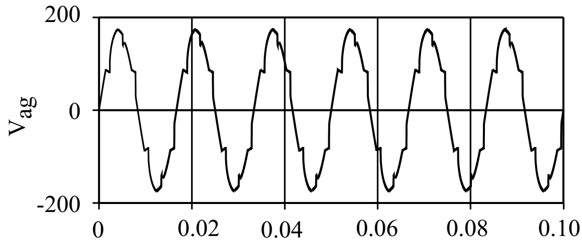

FIGURE 17.12 Voltage notching of the AC source voltage due to commutation of diodes in a three-phase rectifier.

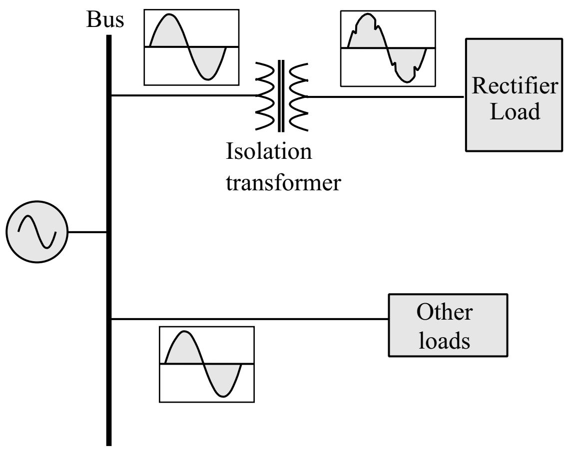

FIGURE 17.13 Use of an isolation transformer to keep notching from affecting other loads.

An isolation transformer can be used to supply the offending equipment and thus reduce the amount of notching seen by other loads. Figure 17.13 shows a rectifier load and other loads fed from a common bus with an isolation transformer between the rectifier load and the bus. The voltage on the secondary of the transformer is notched; however, the voltage on the primary side is relatively unaffected because the impedance of the isolation transformer tends to smooth out the notches. Thus, the other loads do not see the notching or at least see much smaller notches in the voltage waveform.

Effects of Harmonics on Power System Components

Harmonic currents from nonlinear loads can seriously affect electric power distribution equipment. Components that may be affected include transformers, conductors, circuit breakers, bus bars and connecting lugs, and electrical panels. Harmonic problems can occur in both single-phase and three-phase systems.

Conductors

Higher-order harmonic current components cause additional I^2R heating in every conductor through which they flow, because conductor resistance increases with frequency as a result of the skin effect. This means that as the frequency of a current increases, its ability to “soak” into a conductor is reduced, resulting in a higher current density at the edge of the conductor than at its center. A conductor can be carrying rated current (rms amps) and still overheat if the current contains significant higher-order harmonics. Because every conductor carrying the harmonic currents will have increased losses, there will be more heat to be dissipated in the system and the overall efficiency of the system will be reduced.

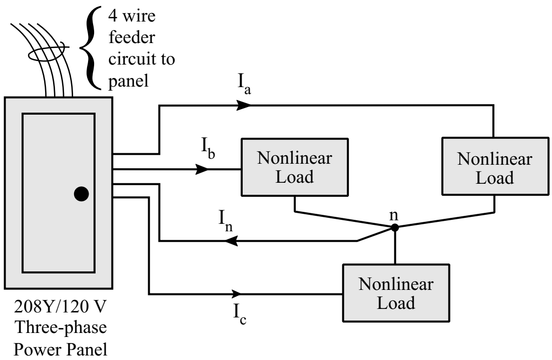

FIGURE 17.14 Three-phase power system with balanced, single-phase nonlinear loads.

Three-Phase Neutral Conductors

Triplen harmonics pose a problem for the neutral conductor in three-phase, wye-connected systems, such as the one shown in Fig. 17.14. Therein, a feeder circuit provides three-phase power to a circuit breaker panel board from which branch circuits provide power to outlets and lighting, including three single-phase loads connected via a four-wire branch circuit. When identical linear loads are placed on each of the three phases, the phase currents add to zero at point "*n*" and no current flows on the neutral wire. Again assuming linear loads, then even if the load are not identical, the current in the neutral could not be higher than the highest phase current.

If the loads in Fig. 17.14 are nonlinear, there will be harmonic currents in each phase. For balanced loads, the fundamentals and all non-triplen harmonic currents add to zero at the neutral point. If triplen harmonics are present in the phase currents, however, they will be in phase and add directly on the branch circuit and feeder neutrals. Since the neutral conductors carry the sum of the triplens from the three phases, the neutral current can actually exceed the current in the phase conductors. Since neutral conductors are not protected by circuit breakers, this can damage to the conductors.

To find the current in the neutral, we must recognize that all positive and negative sequence harmonics from the three phases will cancel out at the neutral point. The triplen harmonics, on the other hand, will *add* together at the neutral:

$$I_{N\mathrm{rms}} = 3 \times (I_{3\mathrm{rms}}^2 + I_{6\mathrm{rms}}^2 + I_{9\mathrm{rms}}^2 + \cdots)^{1/2} \tag{17.13}$$

From Eq. (17.13), it is evident that the neutral current is three times the rms of all the triplens on one phase of the system.

Transformers

Current flow in the windings and flux in the ferromagnetic core cause real power losses in a transformer. Because of their higher frequencies, harmonic currents cause additional losses in every conductor through which they flow, including the conductors of the transformer coils. Harmonic currents in the windings create harmonic flux components in the core of the transformer, which cause additional hysteresis and eddy current losses in the steel. Hysteresis loss is proportional to the frequency of the magnetic flux, and eddy currents are proportional to frequency squared. Thus, harmonic currents can cause significant increases in the core loss of the transformer. These additional losses may result in transformer overheating and electrical insulation failure.

To provide for the effects of nonlinear loads, manufacturers build specially designed transformers, called "K-factor rated," that are capable of supplying rated output current to loads with a specific level of harmonic content. K-factor-rated transformers have larger conductors in the windings and thinner, low-loss steel laminations in the core to reduce the losses. A transformer with a K-factor of K-1 is rated only for single-frequency current; thus, if the load is nonlinear, the transformer cannot provide rated current without overheating. Transformers rated K-4, K-9, K-13, and higher are available to provide power to nonlinear loads. K-factors of K-4 or K-9 indicate the transformer can supply rated current to loads that would increase the eddy current loss of a K-1 transformer by a factor of 4 or 9, respectively. Transformers rated K-9 or K-13 would likely be required for office areas containing many desktop computers, copy machines, fax machines, and electronic lighting ballasts. A large variable-speed motor drive could require a transformer rated K-30 or higher. The K-factor of a load can be calculated, if the harmonic components are known, as follows:

$$K = \sum h^2 \left(\frac{I_{h,\mathrm{rms}}}{I_{\mathrm{tot,rms}}} \right)^2 \tag{17.14}$$

where h is the harmonic order number, $I_{h,\mathrm{rms}}$ is the rms of the harmonic current whose frequency is "h" times the fundamental frequency, and $I_{\mathrm{tot,rms}}$ is the rms of the total current.

Effects of Harmonics on System Power Factor

Earlier, Eq. (17.6) showed that the addition of harmonic currents to the fundamental component increases the total rms current. Because they affect the rms value of the current, harmonics will affect the power factor of the circuit. Consider the voltage and current waveforms shown in Fig. 17.15 in which current lags the voltage by an angle θ. The apparent power of the circuit would be found by multiplying the rms voltage magnitude by the rms current magnitude. Power factor, F_p, is then defined as the ratio of the real power to the apparent power:

$$F_p = \frac{P}{V_{\mathrm{rms}} I_{\mathrm{rms}}} = \cos(\theta) \tag{17.15}$$

For linear loads, the phase shift (time displacement) between voltage and current results in different values for real power and apparent power. Since the current can only lag or lead the voltage by 0 to 90°, the power factor will always be positive and less than or equal to 1.

Instead of a sinusoidal current, suppose the current and voltage shown by Fig. 17.16. The current is the quasi-square wave, consisting of the Fourier series shown in Eq. (17.4). The voltage is a sine wave,

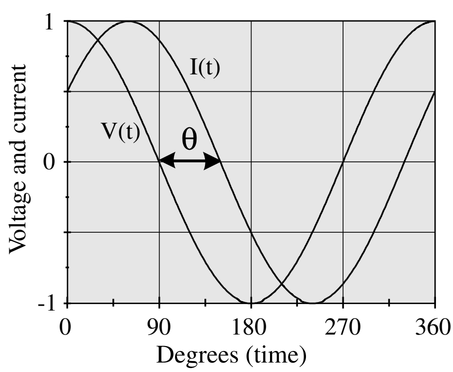

FIGURE 17.15 Voltage and current for a lagging load.

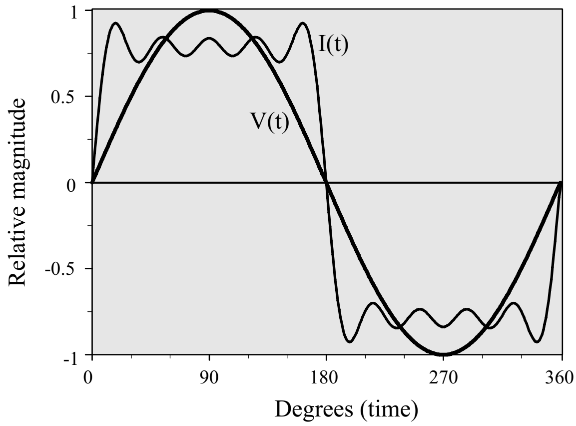

FIGURE 17.16 Sinusoidal voltage and quasi-square wave current.

which is in phase with the fundamental harmonic component of the current. The power can be found as a function of time by multiplying the voltage times the current at each time step.

Because the voltage consists of a single component, the power is a series of terms consisting of the voltage times each harmonic component of current. The first term of the series is of the form $\sin^2 \omega t$ since the voltage is in phase with the fundamental current component. Obviously, this term is always positive; therefore, it indicates real power (energy) being delivered to the load.

The remaining terms contain the product of the fundamental frequency voltage and one of the higher-order harmonic current components. Multiplying two sinusoidal waveforms of different frequencies creates a sinusoidal waveform, which has a zero average value. Thus, none of higher-order harmonic currents produces real power if the voltage is a single frequency. Substituting Eq. (17.7) for the total rms current into Eq. (17.15) yields a new expression for the power factor:

$$F_{p\,\mathrm{tot}} = \frac{P}{V_{1\mathrm{rms}} I_{\mathrm{rms}}} = \frac{P}{V_{1\mathrm{rms}} I_{1\mathrm{rms}} \sqrt{1 + \left(\dfrac{\%\mathrm{THD_F}}{100}\right)^2}} \tag{17.16}$$

where the current $\%\mathrm{THD_F}$ is used in the denominator of Eq. (17.16). Rewriting Eq. (17.16):

$$F_{p\,\mathrm{tot}} = \frac{P}{V_{1\mathrm{rms}} I_{1\mathrm{rms}}} \times \frac{1}{\sqrt{1 + \left(\dfrac{\%\mathrm{THD_F}}{100}\right)^2}} \tag{17.17}$$

In Eqs. (17.16) and (17.17), the subscript "tot" indicates the *total power factor*, which is sometimes called the *true power factor*. The total power factor in Eq. (17.17) is the product of two components, the first of which is called the *displacement power factor*:

$$F_{p\mathrm{disp}} = \frac{P}{V_{1\mathrm{rms}} I_{1\mathrm{rms}}} \tag{17.18}$$

The second component of the total power factor is the *distortion power factor*, which results from the harmonic components in the current:

$$F_{p\,\mathrm{disp}} = \frac{1}{\sqrt{1 + \left(\dfrac{\%\mathrm{THD_F}}{100}\right)^2}} \tag{17.19}$$

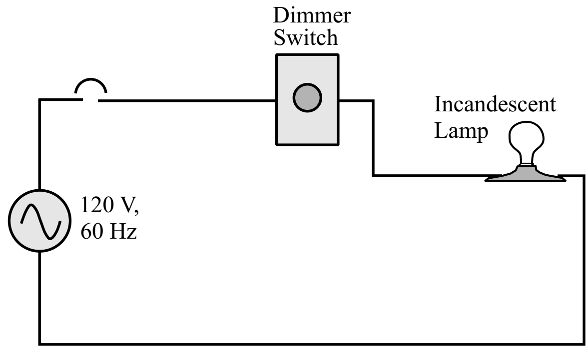

FIGURE 17.17 Simple incandescent lamp dimmer circuit.

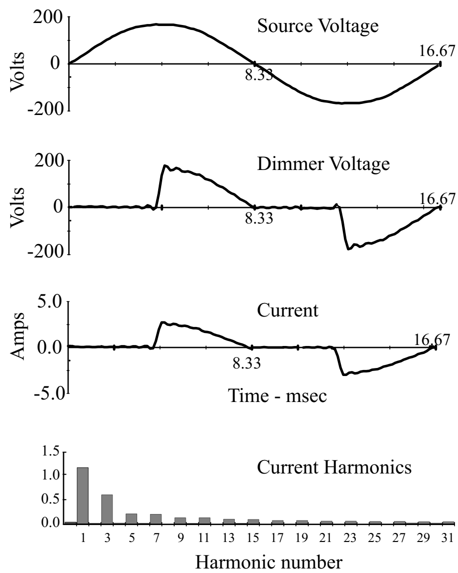

FIGURE 17.18 Source voltage, lamp voltage, lamp current, and current harmonic spectrum for the system shown in Fig. 17.17.

In the event that the voltage also has harmonic components, then the distortion power factor would be the product of two terms similar to the right side of Eq. (17.19), one for the voltage and one for the current. However, voltage distortion is normally very low compared with the current distortion.

Power electronic devices can cause unusual results with respect to power factor. The circuit shown in Fig. 17.17 consists of an incandescent lamp fed by a simple wall-mounted dimmer. Incandescent lamps operate at essentially unity power. In this case, the lamp voltage was set at 85 V rms by adjusting the dimmer switch. The voltage was observed at the source and at the lamp and the circuit current was measured, all with a harmonics analyzer.

The results are shown in Fig. 17.18, where the top waveform is the source voltage, with a 60-Hz component of 118 V and a 5th harmonic (300-Hz) component of 1.6 V. The second trace shows the lamp voltage, i.e, the dimmer output. At this voltage setting, the dimmer was conducting for approximately one half (90°) of each half-cycle, as shown. The bottom waveform shows the lamp current. Because the incandescent lamp is essentially a resistive load, the current waveform looks identical to the lamp voltage, except for the scale. The bar chart in Fig. 17.18 shows the harmonic spectrum of the current, which, except for the scale, was identical to the voltage spectrum.

Since the lamp voltage and current have the same shape and are in phase, the harmonic components of current and voltage are in phase. With no phase angle between each of the voltage and current components, the power factor of the lamp is unity, and that was found to be the case with the harmonic analyzer. Because the dimmer creates harmonic voltages to the lamp, each harmonic of voltage and its respective current delivers some power to the lamp at unity power factor.

Looking at the source results in a much different picture. The source voltage consists almost solely of a single frequency, whereas the current contains all of the harmonics shown in Fig. 17.18. Since the product of two sine waves of different frequencies is another sine wave, only the fundamental harmonic of current can deliver real power to the circuit. The harmonics analyzer showed that the fundamental component of the current lagged the source voltage by 28°. Taking the cosine of 28° results in a measured displacement power factor of 0.88 at the source.

THD_F for the current was found to be 60.7%, and the distortion power factor was calculated to be 0.855 from Eq. (17.19). The product of the distortion power factor and the displacement power factor yields the total power factor, 0.75 in this case. Thus, the incandescent lamp, a resistive load, appears to the power system as a 0.75 power factor lagging load. Low power factor results in higher losses in the system due to higher I^2R losses. In fact, both I and R increase in this case because the rms current is higher due to the harmonics and because skin effect causes higher resistance in the conductors. While a single lamp on a dimmer switch does not seriously affect the power system, very large nonlinear loads, such as a DC motor drive, could require the installation of harmonic filters to reduce the distortion power factor. Passive harmonic filters are briefly described here and in more detail in Section 17.3.

Power Factor Correction Capacitors

Many industrial loads are inductive, so capacitors are often used to improve the power factor. Although capacitors do not cause harmonics, they can resonate with the inductance of the power system. When resonant frequencies occur near harmonic frequencies, capacitors can amplify the harmonic currents created by nonlinear loads. Figure 17.19 shows a power circuit including power factor correction capacitors. The parallel combination of the system inductance and the power factor correction capacitors has a resonant frequency, f_r. The resonant frequency is given by

$$f_r = \frac{1}{\sqrt{LC}} \tag{17.20}$$

where L is the system inductance (X_s divided by 2π), and C is the capacitance.

Normally, we do not deal with inductance and capacitance, however. It is much more convenient to express the resonant frequency in other terms. In particular, we normally size power factor capacitors in kVAR. Figure 17.19 also shows a switch that can be closed to create a short circuit. If the switch is closed to short out the loads, the source voltage will be dropped across the system impedance, which in this

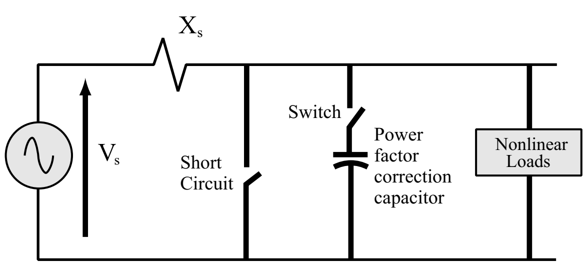

FIGURE 17.19 Circuit demonstrating how resonance can form with power factor correction.

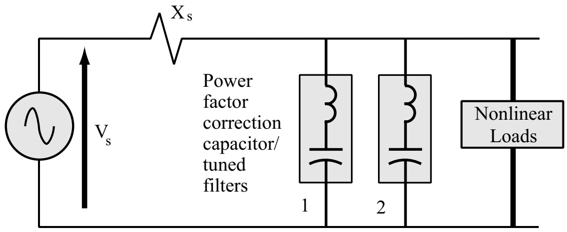

FIGURE 17.20 Use of harmonic filters.

case is considered to be inductive. Thus, the short-circuit kVA can be calculated as

$$\mathrm{kVA_{sc}} = \frac{V^2}{X_s} \tag{17.21}$$

The utility normally provides the available short-circuit capacity upon request. Neglecting the voltage drop across X_s during normal operation, the total kVAR of the capacitance would be

$$\mathrm{kVAR_{cap}} = \frac{V^2}{X_{\mathrm{cap}}} \tag{17.22}$$

Since $X_s = 2\pi f L$ and $X_{\mathrm{cap}} = 1/(2\pi f C)$, it can be shown from Eqs. (17.20) through (17.22) that

$$h_r \cong \sqrt{\frac{\mathrm{kVA_{sc}}}{\mathrm{kVAR_{cap}}}} \tag{17.23}$$

where h_r is the multiple of the system frequency at which the resonance occurs.

For example, if h_r is five, the resonant frequency is 300 Hz for a 60-Hz power system. Unfortunately, it is not uncommon for the value calculated by Eq. (17.23) to be near the 5th harmonic, which, as we have seen in Fig. 17.9, is the dominant harmonic for some three-phase bridge rectifiers. When the capacitors cause a resonance near one of the harmonics, the original harmonic current can be amplified by as much as a factor of 16, which can in turn cause excessive voltage drop and voltage distortion, damage to the capacitors, and lower power factor.

When harmonics cause serious voltage distortion, tuned filters can be used to reduce the amount of harmonic current drawn from the source. Figure 17.20 shows a circuit with two filters, each designed to reduce the effects of one particular harmonic. The inductance added in series with the capacitor should be chosen to create a series resonance frequency that is slightly below the frequency of the harmonic that is to be reduced. For example, if it was desired to reduce the 5th and 7th harmonics, then the filters would be designed to have resonance frequencies about 4.7 and 6.7 times the normal system frequency. This allows for tolerances in the actual values of the devices and causes the majority of the 5th and 7th harmonic currents to be diverted through the filters. A small portion of the harmonic current is still supplied by the source.

IEEE Standard 519

Recognizing the problems caused by nonlinear loads, the IEEE Standards board approved a revised and renamed Standard 519 in the fall of 1992. The 1981 version of the standard was titled, "Guide for Harmonic Control and Reactive Compensation of Static Power Converters." The 1981 version recommended specific limits for voltage THD from the utility, but did not recognize the possibility of customer load currents causing voltage distortion. The 1992 version was titled, "IEEE Recommended Practices and

TABLE 17.1 Harmonics Allowed by IEEE Standard 519

SCR	$h < 11$	11–15	17–21	23–33	$h > 33$	% THD
<20	4.0	2.0	1.5	0.6	0.3	5.0
20–50	7.0	3.5	2.5	1.0	0.5	8.0
50–100	10.0	4.5	4.0	1.5	0.7	12.0
100–1000	12.0	5.5	5.0	2.0	1.0	15.0
>1000	15.0	7.0	6.0	2.5	1.4	20.0

Requirements for Harmonic Control in Electrical Power Systems." The new version places the responsibility for ensuring power quality on both the utility and the customer.

As indicated by the title, IEEE Standard 519 is a "recommended practice," which means it is not a law or rule for all utility–customer interfaces, but it may be used as a design guideline for new installations. Utilities may also include the requirements from Standard 519 in service agreements with their customers, which could result in financial penalties for customers that do not comply. The standard makes the customer responsible for limiting the harmonic currents injected into the power system and the utility responsible for avoiding unacceptable voltage distortion.

IEEE Standard 519 defines harmonic current limits (shown in Table 17.1) for individual customers at the point of common coupling (PCC). Because voltage distortion is caused by the amount of harmonic currents in the system, larger customers are capable of causing more voltage distortion than smaller ones. Recognizing this, the standard allows a higher current THD for smaller customers' loads. The short-circuit ratio (SCR) is used to differentiate customer size.

When the load of Fig. 17.19 was shorted, the only impedance limiting the current was the system impedance. That current is called the available short-circuit current, and is generally high since the system impedance is much lower than the load impedance. SCR is defined as the "average maximum demand (load) current" for the facility divided by the available short-circuit current. The maximum load current drawn by a large customer would be a higher fraction of the available short-circuit current, so the large customer's SCR would be lower. The lower the SCR, the more stringent are the IEEE 519 limitations on harmonic currents.

IEEE Standard 519 also provides limits for specific ranges of frequencies, as shown in Table 17.1. Higher-order harmonics are constrained to have lower amplitudes for two reasons. First, higher-order harmonics cause greater voltage distortion than lower-order harmonics, even if they have the same amplitude, because the system inductive reactance is proportional to frequency. Second, interference with telecommunication equipment is more severe for higher-frequency harmonics. Note that Table 17.1 applies only to odd harmonics; even harmonics are limited to 25% of the values for the ranges they would occupy in Table 17.1.

The utility is required by Standard 519 to maintain acceptable levels of voltage distortion. Below 69 kV, individual harmonic components in the voltage should not exceed 3% of the fundamental, and the voltage THD must be less than 5%. Higher voltages have even lower limits, but those apply primarily to utility interconnections.

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17.3 Passive Harmonic Filters

Badrul H. Chowdhury

Currently in the United States, only 15 to 20% of the utility distribution loading consists of nonlinear loads. Loads, such as AC and DC adjustable speed drives (ASD), power rectifiers and inverters, arc furnaces, and discharge lighting (metal halide, fluorescent, etc.), and even saturated transformers, can be considered nonlinear devices. It is projected over the next 10 years that such nonlinear loads will comprise approximately 70 to 85% of the loading on utility distribution systems in the United States. These loads may generate enough harmonics to cause distorted current and voltage waveshapes.

The deleterious effects of harmonics are many. A significant impact is equipment overheating because of the presence of harmonics in addition to the fundamental. Harmonics can also create resonance conditions with power factor correction capacitors, resulting in higher than normal currents and voltages. This can lead to improper operation of protective devices, such as relays and fuses.

Harmonic frequency currents can cause additional rotating fields in AC motors. Depending on the frequency, the motor will rotate in the opposite direction (countertorque). In particular, the 5th harmonic, which is the most prevalent harmonic in three-phase power systems, is a negative sequence harmonic causing the motor to have a backward rotation, thus shortening the service life.

A typical current wave, as drawn by a three-phase AC motor drive, may look like the waveshape shown in Fig. 17.21. A Fourier analysis of the current would reveal the nature of the harmonics present. Three-phase ASDs generate primarily the 5th and 7th current harmonics and a lesser amount of 11th, 13th, and higher orders. The triplen harmonics (3rd, 9th, 15th, i.e., odd multiples of three) are conspicuously missing, as is usually the case in six-pulse converters, giving them an added advantage over single-phase converters. However, the triplen harmonics are additive in the neutral and can cause dangerous overheating.

In general, the characteristic harmonics generated by a converter is given by

$$h = pn \pm 1 \begin{cases} p = 6, 12, 18, \ldots \\ n = 1, 2, \ldots \end{cases} \tag{17.24}$$

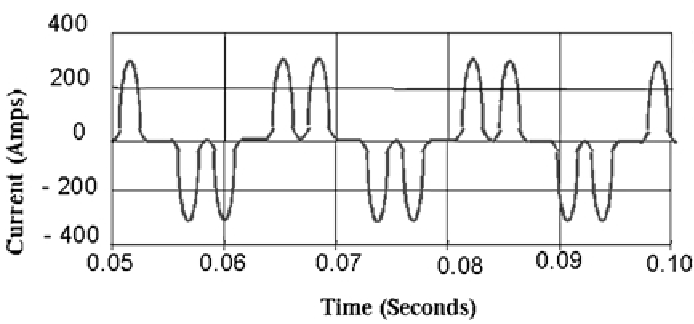

FIGURE 17.21 Typical current waveform of a three-phase adjustable-speed drive.

where h is the order of harmonics, n is any integer, and p is the number of pulses generated in each cycle (six for a three-phase converter).

To understand the impact of harmonics and to design remedies, one must quantify the amount of harmonics present. This is done by combining all of the harmonic frequency components (voltage or current) with the fundamental component (voltage or current) to form the total harmonic distortion, or THD. A commonly accepted definition of THD is as follows:

$$\mathrm{THD_I} = \frac{\sqrt{I_2^2 + I_3^2 + I_4^2 + \cdots}}{I_1} \times 100\% = \frac{\sqrt{\sum_{h=2}^{\infty} I_h^2}}{I_1} \times 100\% \tag{17.25}$$

where I_1 is the fundamental component of the current, I_2 is the second harmonic, I_3 the third harmonic, and so on. A similar equation can be written for voltage distortion.

Any THD values over 5% are significant enough for concern. Harmonic current distortion greater than 5% will contribute to the additional heating of a power transformer, so it must be derated for harmonics. It is not uncommon for THD levels in industrial plants to reach 25%. Normally, THD levels in office settings will be lower than in industrial plants, but office equipment is much more susceptible to variations in power quality. Odd-number harmonics (3rd, 5th, 7th, etc.) are of the greatest concern in the electrical distribution system. Even-number harmonics are usually mitigated because the harmonics swing equally in both the positive and negative direction. Pesky harmonics can be mitigated by the use of passive and active filters. Passive filters, consisting of tuned series L-C circuits, are the most popular. However, they require careful application, and may produce unwanted side effects, particularly in the presence of power factor correction capacitors.

The active filter concept uses power electronics to produce harmonic components that cancel the harmonic components from the nonlinear loads so that the current supplied from the source is sinusoidal. These filters are costly and relatively new.

Passive harmonic filters are constructed from passive elements (resistors, inductors, and capacitors) and thus the name. These filters are highly suited for use in three-phase, four-wire electrical power distribution systems. They should be applied as close as possible to the offending loads, preferably at the farthest three- to single-phase point of distribution. This will ensure maximum protection for the upstream system. Harmonics can be substantially reduced to as low as 30% by use of passive filters.

Passive filters can be categorized as parallel filters and series filters. A parallel filter is characterized as a series resonant and trap-type exhibiting a low impedance at its tuned frequency. Deployed close to the source of distortion, this filter keeps the harmonic currents out of the supply system. It also provides some smoothing of the load voltage. This is the most common type of filter.

The series filter is characterized as a parallel resonant and blocking type with high impedance at its tuned frequency. It is not very common because the load voltage can be distorted.

Series Passive Filter

This configuration is popular for single-phase applications for the purpose of minimizing the 3rd harmonic. Other specific tuned frequencies can also be filtered. Figure 17.22 shows the basic diagram of a series passive filter.

The advantages of a series filter are that it:

- Provides high impedance to tuned frequency;
- Does not introduce any system resonance;
- Does not import harmonics from other sources;
- Improves displacement power factor and true power factor.

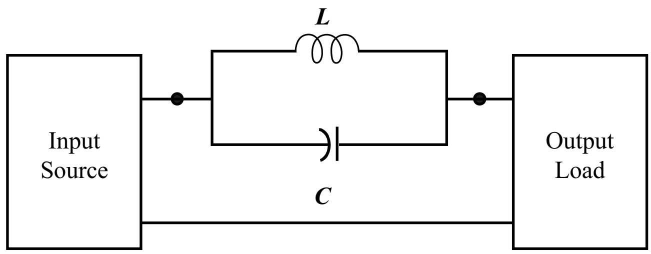

FIGURE 17.22 A series passive filter.

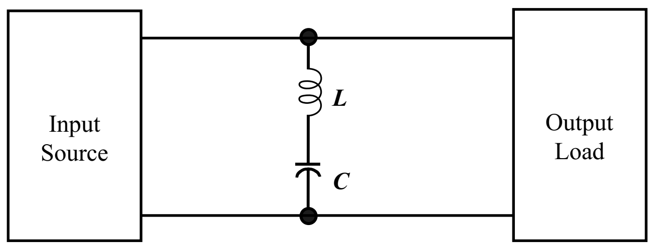

FIGURE 17.23 A shunt passive filter.

Some disadvantages are that it:

- Must handle the rated full load current;
- Is only minimally effective other than tuned harmonic frequencies;
- Can supply nonlinear loads only.

Shunt Passive Filter

The shunt passive filter is also capable of filtering specific tuned harmonic frequencies such as, 5th, 7th, 11th, etc. Figure 17.23 shows a commonly used diagram of a shunt filter. The advantages of a parallel filter are that it:

- Provides low impedance to tuned frequency;
- Supplies specific harmonic component to load rather than from AC source;
- Is only required to carry harmonic current and not the full load current;
- Improves displacement power factor and true power factor.

Some disadvantages are that:

- It only filters a single (tuned) harmonic frequency;
- It can create system resonance;
- It can import harmonics form other nonlinear loads;
- Multiple filters are required to satisfy typical desired harmonic limits.

Series Passive AC Input Reactor

The basic configuration is shown in Fig. 17.24. This type filters all harmonic frequencies, by varying amounts. The advantages of a series reactor are:

- Low cost;
- Higher true power factor;
- Small size;

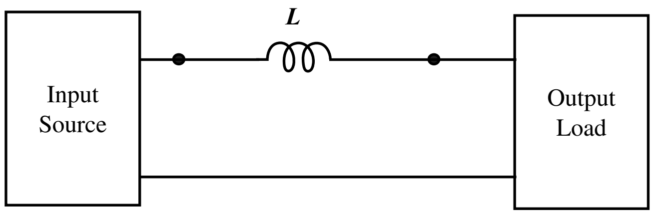

FIGURE 17.24 A series passive AC input reactor.

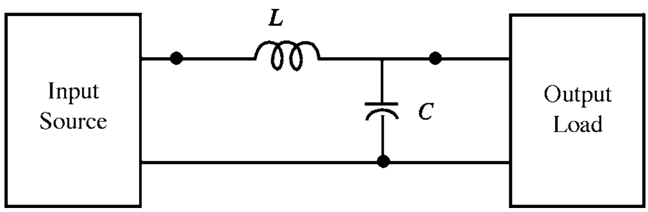

FIGURE 17.25 Low-pass filter.

- Filter does not create system resonance;
- It protects against power line disturbances.

Some disadvantages are that it:

- Must handle the rated full load current;
- Can only improve harmonic current distortion to 30 to 40% at best;
- Only slightly reduces displacement power factor.

Low-Pass (Broadband) Filter

The basic configuration is shown in Fig. 17.25. It is capable of eliminating all harmonic frequencies above the resonant frequency. The specific advantages of a low-pass filter are that it:

- Minimizes all harmonic frequencies;
- Supplies all harmonic frequencies as opposed to the AC source supplying those frequencies;
- Does not introduce any system resonance;
- Does not import harmonics from other sources;
- Improves true power factor.

Some of the disadvantages are that it:

- Must handle the rated full load current;
- Can supply nonlinear loads only.

Passive Filter Design

The filter design process involves a number of steps that will ensure lowest possible cost and proper performance under the THD limits. Figure 17.26 shows a flowchart of the entire process.

Characterizing Harmonic-Producing Loads

This is the first step in the process that will produce a summary of the level of harmonics being generated by nonlinear loads, such as AC adjustable speed drives, power rectifiers, arc furnaces, etc. Harmonic measurements must be used to characterize the level of harmonic generation for an existing nonlinear load.

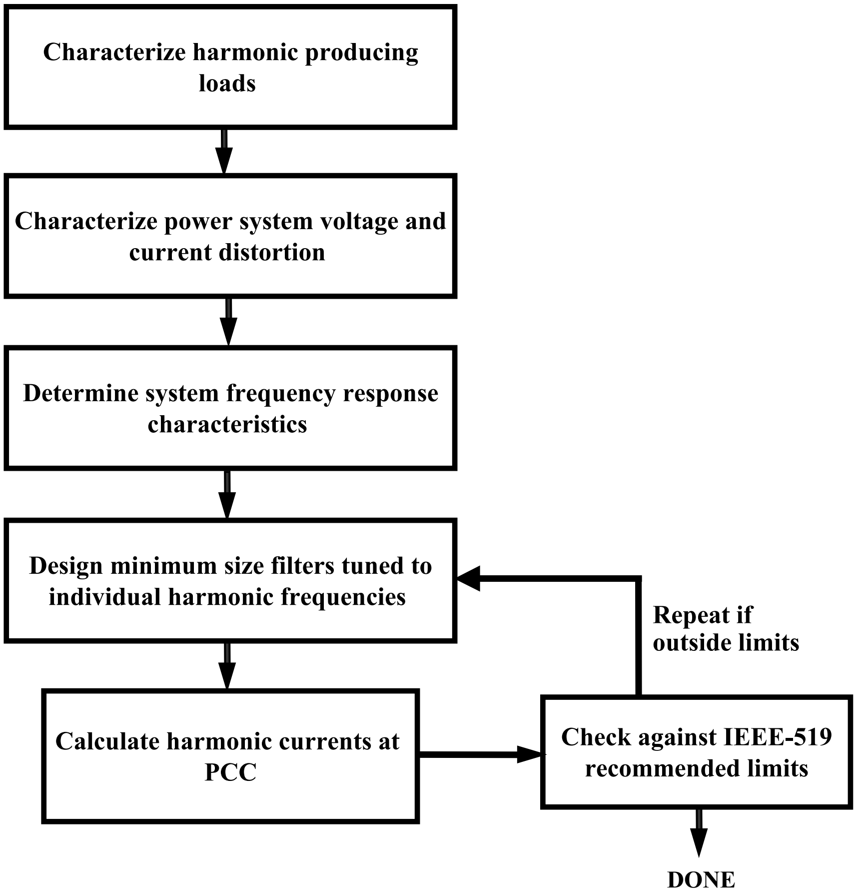

FIGURE 17.26 Flowchart for harmonic filter design.

Characterizing Power System Voltage and Current Distortion

In this step, a power system model is developed for analysis. The model is developed from one-line diagrams, manufacturer's data for various electrical equipment, the utility system characteristics, such as fault MVA, representative impedance, nominal voltage level, and the loading information. Figure 17.27 shows a sample representation of a utility system and an industrial plant supplied by a step-down transformer. The equivalent utility system can be represented as a simple impedance consisting of a resistance and an inductive reactance.

Determining System Frequency Response Characteristics

Switching transients created from regular utility operations as well as harmonics emanating from non-linear loads can both be magnified by power factor correction capacitors if resonant conditions exist. Therefore, it is necessary to perform simulations or frequency scans to determine the frequency response characteristics, looking from the low voltage bus. Simulations can be easily carried out by representing the system as a Thevenin's equivalent circuit. Such a circuit is shown in Fig. 17.28.

In the figure, L_{eq} and R_{eq} represent the combined inductance and resistance of the utility system and the step-down transformer.

$$Z_{in} = \frac{(R_{eq} + j\omega L_{eq})(-j/\omega C)}{R_{eq} + j\omega L_{eq} - j/\omega C} \tag{17.26}$$

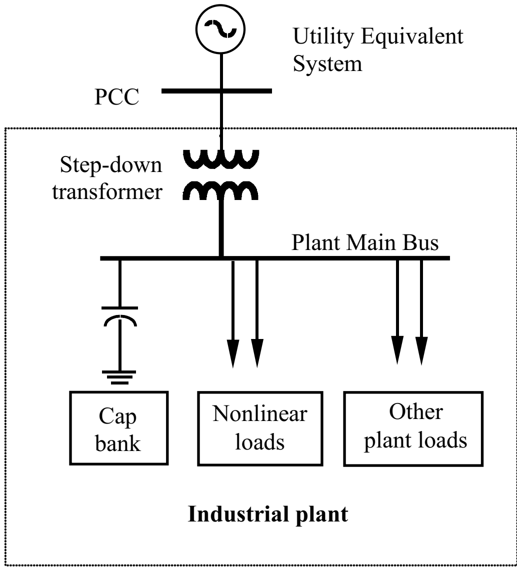

FIGURE 17.27 A typical representation of an industrial plant being supplied by a utility system.

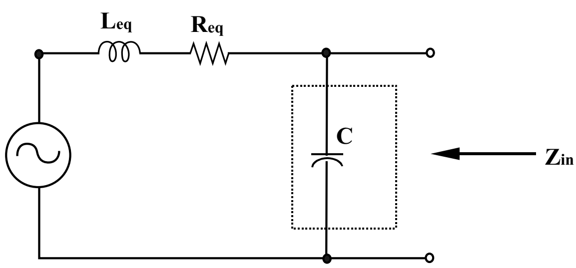

FIGURE 17.28 System equivalent circuit with reactive compensation at the load.

Parallel resonance occurs when the imaginary part of the denominator is equal to zero. That is,

$$\omega_0 L_{eq} - 1/\omega_0 C = 0 \tag{17.27}$$

Solving for ω_0:

$$\omega_0 = \frac{1}{[L_{eq}C]^{1/2}} \tag{17.28}$$

In hertz:

$$f_0 = \frac{1}{2\pi[L_{eq}C]^{1/2}} \tag{17.29}$$

Frequency scan output consists of magnitude and phase angle for the driving point impedance. The effect of important system parameters, such as a capacitor, is evaluated and the potential for problem resonance conditions is determined. Figure 17.29 shows a typical output of a frequency scan simulation for studying the impact of power factor correction capacitors. Figure 17.30 depicts the proximity of the resonance points to some of the important harmonic characteristics, such as the 5th, the 7th, the 11th, and the 13th harmonics for varying levels of capacitive compensation.

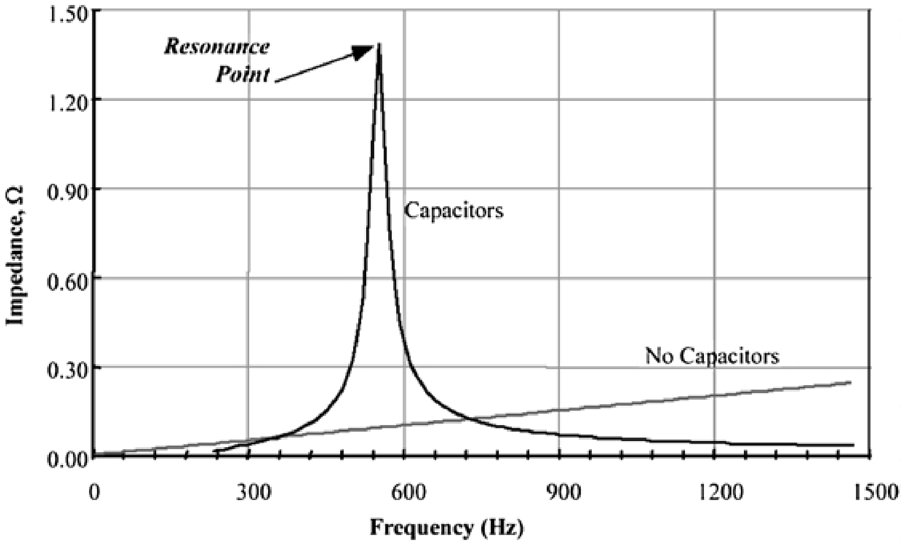

FIGURE 17.29 Frequency scan showing resonance point.

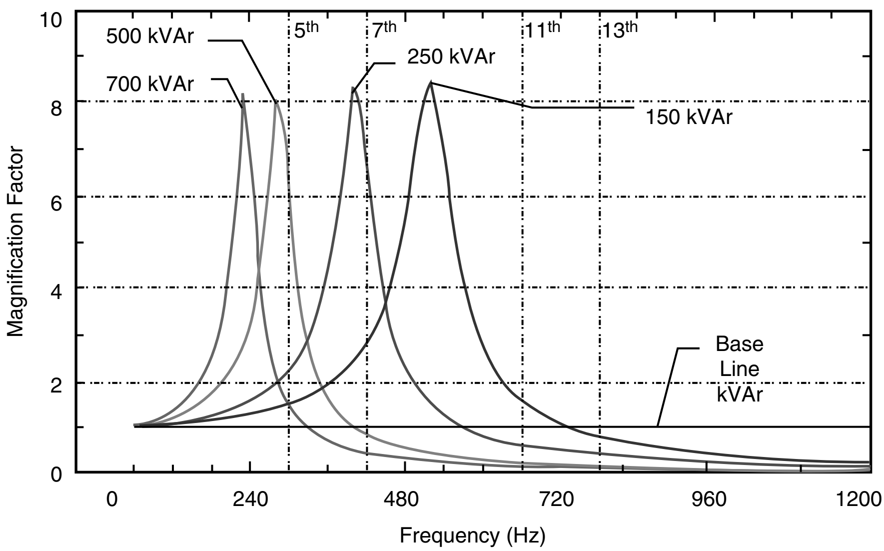

FIGURE 17.30 Resonance magnification due to low voltage reactive compensation.

Designing Minimum Size Filters Tuned to Individual Harmonic Frequencies

If not filtered, the harmonics generated by the industrial customer, downstream of the plant main bus, are returned upstream to the point of common coupling (PCC). If the short-circuit level available at the PCC is high enough, the voltage distortion will be significantly high and will thus affect other customers.

One of the most widely adopted solutions for reducing the impact of harmonics is the application of capacitor banks as tuned harmonic filters. This amounts to a very cost-effective solution since power factor correction capacitors are quite commonly installed in industrial facilities. To avoid resonance, one can use inductors in series with power factor capacitors to produce a harmonic filter. The inductor allows the parallel resonant frequency to be shifted somewhat. Figure 17.31a and b show a tuned delta-connected and a wye-connected filter bank, respectively. Without the series inductor, the bank simply becomes a

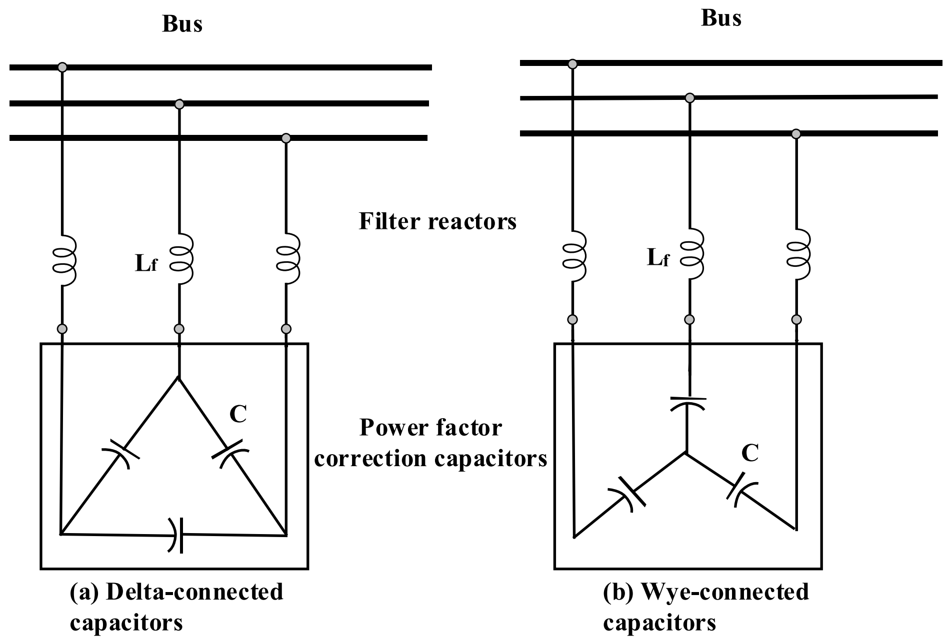

FIGURE 17.31 Application of capacitor banks as tuned harmonic filters.

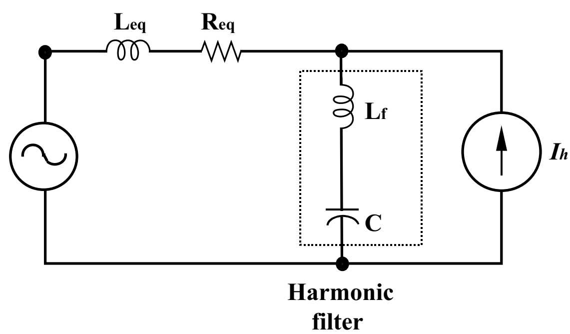

FIGURE 17.32 Equivalent circuit of the compensated system showing the harmonic filter equivalent.

power factor correction capacitor bank. This combination is referred to as *harmonic filter bank* or *detuned capacitor bank.*

The *series resonant frequency* or tuning frequency of the filter is selected to be about 3 to 10% below the lowest-order harmonic produced by the load. For typical six-pulse converters, this happens to be the 5th harmonic or a frequency of 300 Hz (on a 60-Hz system).

Typically, the tuning frequency of the filter is 282 Hz, corresponding to the 4.7th harmonic. In addition to shifting the parallel resonant frequency, the filter also supplies a portion of the harmonic current demanded by the load. Hence, the source current has less of the 5th harmonic content.

Figure 17.32 shows the equivalent circuit of the compensated system where

I_h = hth harmonic current source representing contributions from all harmonic sources at the plant main bus
L_f = inductance of the series reactor in the filter
C = capacitance of the harmonic filter

Depending on the severity of the individual harmonic level, harmonic filters should be tuned to a harmonic that causes the resonance condition. Some useful equations are given below.

Series resonant frequency of the filter element is

$$f_0 = \frac{1}{2\pi[L_f C]^{1/2}} \tag{17.30}$$

where L_f = inductance of the tuning inductor and C = capacitance of the capacitor bank.

Solving for L_f:

$$L_f = \frac{1}{C[2\pi f_0]^2} \tag{17.31}$$

Since

$$C = \frac{1}{2\pi f_{\text{sys}} X_{\text{cap}}} = \frac{1}{\omega_{\text{sys}} X_{\text{cap}}} \tag{17.32}$$

where

$$X_{\text{cap}} = \frac{KV_{\text{cap}}^2}{MVA_{\text{cap}}} \tag{17.33}$$

$$L_f = \frac{f_{\text{sys}} X_{\text{cap}}}{2\pi(f_0)^2} = \frac{f_{\text{sys}} kV_{\text{cap}}^2}{2\pi(f_0)^2 MVA_{\text{cap}}} \tag{17.34}$$

Thus, the driving point impedance, as seen in Fig. 17.32, is

$$Z_{\text{in}} = \frac{(R_{\text{eq}} + j\omega L_{\text{eq}})(j\omega L_f - j/\omega C)}{R_{\text{eq}} + j\omega L_{\text{eq}} + j\omega L_f - j/\omega C} \tag{17.35}$$

Parallel resonance occurs when the imaginary part of the denominator is equal to zero. That is,

$$\omega_0 L_{\text{eq}} + \omega_0 L_f - 1/\omega_0 C = 0 \tag{17.36}$$

Solving for ω_0:

$$\omega_0 = \frac{1}{[(L_{\text{eq}} + L_f)C]^{1/2}} \tag{17.37}$$

In hertz:

$$f_0 = \frac{1}{2\pi[(L_{\text{eq}} + L_f)C]^{1/2}} \tag{17.38}$$

If the design is correct, then the frequency given by Eq. (17.38) should be farther away from the resonance point.

Calculating Harmonic Currents at PCC

To carry out simulations to estimate actual harmonic distortion levels, one needs to represent each harmonic-generating device, the system parameters, and the tuned filter characteristics. The output for these simulations consists of individual harmonic levels, bus voltage distortion and current distortion levels, and rms voltage and current levels. Harmonic levels are calculated at the PCC where the consumer's load connects to other loads in the power system.

Checking against IEEE-519 Recommended Limits

Current and voltage distortion levels as determined through simulations are compared with recommended limits outlined in IEEE Standard 519-1992. This standard is explained in the Appendix to this section. If harmonic voltage distortion levels are still not within acceptable limits, it is easy to change capacitor sizes and/or locations, or the size of the series reactor. These changes affect the frequency response characteristics of the industrial facility such that proximity to resonance points can be altered.

Further Information

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Appendix—IEEE Recommended Practices and Requirements for Harmonic Control in Electric Power Systems

IEEE 519-1992 recommends limits on harmonic distortion according to two distinct criteria:

1. There is a limitation on the amount of harmonic current that a consumer can inject into a utility network.
2. A limitation is placed on the level of harmonic voltage that a utility can supply to a consumer.

$$\mathrm{SCR} = \frac{\text{Short-circuit MVA}}{\text{Load MW}} = \frac{I_{\mathrm{SC}}}{I_L}$$

IEEE 519 is limited to being a collection of *Recommended Practices* that serve as a guide to both suppliers and consumers of electrical energy with regard to excessive harmonic current injection or excessive voltage distortion. All of the current distortion values are given in terms relative to the maximum demand load current. The total distortion is in terms of total demand distortion (TDD) instead of the more common THD term. I_{SC} = maximum short circuit current at PCC; I_L = maximum demand load current (fundamental frequency) at point of common coupling; TDD = Total demand distortion in % of max demand and given by

$$\mathrm{TDD} = \frac{\sqrt{\sum_{h=2}^{\infty} I_h^2}}{I_L} \times 100\%$$

17.4 Active Filters for Power Conditioning

Hirofumi Akagi

Much research has been performed on active filters for power conditioning and their practical applications since their basic principles of compensation were proposed around 1970 [Bird et al., 1969; Gyugyi and Strycula, 1976; Kawahira et al., 1983]. In particular, recent remarkable progress in the capacity and switching speed of power semiconductor devices such as insulated-gate bipolar transistors (IGBTs) has spurred interest in active filters for power conditioning. In addition, state-of-the-art power electronics technology has enabled active filters to be put into practical use. More than one thousand sets of active filters consisting of voltage-fed pulse width modulation (PWM) inverters using IGBTs or gate turn-off (GTO) thyristors are operating successfully in Japan.

Active filters for power conditioning provide the following functions:

- Reactive-power compensation,
- Harmonic compensation, harmonic isolation, harmonic damping, and harmonic termination,
- Negative sequence current/voltage compensation,
- Voltage regulation.

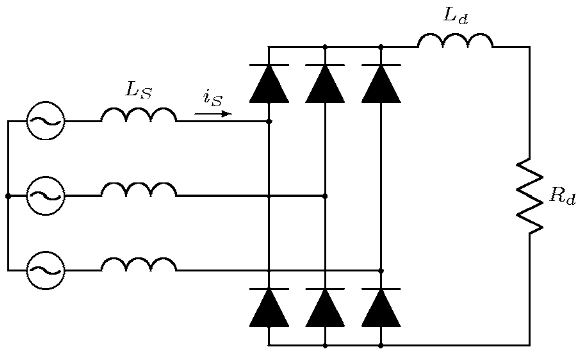
The term *active filters* is also used in the field of signal processing. In order to distinguish active filters in power processing from active filters in signal processing, the term *active power filters* often appears in many technical papers or literature. However, the author prefers *active filters for power conditioning* to active power filters, because the term active power filters is misleading to either active filters for power or filters for active power. Therefore, this section takes the term active filters for power conditioning or simply uses the term active filters as long as no confusion occurs.

Harmonic-Producing Loads

Identified Loads and Unidentified Loads

Nonlinear loads drawing nonsinusoidal currents from utilities are classified into identified and unidentified loads. High-power diode/thyristor rectifiers, cycloconverters, and arc furnaces are typically characterized as identified harmonic-producing loads because utilities identify the individual nonlinear loads installed by high-power consumers on power distribution systems in many cases. The utilities determine the point of common coupling with high-power consumers who install their own harmonic-producing loads on power distribution systems, and also can determine the amount of harmonic current injected from an individual consumer.

A “single” low-power diode rectifier produces a negligible amount of harmonic current. However, multiple low-power diode rectifiers can inject a large amount of harmonics into power distribution



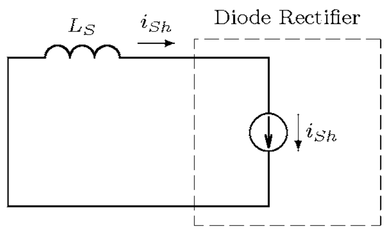


FIGURE 17.33 Diode rectifier with inductive load. (a) Power circuit; (b) equivalent circuit for harmonic on a per-phase base.

systems. A low-power diode rectifier used as a utility interface in an electric appliance is typically considered as an unidentified harmonic-producing load. Attention should be paid to unidentified harmonic-producing loads as well as identified harmonic-producing loads.

Harmonic Current Sources and Harmonic Voltage Sources

In many cases, a harmonic-producing load can be represented by either a harmonic current source or a harmonic voltage source from a practical point of view. Figure 17.33a shows a three-phase diode rectifier with a DC link inductor L_d. When attention is paid to voltage and current harmonics, the rectifier can be considered as a harmonic current source shown in Fig. 17.33b. The reason is that the load impedance is much larger than the supply impedance for harmonic frequency ω_h, as follows:

$$\sqrt{R_L^2 + (\omega_h L_d)^2} \gg \omega_h L_S$$

Here, L_S is the sum of supply inductance existing upstream of the point of common coupling (PCC) and leakage inductance of a rectifier transformer. Note that the rectifier transformer is disregarded from Fig. 17.33a. Figure 17.33b suggests that the supply harmonic current i_{Sh} is independent of L_S.

Figure 17.34a shows a three-phase diode rectifier with a DC link capacitor. The rectifier would be characterized as a harmonic voltage source shown in Fig. 17.34b if it is seen from its AC terminals. The reason is that the following relation exists:

$$\frac{1}{\omega_h C_d} \ll \omega_h L_S$$

This implies that i_{Sh} is strongly influenced by the inductance value of L_S.

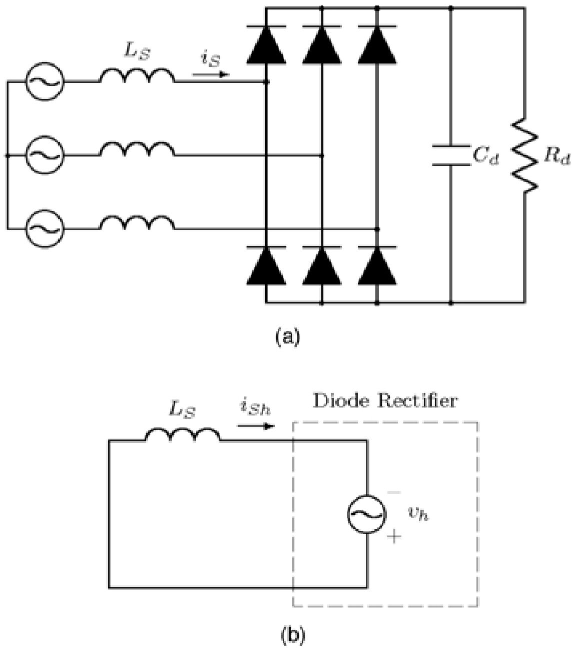


FIGURE 17.34 Diode rectifier with capacitive load. (a) Power circuit; (b) equivalent circuit for harmonic on a per-phase base.

Theoretical Approach to Active Filters for Power Conditioning

The Akagi-Nabae Theory

The theory of instantaneous power in three-phase circuits is referred to as the "Akagi-Nabae theory" [Akagi et al., 1983; 1984]. Figure 17.35 shows a three-phase three-wire system on the *a-b-c* coordinates, where no zero-sequence voltage is included in the three-phase three-wire system. Applying the theory to Fig. 17.35 can transform the three-phase voltages and currents on the *a-b-c* coordinates into the two-phase voltages and currents on the α-β coordinates, as follows:

$$\begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1\ 2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} e_a \\ e_b \\ e_c \end{bmatrix} \tag{17.39}$$

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & -1\ 2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} \tag{17.40}$$

As is well known, the instantaneous real power either on the *a-b-c* coordinates or on the α-β coordinates is defined by

$$p = e_a i_a + e_b i_b + e_c i_c = e_\alpha i_\alpha + e_\beta i_\beta \tag{17.41}$$

To avoid confusion, p is referred to as three-phase instantaneous real power. According to the theory, the three-phase instantaneous imaginary power, q, is defined by

$$q = e_\alpha i_\beta - e_\beta i_\alpha \tag{17.42}$$

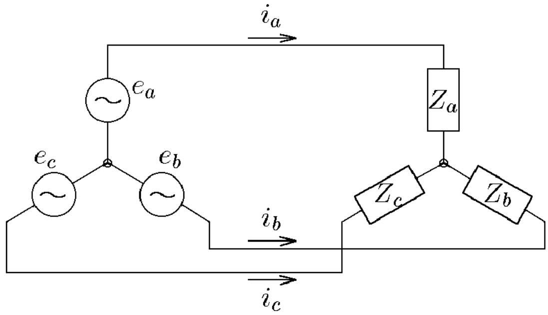


FIGURE 17.35 Three-phase three-wire system.

The combination of the above two equations bears the following basic formulation:

$$\begin{bmatrix} p \\ q \end{bmatrix} = \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \tag{17.43}$$

Here, $e_\alpha \cdot i_\alpha$ or $e_\beta \cdot i_\beta$ obviously means instantaneous power in the α-phase or the β-phase because either is defined by the product of the instantaneous voltage in one phase and the instantaneous current in the same phase. Therefore, p has a dimension of [W]. Conversely, neither $e_\alpha \cdot i_\beta$ nor $e_\beta \cdot i_\alpha$ means instantaneous power because either is defined by the product of the instantaneous voltage in one phase and the instantaneous current in the other phase. Accordingly, q is quite different from p in dimension and electric property although q looks similar in formulation to p. A common dimension for q should be introduced from both theoretical and practical points of view. A good candidate is [IW], that is, "imaginary watt."

Equation (17.43) is changed into the following equation:

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} p \\ q \end{bmatrix} \tag{17.44}$$

Note that the determinant with respect to e_α and e_β in Eq. (17.43) is not zero. The instantaneous currents on the α-β coordinates, i_α and i_β, are divided into two kinds of instantaneous current components, respectively:

$$\begin{aligned} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} &= \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} p \\ 0 \end{bmatrix} + \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} 0 \\ q \end{bmatrix} \\ &\equiv \begin{bmatrix} i_{\alpha p} \\ i_{\beta p} \end{bmatrix} + \begin{bmatrix} i_{\alpha q} \\ i_{\beta q} \end{bmatrix} \end{aligned} \tag{17.45}$$

Let the instantaneous powers in the α-phase and the β-phase be p_α and p_β, respectively. They are given by the conventional definition as follows:

$$\begin{bmatrix} p_\alpha \\ p_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha i_\alpha \\ e_\beta i_\beta \end{bmatrix} = \begin{bmatrix} e_\alpha i_{\alpha p} \\ e_\beta i_{\beta p} \end{bmatrix} + \begin{bmatrix} e_\alpha i_{\alpha q} \\ e_\beta i_{\beta q} \end{bmatrix} \tag{17.46}$$

The three-phase instantaneous real power, p, is given as follows, by using Eqs. (17.45) and (17.46):

$$\begin{aligned} p &= p_\alpha + p_\beta = e_\alpha i_{\alpha p} + e_\beta i_{\beta p} + e_\alpha i_{\alpha q} + e_\beta i_{\beta q} \\ &= \frac{e_\alpha^2}{e_\alpha^2 + e_\beta^2} p + \frac{e_\beta^2}{e_\alpha^2 + e_\beta^2} p + \frac{-e_\alpha e_\beta}{e_\alpha^2 + e_\beta^2} q + \frac{e_\alpha e_\beta}{e_\alpha^2 + e_\beta^2} q \end{aligned} \tag{17.47}$$

The sum of the third and fourth terms on the right-hand side in Eq. (17.47) is always zero. From Eqs. (17.46) and (17.47), the following equations are obtained:

$$p = e_\alpha i_{\alpha p} + e_\beta i_{\beta p} \equiv p_{\alpha p} + p_{\beta p} \tag{17.48}$$

$$0 = e_\alpha i_{\alpha q} + e_\beta i_{\beta q} \equiv p_{\alpha q} + p_{\beta q} \tag{17.49}$$

Inspection of Eqs. (17.48) and (17.49) leads to the following essential conclusions:

- The sum of the power components, $p_{\alpha p}$ and $p_{\beta p}$, coincides with the three-phase instantaneous real power, p, which is given by Eq. (17.41). Therefore, $p_{\alpha p}$ and $p_{\beta p}$ are referred to as the α-phase and β-phase instantaneous active powers.
- The other power components, $p_{\alpha q}$ and $p_{\beta q}$, cancel each other and make no contribution to the instantaneous power flow from the source to the load. Therefore, $p_{\alpha q}$ and $p_{\beta q}$ are referred to as the α-phase and β-phase instantaneous reactive powers.
- Thus, a shunt active filter without energy storage can achieve instantaneous compensation of the current components, $i_{\alpha q}$ and $i_{\beta q}$ or the power components, $p_{\alpha q}$ and $p_{\beta q}$. In other words, the Akagi–Nabae theory based on Eq. (17.43) exactly reveals what components the active filter without energy storage can eliminate from the α-phase and β-phase instantaneous currents, i_α and i_β or the α-phase and β-phase instantaneous real powers, p_α and p_β.

Energy Storage Capacity

Figure 17.36 shows a system configuration of a shunt active filter for harmonic compensation of a diode rectifier, where the main circuit of the active filter consists of a three-phase voltage-fed PWM inverter and a DC capacitor, C_d. The active filter is controlled to draw the compensating current, i_{AF}, from the utility, so that the compensating current cancels the harmonic current flowing on the AC side of the diode rectifier with a DC link inductor.

Referring to Eq. (17.44) yields the α-phase and β-phase compensating currents,

$$\begin{bmatrix} i_{AF\alpha} \\ i_{AF\beta} \end{bmatrix} = \begin{bmatrix} e_\alpha & e_\beta \\ -e_\beta & e_\alpha \end{bmatrix}^{-1} \begin{bmatrix} p_{AF} \\ q_{AF} \end{bmatrix} \tag{17.50}$$

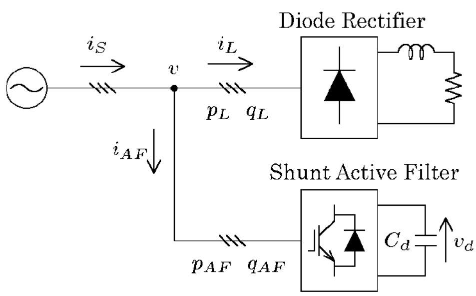


FIGURE 17.36 Shunt active filter.

Here, p_{AF} and q_{AF} are the three-phase instantaneous real and imaginary power on the AC side of the active filter, and they are usually extracted from p_L and q_L. Note that p_L and q_L are the three-phase instantaneous real and imaginary power on the AC side of a harmonic-producing load. For instance, when the active filter compensates for the harmonic current produced by the load, the following relationships exist:

$$p_{AF} = -\tilde{p}_L, \qquad q_{AF} = -\tilde{q}_L \tag{17.51}$$

Here, $\tilde{p}_L$ and $\tilde{q}_L$ are AC components of p_L and q_L, respectively. Note that the DC components of p_L and q_L correspond to the fundamental current present in i_L and the AC components to the harmonic current. In general, two high-pass filters in the control circuit extract $\tilde{p}_L$ from p_L and $\tilde{q}_L$ from q_L.

The active filter draws p_{AF} from the utility, and delivers it to the DC capacitor if no loss is dissipated in the active filter. Thus, p_{AF} induces voltage fluctuation of the DC capacitor. When the amplitude of p_{AF} is assumed to be constant, the lower the frequency of the AC component, the larger the voltage fluctuation [Akagi et al., 1984; 1986]. If the period of the AC component is one hour, the DC capacitor has to absorb or release electric energy given by integration of p_{AF} with respect to time. Thus, the following relationship exists between the instantaneous voltage across the DC capacitor, v_d and p_{AF}:

$$\frac{1}{2}C_d v_d^2(t) = \frac{1}{2}C_d v_d^2(0) + \int_0^t p_{AF}\, dt \tag{17.52}$$

This implies that the active filter needs an extremely large-capacity DC capacitor to suppress the voltage fluctuation coming from achieving "harmonic" compensation of $\tilde{p}_L$. Hence, the active filter is no longer a harmonic compensator, and thereby it should be referred to as a "DC capacitor-based energy storage system," although it is impractical at present. In this case, the main purpose of the voltage-fed PWM inverter is to perform an interface between the utility and the bulky DC capacitor.

The active filter seems to "draw" q_{AF} from the utility, as shown in Fig. 17.36. However, q_{AF} makes no contribution to energy transfer in the three-phase circuit. No energy storage, therefore, is required to the active filter, independent of q_{AF}, whenever $p_{AF} = 0$.

Classification of Active Filters

Various types of active filters have been proposed in technical literature [Moran, 1989; Grady et al., 1990; Akagi, 1994; Akagi and Fujita, 1995; 1997; Aredes et al., 1998]. Classification of active filters is made from different points of view [Akagi, 1996]. Active filters are divided into AC and DC filters. Active DC filters have been designed to compensate for current and/or voltage harmonics on the DC side of thyristor converters for high-voltage DC transmission systems [Watanabe, 1990; Zhang et al., 1993] and on the DC link of a PWM rectifier/inverter for traction systems. Emphasis, however, is put on active AC filter in the following because the term "active filters" refers to active AC filters in most cases.

Classification by Objectives: Who Is Responsible for Installing Active Filters?

The objective of "who is responsible for installing active filters" classifies them into the following two groups:

1. Active filters installed by *individual consumers* on their own premises in the vicinity of one or more identified harmonic-producing loads.
2. Active filters being installed by *electric power utilities* in substations and/or on distribution feeders.

Individual consumers should pay attention to current harmonics produced by their own harmonic-producing loads, and thereby the active filters installed by the individual consumers are aimed at compensating for current harmonics.

TABLE 17.2 Comparison of Shunt Active Filters and Series Active Filters

	Shunt Active Filter	Series Active Filter
System configuration	Fig. 17.36	Fig. 17.37
Power circuit of active filter	Voltage-fed PWM inverter *with* current minor loop	Voltage-fed PWM inverter *without* current minor loop
Active filter acts as	Current source: i_{AF}	Voltage source: v_{AF}
Harmonic-producing load suitable	Diode/thyristor rectifiers with *inductive* loads, and cycloconverters	Large-capacity diode rectifiers with *capacitive* loads
Additional function	Reactive power compensation	AC voltage regulation
Present situation	Commercial stage	Laboratory stage

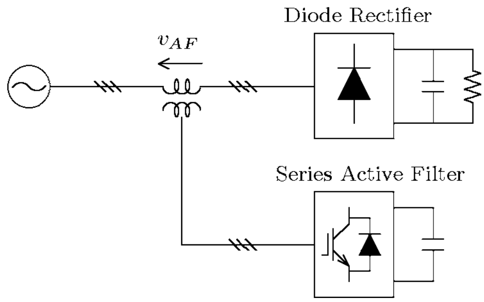


FIGURE 17.37 Series active filter.

Utilities should concern themselves with voltage harmonics, and therefore active filters will be installed by utilities in the near future for the purpose of compensating for voltage harmonics and/or of achieving "harmonic damping" throughout power distribution systems or "harmonic termination" of a radial power distribution feeder.

Classification by System Configuration

Shunt Active Filters and Series Active Filters

A stand-alone shunt active filter shown in Fig. 17.36 is one of the most fundamental system configurations. The active filter is controlled to draw a compensating current, i_{AF}, from the utility, so that it cancels current harmonics on the AC side of a general-purpose diode/thyristor rectifier [Akagi et al., 1990; Peng et al., 1990; Bhattacharya et al., 1998] or a PWM rectifier for traction systems [Krah and Holtz, 1994]. Generally, the shunt active filter is suitable for harmonic compensation of a current harmonic source such as diode/thyristor rectifier with a DC link inductor. The shunt active filter has the capability of damping harmonic resonance between an existing passive filter and the supply impedance.

Figure 17.37 shows a system configuration of a series active filter used alone. The series active filter is connected in series with the utility through a matching transformer, so that it is suitable for harmonic compensation of a voltage harmonic source such as a large-capacity diode rectifier with a DC link capacitor. The series active filter integrated into a diode rectifier with a DC common capacitor is discussed elsewhere. Table 17.2 shows comparisons between the shunt and series active filters. This concludes that the series active filter has a "dual" relationship in each item with the shunt active filter [Akagi, 1996; Peng, 1998].

Hybrid Active/Passive Filters

Figures 17.38 through 17.40 show three types of hybrid active/passive filters, the main purpose of which is to reduce initial costs and to improve efficiency. The shunt passive filter consists of one or more tuned LC filters and/or a high-pass filter. Table 17.3 shows comparisons among the three hybrid filters in which

TABLE 17.3 Comparison of Hybrid Active/Passive Filters

	Shunt Active Filter Plus Shunt Passive Filter	Series Active Filter Plus Shunt Passive Filter	Series Active Filter Connected in Series with Shunt Passive Filter
System configuration	Fig. 17.38	Fig. 17.39	Fig. 17.40
Power circuit of active filter	• Voltage-fed PWM inverter *with* current minor loop	• Voltage-fed PWM inverter *without* current minor loop	• Voltage-fed PWM inverter *with* or *without* current minor loop
Function of active filter	• Harmonic compensation	• Harmonic isolation	• Harmonic isolation or harmonic compensation
Advantages	• General shunt active filters applicable • Reactive power controllable	• Already existing shunt passive filters applicable • No harmonic current flowing through active filter	• Already existing shunt passive filters applicable • Easy protection of active filter
Problems or issues	• Share compensation in frequency domain between active filter and passive filter	• Difficult to protect active filter against overcurrent • No reactive power control	• No reactive power control
Present situation	• Commercial stage	• A few practical applications	• Commercial stage

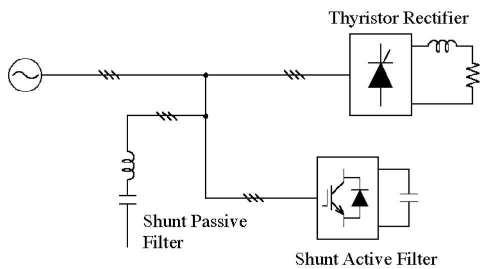


FIGURE 17.38 Combination of shunt active filter and shunt passive filter.

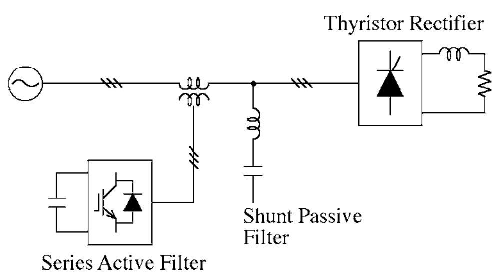


FIGURE 17.39 Combination of series active filter and shunt passive filter.

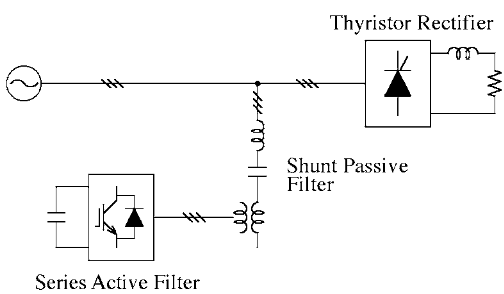


FIGURE 17.40 Series active filter connected in series with shunt passive filter.

the active filters are different in function from the passive filters. Note that the hybrid filters are applicable to any current harmonic source, although a harmonic-producing load is represented by a thyristor rectifier with a DC link inductor in Figs. 17.38 through 17.40.

Such a combination of a shunt active filter and a shunt passive filter as shown in Fig. 17.38 has already been applied to harmonic compensation of naturally-commutated twelve-pulse cycloconverters for steel mill drives [Takeda et al., 1987]. The passive filters absorbs 11th and 13th harmonic currents while the active filter compensates for 5th and 7th harmonic currents and achieves damping of harmonic resonance between the supply and the passive filter. One of the most important considerations in system design is to avoid competition for compensation between the passive filter and the active filter.

The hybrid active filters, shown in Fig. 17.39 [Peng et al., 1990; 1993; Kawaguchi et al., 1997] and in Fig. 17.40 [Fujita and Akagi, 1991; Balbo et al., 1994; van Zyl et al., 1995], are right now on the commercial stage, not only for harmonic compensation but also for harmonic isolation between supply and load, and for voltage regulation and imbalance compensation. They are considered prospective alternatives to pure active filters used alone. Other combined systems of active filters and passive filters or LC circuits have been proposed in Bhattacharya et al. [1997].

Classification by Power Circuit

There are two types of power circuits used for active filters: a voltage-fed PWM inverter [Akagi et al., 1986; Takeda et al., 1987] and a current-fed PWM inverter [Kawahira et al., 1983; van Schoor and van Wyk, 1987]. These are similar to the power circuits used for AC motor drives. They are, however, different in their behavior because active filters act as nonsinusoidal current or voltage sources. The author prefers the voltage-fed to the current-fed PWM inverter because the voltage-fed PWM inverter is higher in efficiency and lower in initial costs than the current-fed PWM inverter [Akagi, 1994]. In fact, almost all active filters that have been put into practical application in Japan have adopted the voltage-fed PWM inverter as the power circuit.

Classification by Control Strategy

The control strategy of active filters has a great impact not only on the compensation objective and required kVA rating of active filters, but also on the filtering characteristics in transient state as well as in steady state [Akagi et al., 1986].

Frequency-Domain and Time-Domain

There are mainly two kinds of control strategies for extracting current harmonics or voltage harmonics from the corresponding distorted current or voltage; one is based on the Fourier analysis in the frequency domain [Grady et al., 1990], and the other is based on the Akagi–Nabae theory in the time-domain. The concept of the Akagi–Nabae theory in the time-domain has been applied to the control strategy of almost all the active filters installed by individual high-power consumers over the last 10 years in Japan.

Harmonic Detection Methods

Three kinds of harmonic detection methods in the time-domain have been proposed for shunt active filters acting as a current source i_{AF}. Taking into account the polarity of the currents i_S, i_L and i_{AF} in Fig. 17.36 gives

$$\begin{aligned}
&\text{load-current detection:} && i_{AF} = -i_{Lh} \\
&\text{supply-current detection:} && i_{AF} = -K_S \cdot i_{Sh} \\
&\text{voltage detection:} && i_{AF} = K_V \cdot v_h
\end{aligned}$$

Note that load-current detection is based on feedforward control, while supply-current detection and voltage detection are based on feedback control with gains of K_S and K_V, respectively. Load-current detection and supply-current detection are suitable for shunt active filters installed in the vicinity of one or more harmonic-producing loads by individual consumers. Voltage detection is suitable for shunt active filters that will be dispersed on power distribution systems by utilities, because the shunt active filter based on voltage detection is controlled in such a way to present infinite impedance to the external circuit for the fundamental frequency, and to present a resistor with low resistance of $1/K_V$ [Ω] for harmonic frequencies [Akagi et al., 1999].

Supply-current detection is the most basic harmonic detection method for series active filters acting as a voltage source v_{AF}. Referring to Fig. 17.37 yields

$$\text{supply-current detection:} \quad v_{AF} = G \cdot i_{Sh}$$

The series active filter based on supply-current detection is controlled in such a way to present zero impedance to the external circuit for the fundamental frequency and to present a resistor with high resistance of G [Ω] for the harmonic frequencies. The series active filters shown in Fig. 17.37 [Fujita and Akagi, 1997] and Fig. 17.39 [Peng et al., 1990] are based on supply-current detection.

Integrated Series Active Filters

A small-rated series active filter integrated with a large-rated double-series diode rectifier has the following functions [Fujita and Akagi, 1997]:

- Harmonic compensation of the diode rectifier,
- Voltage regulation of the common DC bus,
- Damping of harmonic resonance between the communication capacitors connected across individual diodes and the leakage inductors including the AC line inductors,
- Reduction of current ripples flowing into the electrolytic capacitor on the common DC bus.

System Configuration

Figure 17.41 shows a harmonic current-free AC/DC power conversion system described below. It consists of a combination of a double-series diode rectifier of 5 kW and a series active filter with a peak voltage and current rating of 0.38 kVA. The AC terminals of a single-phase H-bridge voltage-fed PWM inverter are connected in "series" with a power line through a single-phase matching transformer, so that the combination of the matching transformers and the PWM inverters forms the "series" active filter. For small to medium-power systems, it is economically practical to replace the three single-phase inverters with a single three-phase inverter using six IGBTs. A small-rated high-pass filter for suppression of switching ripples is connected to the AC terminals of each inverter in the experimental system, although it is eliminated from Fig. 17.41 for the sake of simplicity.

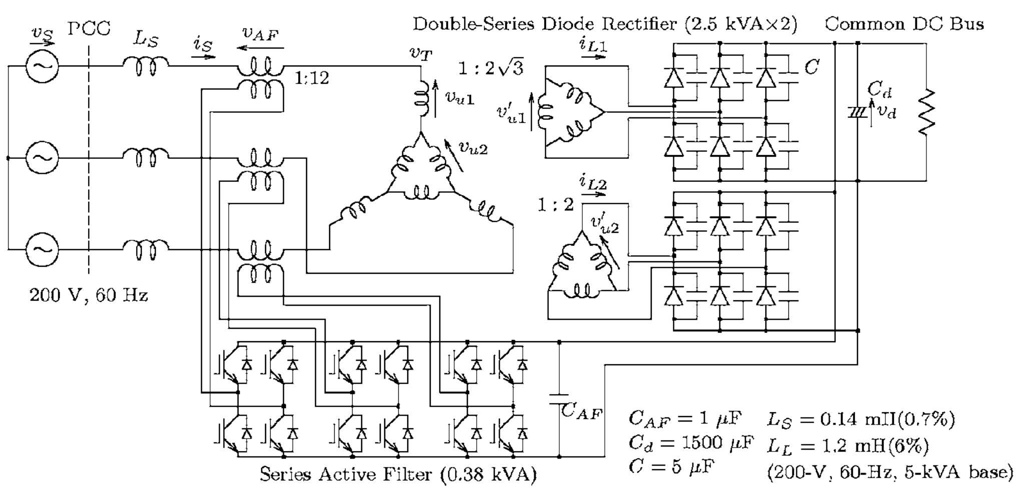


FIGURE 17.41 The harmonic current-free AC/DC power conversion system.

The primary windings of the Y-Δ and Δ-Δ connected transformers are connected in "series" with each other, so that the combination of the three-phase transformers and two three-phase diode rectifiers forms the "double-series" diode rectifier, which is characterized as a three-phase twelve-pulse rectifier. The DC terminals of the diode rectifier and the active filter form a common DC bus equipped with an electrolytic capacitor. This results not only in eliminating any electrolytic capacitor from the active filter, but also in reducing current ripples flowing into the electrolytic capacitor across the common DC bus.

Connecting only a commutation capacitor C in parallel with each diode plays an essential role in reducing the required peak voltage rating of the series active filter.

Operating Principle

Figure 17.42 shows an equivalent circuit for the power conversion system on a per-phase basis. The series active filter is represented as an AC voltage source v_{AF}, and the double-series diode rectifier as the series connection of a leakage inductor L_L of the transformers with an AC voltage source v_L. The reason for providing the AC voltage source to the equivalent model of the diode rectifier is that the electrolytic capacitor C_d is directly connected to the DC terminal of the diode rectifier, as shown in Fig. 17.41.

The active filter is controlled in such a way as to present zero impedance for the fundamental frequency and to act as a resistor with high resistance of K [Ω] for harmonic frequencies. The AC voltage of the active filter, which is applied to a power line through the matching transformer, is given by

$$v_{AF}{}^* = K \cdot i_{Sh} \tag{17.53}$$

where i_{Sh} is a supply harmonic current drawn from the utility. Note that v_{AF} and i_{Sh} are instantaneous values. Figure 17.43 shows an equivalent circuit with respect to current and voltage harmonics in Fig. 17.42. Referring to Fig. 17.43 enables derivation of the following basic equations:

$$I_{Sh} = \frac{V_{Sh} - V_{Lh}}{Z_S + Z_L + K} \tag{17.54}$$

$$V_{AF} = \frac{K}{Z_S + Z_L + K}(V_{Sh} - V_{Lh}) \tag{17.55}$$

where V_{AF} is equal to the harmonic voltage appearing across the resistor K in Fig. 17.43.

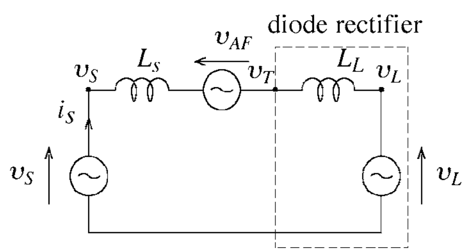


FIGURE 17.42 Single-phase equivalent circuit.

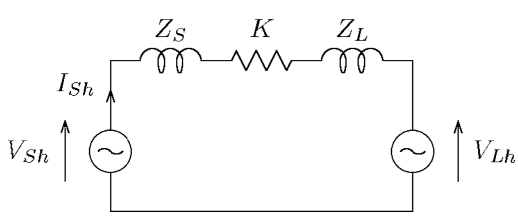


FIGURE 17.43 Single-phase equivalent circuit with respect to harmonics.

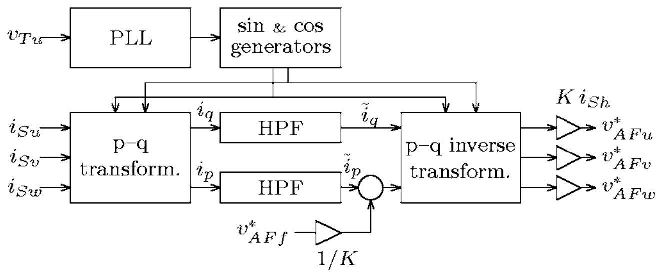


FIGURE 17.44 Control circuit for the series active filter.

If $K \gg Z_s + Z_L$, Eqs. (17.54) and (17.55) are changed into the following simple equations.

$$I_{Sh} \approx 0 \tag{17.56}$$

$$V_{AF} \approx V_{Sh} - V_{Lh} \tag{17.57}$$

Equation (17.56) implies that an almost purely sinusoidal current is drawn from the utility. As a result, each diode in the diode rectifier continues conducting during a half cycle. Eqution (17.57) suggests that the harmonic voltage V_{Lh}, which is produced by the diode rectifier, appears at the primary terminals of the transformers in Fig. 17.41, although it does not appear upstream of the active filter or at the utility–consumer point of common coupling (PCC).

Control Circuit

Figure 17.44 shows a block diagram of a control circuit based on hybrid analog/digital hardware. The concept of the Akagi–Nabae theory [Akagi, 1983; 1984] is applied to the control circuit implementation. The *p-q* transformation circuit executes the following calculation to convert the three-phase supply current i_{Sv}, i_{Sv}, and i_{Sw} into the instantaneous active current i_p and the instantaneous reactive current i_q.

$$\begin{bmatrix} i_p \\ i_q \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} \cos\omega t & \sin\omega t \\ -\sin\omega t & \cos\omega t \end{bmatrix} \cdot \begin{bmatrix} 1 & -1\ 2 & -1/2 \\ 0 & \sqrt{3}/2 & -\sqrt{3}/2 \end{bmatrix} \begin{bmatrix} i_{Su} \\ i_{Sv} \\ i_{Sw} \end{bmatrix} \tag{17.58}$$

The fundamental components in i_{Su}, i_{Sv}, and i_{Sw} correspond to DC components in i_p and i_q, and harmonic components to AC components. Two first-order high-pass-filters (HPFs) with the same cut-off frequency of 10 Hz as each other extract the AC components $\tilde{i}_p$ and $\tilde{i}_q$ from i_p and i_q, respectively. Then, the *p-q* transformation/inverse transformation of the extracted AC components produces the following supply harmonic currents:

$$\begin{bmatrix} i_{Shu} \\ i_{Shv} \\ i_{Shw} \end{bmatrix} = \sqrt{\frac{2}{3}} \begin{bmatrix} 1 & 0 \\ -1\ 2 & \sqrt{3}/2 \\ -1/2 & -\sqrt{3}/2 \end{bmatrix} \cdot \begin{bmatrix} \cos\omega t & -\sin\omega t \\ \sin\omega t & \cos\omega t \end{bmatrix} \begin{bmatrix} \tilde{i}_p \\ \tilde{i}_q \end{bmatrix} \tag{17.59}$$

Each harmonic current is amplified by a gain of K, and then it is applied to the gate control circuit of the active filter as a voltage reference v^*_{AF} in order to regulate the common DC bus voltage, v^*_{AFf} is divided by the gain of K, and then it is added to $\tilde{i}_p$.

The PLL (phase locked loop) circuit produces phase information ωt which is a 12-bit digital signal of 60×2^{12} samples per second. Digital signals, sin ωt and cos ωt, are generated from the phase information, and then they are applied to the p-q (inverse) transformation circuits. Multifunction in the transformation circuits is achieved by means of eight multiplying D/A converters. Each voltage reference, v_{AF}^* is compared with two repetitive triangular waveforms of 10 kHz in order to generate the gate signals for the IGBTs. The two triangular waveforms have the same frequency, but one has polarity opposite to the other, so that the equivalent switching frequency of each inverter is 20 kHz, which is twice as high as that of the triangular waveforms.

Experimental Results

In the following experiment, the control gain of the active filter, K, is set to 27 Ω, which is equal to 3.3 p.u. on a 3ϕ 200-V, 15-A, 60-Hz basis. Equation (17.54) suggests that the higher the control gain, the better the performance of the active filter. An extremely high gain, however, may make the control system unstable, and thereby a trade-off between performance and stability exists in determining an optimal control gain. A constant load resistor is connected to the common DC bus, as shown in Fig. 17.41.

Figures 17.45 and 17.46 show experimental waveforms, where a 5-μF commutation capacitor is connected in parallel with each diode used for the double-series diode rectifier. Table 17.4 shows the THD of i_S and the ratio of each harmonic current with respect to the fundamental current contained in i_S. Before starting the active filter, the supply 11th and 13th harmonic currents in Fig. 17.45 are slightly magnified due to resonance between the commutation capacitors C and the AC line and leakage inductors, L_S and L_L. Nonnegligible amounts of 3rd, 5th, and 7th harmonic currents, which are so-called "non-characteristic current harmonics" for the three-phase twelve-pulse diode rectifier, are drawn from the utility.

Figure 17.46 shows experimental waveforms where the peak voltage of the series active filter is imposed on a limitation of ±12 V inside the control circuit based on hybrid analog/digital hardware. Note that

TABLE 17.4 Supply Current THD and Harmonics Expressed as the Harmonic-to-Fundamental Current Ratio [%], Where Commutation Capacitors of 5 μF are Connected

	THD	3rd	5th	7th	11th	13th
Before (Fig. 17.46)	16.8	5.4	2.5	2.2	12.3	9.5
After (Fig. 17.47)	1.6	0.7	0.2	0.4	0.8	1.0

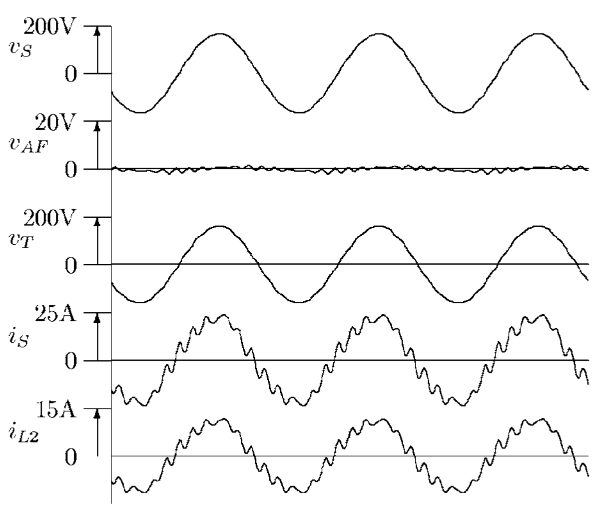


FIGURE 17.45 Experimental waveforms before starting the series active filter.

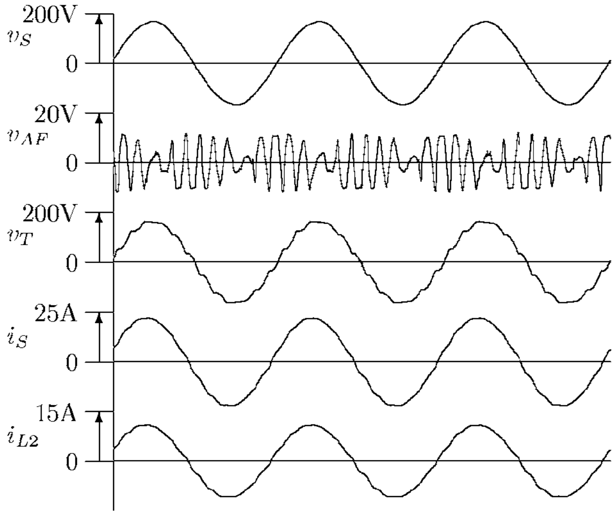


FIGURE 17.46 Experimental waveforms after starting the series active filter.

the limitation of ±12 V to the peak voltage is equivalent to the use of three single-phase matching transformers with turn ratios of 1:20 under the common DC-link voltage of 240 V. After starting the active filter, a sinusoidal current with a leading power factor of 0.96 is drawn because the active filter acts as a high resistor of 27 Ω, having the capability of compensating for both voltage harmonics V_{Sh} and V_{Lh}, as well as of damping the resonance. As shown in Fig. 17.46, the waveforms of i_S and v_T are not affected by the voltage limitation, although the peak voltage v_{AF} frequently reaches the saturation or limitation voltage of ±12 V.

The required peak voltage and current rating of the series active filter in Fig. 17.46 is given by

$$3 \times 12^{\mathrm{V}} / \sqrt{2} \times 15^{\mathrm{A}} = 0.38 \text{ kVA} \tag{17.60}$$

which is only 7.6% of the kVA-rating of the diode rectifier.

The harmonic current-free AC-to-DC power conversion system has both practical and economical advantages. Hence, it is expected to be used as a utility interface with large industrial inverter-based loads such as multiple adjustable speed drives and uninterruptible power supplies in the range of 1 to 10 MW.

Practical Applications of Active Filters for Power Conditioning

Present Status and Future Trends

Shunt active filters have been put into practical applications mainly for harmonic compensation, with or without reactive-power compensation. Table 17.5 shows ratings and application examples of shunt active filters classified by compensation objectives.

Applications of shunt active filters are expanding, not only into industry and electric power utilities but also into office buildings, hospitals, water supply utilities, and rolling stock. At present, voltage-fed PWM inverters using IGBT modules are usually employed as the power circuits of active filters in a range of 10 kVA to 2 MVA, and DC capacitors are used as the energy storage components.

Since a combined system of a series active filter and a shunt passive filter was proposed in 1988 [Peng et al., 1990], much research has been done on hybrid active filters and their practical applications [Bhattacharya et al., 1997; Aredes et al., 1998]. The reason is that hybrid active filters are attractive from both practical and economical points of view, in particular, for high-power applications. A hybrid active filter for harmonic damping has been installed at the Yamanashi test line for high-speed magnetically

TABLE 17.5 Shunt Active Filters on Commercial Base in Japan

Objective	Rating	Switching Devices	Applications
Harmonic compensation with or without reactive/negative-sequence current compensation	10 kVA ~ 2 MVA	IGBTs	Diode/thyristor rectifiers and cycloconverters for industrial loads
Voltage flicker compensation	5 MVA ~ 50 MVA	GTO thyristors	Arc furnaces
Voltage regulation	40 MVA ~ 60 MVA	GTO thyristors	Shinkansen (Japanese "bullet" trains)

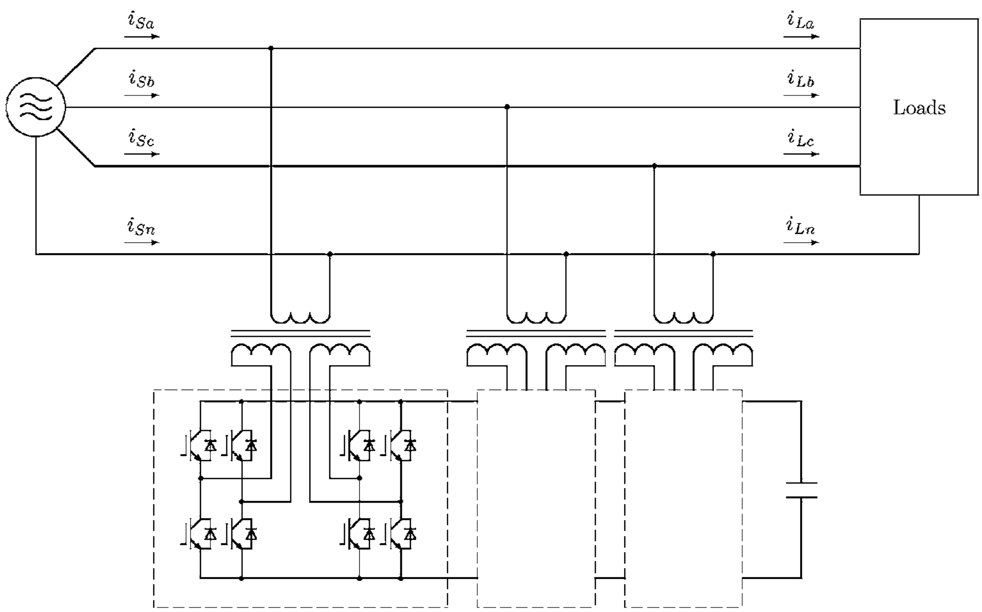


FIGURE 17.47 Shunt active filter for three-phase four-wire system.

levitated trains [Kawaguchi et al., 1997]. The hybrid filter consists of a combination of a 5-MVA series active filter and a 25-MVA shunt passive filter. The series active filter makes a great contribution to damping of harmonic resonance between the supply inductor and the shunt passive filter.

Shunt Active Filters for Three-Phase Four-Wire Systems

Figure 17.47 depicts the system configuration of a shunt active filter for a three-phase four-wire system. The 300-kVA active filter developed by Meidensha has been installed in a broadcasting station [Yoshida et al., 1998]. Electronic equipment for broadcasting requires single-phase 100-V AC power supply in Japan, and therefore the phase-neutral rms voltage is 100 V in Fig. 17.47. A single-phase diode rectifier is used as an AC-to-DC power converter in an electronic device for broadcasting. The single-phase diode rectifier generates an amount of third-harmonic current that flows back to the supply through the neutral line. Unfortunately, the third-harmonic currents injected from all of the diode rectifiers are in phase, thus contributing to a large amount of third-harmonic current flowing in the neutral line. The current harmonics, which mainly contain the 3rd, 5th, and 7th harmonic frequency components, may cause voltage harmonics at the secondary of a distribution transformer. The induced harmonic voltage may produce a serious effect on other harmonic-sensitive devices connected at the secondary of the transformer.

Figure 17.48 shows actually measured current waveform in Fig. 17.47. The load currents, i_{La}, i_{Lb}, and i_{Lc}, and the neutral current flowing on the load side, i_{Ln}, are distorted waveforms including a large amount of harmonic current, while the supply currents, i_{Sa}, i_{Sb}, and i_{Sc}, and the neutral current flowing on the supply side, i_{Sn}, are almost sinusoidal waveforms with the help of the active filter.

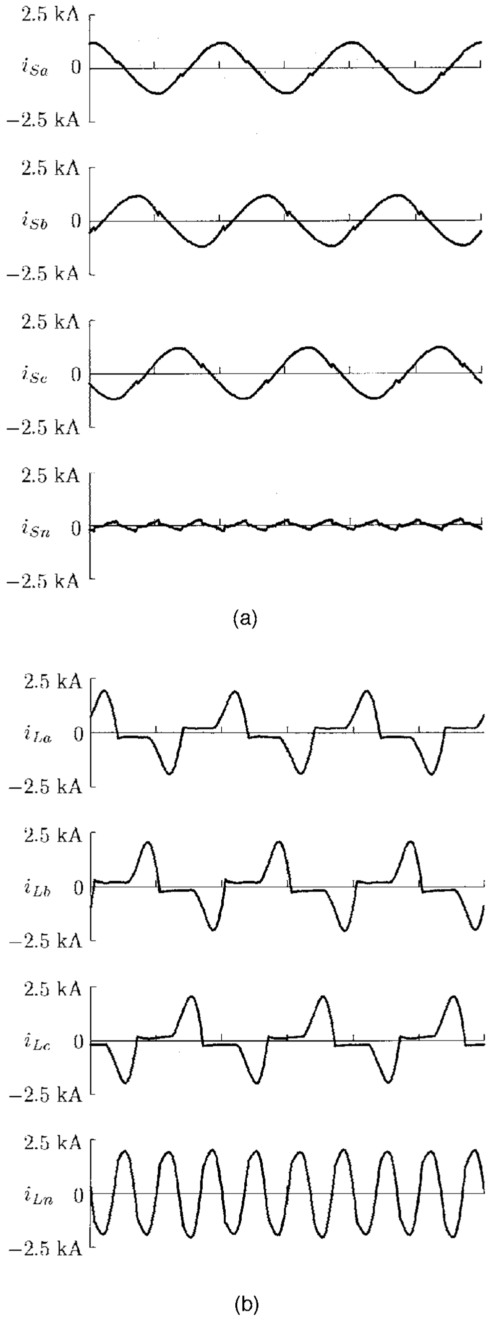


FIGURE 17.48 Actual current waveforms: (a) supply currents; (b) load currents.

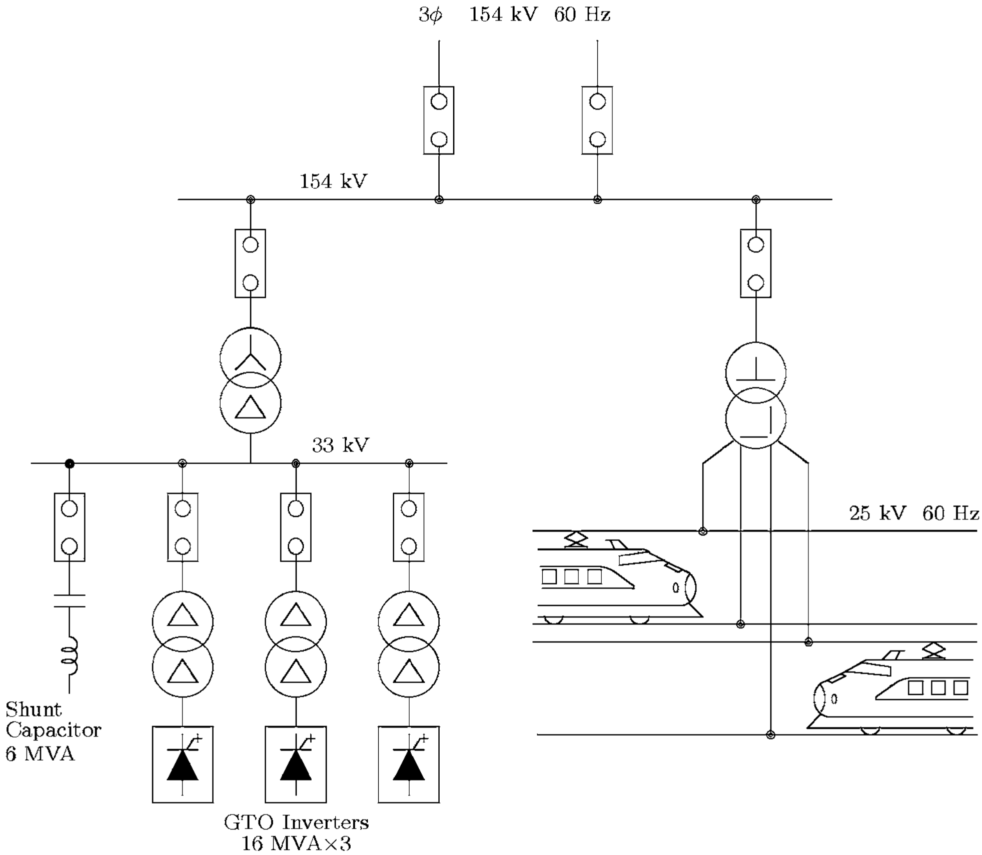
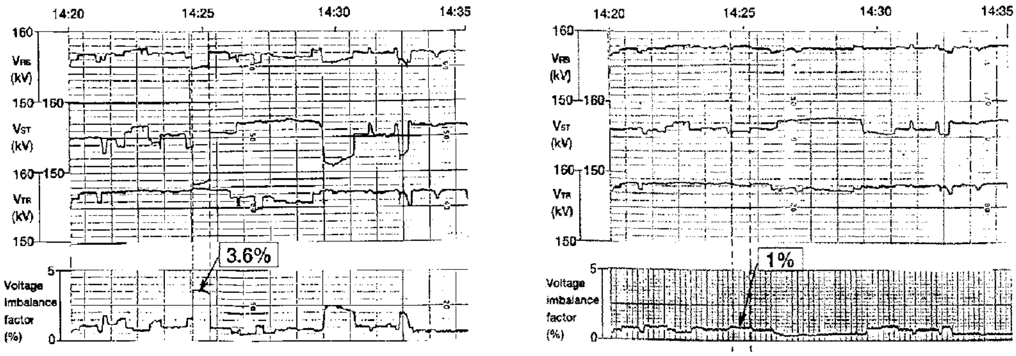


FIGURE 17.49 The 48-MVA shunt active filter installed in the Shintakatsuki substation.

The 48-MVA Shunt Active Filter for Compensation of Voltage Impact Drop, Variation, and Imbalance

Figure 17.49 shows a power system delivering electric power to the Japanese "bullet trains" on the Tokaido Shinkansen. Three shunt active filters for compensation of fluctuating reactive current/negative-sequence current have been installed in the Shintakatsuki substation by the Central Japan Railway Company [Iizuka et al., 1995]. The shunt active filters, manufactured by Toshiba, consist of voltage-fed PWM inverters using GTO thyristors, each of which is rated at 16 MVA. A high-speed train with maximum output power of 12 MW draws unbalanced varying active and reactive power from the Scott transformer, the primary of which is connected to the 154-kV utility grid. More than twenty high-speed trains pass per hour during the daytime. This causes voltage impact drop, variation, and imbalance at the terminals of the 154-kV utility system, accompanied by a serious deterioration in the power quality of other consumers connected to the same power system. The purpose of the shunt active filters with a total rating of 48 MVA is to compensate for voltage impact drop, voltage variation, and imbalance at the terminals of the 154-kV power system, and to improve the power quality. The concept of the instantaneous power theory in the time-domain has been applied to the control strategy for the shunt active filter.

Figure 17.50 shows voltage waveforms on the 154-kV bus and the voltage imbalance factor before and after compensation, measured at 14:20–14:30 on July 27, 1994. The shunt active filters are effective not only in compensating for the voltage impact drop and variation, but also in reducing the voltage imbalance factor from 3.6 to 1%. Here, the voltage imbalance factor is the ratio of the negative to positive sequence component in the three-phase voltages on the 154-kV bus. At present, several active filters in a range of 40 MVA to 60 MVA have been installed in substations along the Tokaido Shinkansen [Takeda et al., 1995].



(a) (b)

FIGURE 17.50 Installation effect: (a) before compensation; (b) after compensation.

Acknowledgment

The author would like to thank Meidensha Corporation and Toshiba Corporation for providing helpful and valuable information of the 300-kVA active filter and the 48-MVA active filter.

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17.5 Unity Power Factor Rectification

Rajapandian Ayyanar and Amit Kumar Jain

The proliferation of power electronic converters with front-end rectifiers has resulted in numerous problems for the utility distribution network. The currents drawn by the rectifier systems are nonsinusoidal and have large harmonic components, which interfere with other loads connected to the utility. Phase displacement of fundamental current and voltage requires the source and distribution equipment to handle reactive power and therefore higher rms currents for a given real output power. This section introduces the problems associated with the rectifier systems and discusses briefly the standards that are being enforced to limit the harmonic content in the line currents to acceptable levels. Three approaches—passive filters, active current-shaping techniques, and active filters—are introduced. Of these, the active current-shaping techniques for both the single-phase and three-phase applications are discussed in detail.

Diode Bridge and Phase-Controlled Rectifiers

In a majority of power electronic applications, for example, in switch-mode power supplies (SMPS), the utility voltage is first converted to an unregulated DC voltage using a single-phase or three-phase diode bridge rectifier. In a typical SMPS, this DC-link voltage is then converted to the desired voltage levels with isolation, using high-frequency DC-DC converters. In adjustable-speed drives, the DC-link voltage is converted to either a variable magnitude DC voltage as in DC drives, or a variable-frequency, variable-magnitude voltage suitable for AC motors. To minimize the ripple in the DC-link voltage, large capacitors are used as shown in Fig. 17.51.

The currents drawn by these diode bridge rectifiers from the utility, as shown in Fig. 17.51a and b, are not sinusoidal. The DC-link capacitor is charged to a value close to the peak of the utility voltage, and draws current only when the utility voltage is near its peak value. Hence, the current drawn from the utility is discontinuous and rich in harmonics. Table 17.6 gives the harmonic spectrum of the current drawn by a typical single-phase diode bridge rectifier with a capacitive filter [1]. As seen, it has a large

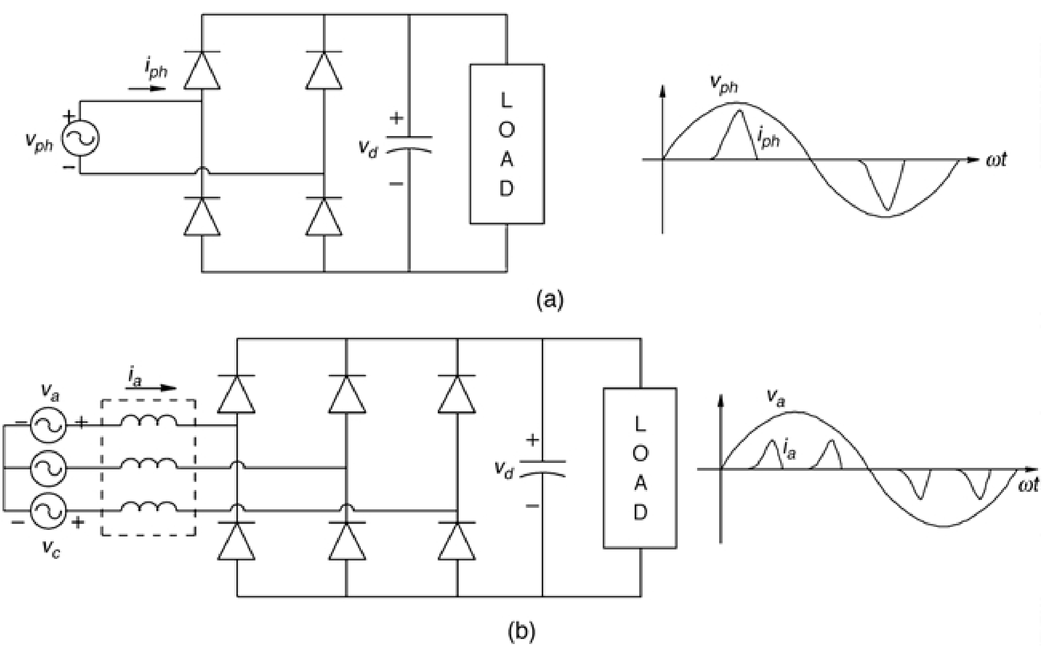

FIGURE 17.51 Diode bridge rectifiers: (a) single phase, (b) three phase.

TABLE 17.6 Typical Harmonic Spectrum of Line Current Drawn by Single-Phase Diode Bridge Rectifier

h	3	5	7	9	11	13	15	17
$\frac{I_h}{I_1}$ %	73.2	36.6	8.1	5.7	4.1	2.9	0.8	0.4

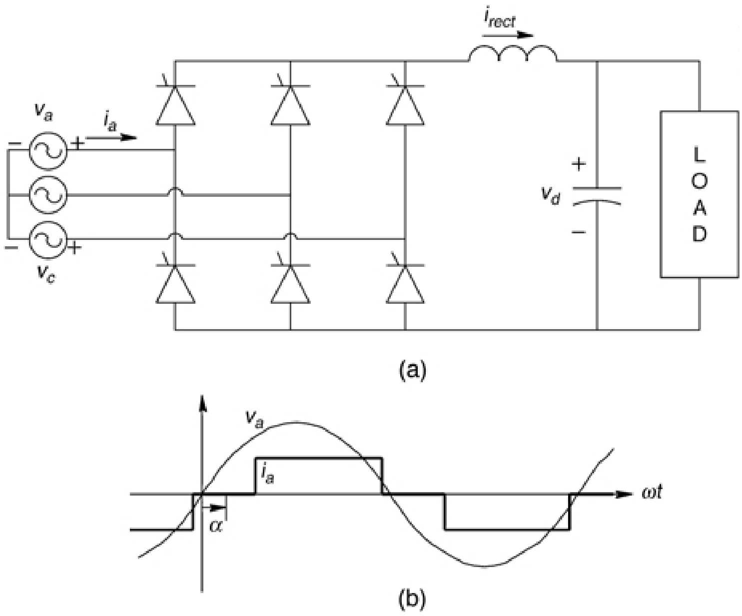

FIGURE 17.52 Three-phase phase-controlled rectifiers.

third harmonic component. For a given output power, the rms value of the line current for the case shown in Table 17.6 is about 30% higher than that of a sinusoidal current at unity power factor.

Figure 17.52a shows the schematic diagram of a three-phase phase-controlled rectifier. Unlike diode bridge rectifiers, the DC-link voltage here can be regulated by controlling the firing angle of the thyristors with respect to the utility voltage. Such phase-controlled three-phase rectifiers with inductive filter are commonly used in high-power applications. Figure 17.52b shows an input phase voltage and the corresponding phase current for a three-phase, phase-controlled rectifier, neglecting the source inductance. The DC side inductor is assumed to be large enough so that the inductor current is pure DC. As seen from Fig. 17.52b, the current drawn from the utility is a quasi-square wave, with its fundamental component displaced from the mains voltage by the firing angle α. The dominant harmonics in the input currents are the 5th and the 7th [1].

Standards for Limiting Harmonic Currents

Various national and international agencies have specified standards and guidelines or recommended practices for the maximum allowable harmonic currents that can be injected into the utility. The standards widely used are those formulated by the International Electrotechnical Commission (IEC), ANSI/IEEE Standards, and the VDE series of German standards [2, 3].

IEC 555 standards, released in 1982, underwent several amendments and eventually became part of the IEC 1000 family, which deals with electromagnetic compliance [4]. IEC 1000-3-2 sets the limits for harmonic emission—both absolute values as well as per-watt magnitude—for equipment with input current less than or equal to 16 A per phase. The standard refers to four different classes of equipment:

TABLE 17.7 IEC 1000-3-2 Limits for Class D Equipment

Harmonic Order n	Maximum Permissible Harmonic Current per Watt mA/W	Maximum Permissible Harmonic Current A
3	3.4	2.30
5	1.9	1.14
7	1.0	0.77
9	0.5	0.40
11	0.35	0.33
$13 \le n \le 39$	$\frac{3.85}{n}$	Refer Class A

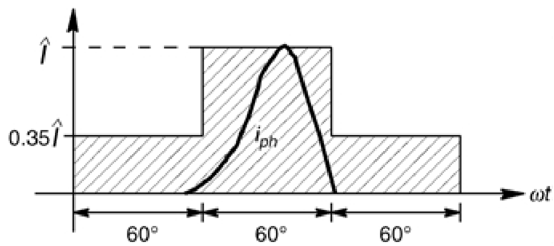

FIGURE 17.53 Envelop of current waveshapes covered in Class D.

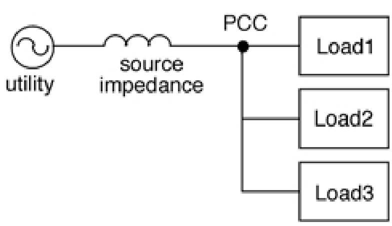

FIGURE 17.54 Point of common coupling.

Class A refers to balanced three-phase equipment, Class B refers to portable tools, Class C to lighting equipment, and Class D deals with equipment that draws currents having a special nonsinusoidal waveshape lying under the envelope shown in Fig. 17.53 with an active input power under 600 W. The diode bridge rectifiers discussed earlier belong to this class of equipment. Table 17.7 lists the 1000-3-2 limits for Class D equipment with maximum current less than or equal to 16 A[4]. IEC 1000-3-3 deals with limits on voltage flicker in low-voltage supply systems with current less than 16 A. IEC 1000-3-4 sets limits for both individual equipment as well as for the whole system installation. IEC 1000-3-2 has been adopted as EN 61000-3-2, and hence any equipment sold in Europe must comply with the standard.

Whereas the IEC standards limit the harmonic emission produced by the equipment, the IEEE standard 519 limits the harmonic currents and voltages at the point of common coupling (PCC) [5]. PCC refers to the point at which all the loads at a particular location are connected to the utility as shown in Fig. 17.54. Table 17.1 lists the harmonic current limits recommended by this standard for power systems below 69 kV. The short circuit ratio (SCR) shown in column 1 is defined as the ratio of maximum short-circuit current, I_s, to the maximum load current, I_L. As seen, stricter limits are applied to low SCR, since low SCR, implying high source impedance, results in higher voltage distortion for a given harmonic current magnitude.

Definitions of Some Common Terms

This section discusses definitions of some of the commonly used terms related to power factor. For a single-phase system, *power factor* is defined as

$$\text{Power factor} = \frac{\text{Average power}}{\text{Apparent power}} = \frac{\frac{1}{T}\int_0^T v_s(t)i_s(t)}{V_{\text{rms}}I_{\text{rms}}} \tag{17.61}$$

where $v_s(t)$ and $i_s(t)$ are the instantaneous phase voltage and current, and V_{rms} and I_{rms} are the corresponding rms quantities.

If the voltage is sinusoidal,

$$\text{Power factor} = \frac{V_{\text{rms}}I_{1,\,\text{rms}}}{V_{\text{rms}}I_{\text{rms}}}\cos\theta = \frac{I_{1,\,\text{rms}}}{I_{\text{rms}}}\cos\theta \tag{17.62}$$

In Eq. (17.62), cos θ is called the *displacement power factor*, and θ is the phase angle between the voltage and the fundamental component of the current. The ratio $I_{1,\text{rms}}/I_{\text{rms}}$ is called the *distortion power factor* and it accounts for the additional losses due to harmonics. Hence, the power factor with a sinusoidal voltage waveform and an arbitrary current waveform is equal to the product of the distortion power factor and the displacement power factor.

The harmonic content in the current waveform is usually quantified by the *total harmonic distortion* (THD) defined as:

$$\text{THD} = \frac{\sqrt{I_{\text{rms}}^2 - I_{1,\,\text{rms}}^2}}{I_{1,\,\text{rms}}} \tag{17.63}$$

Passive Solutions

To achieve compliance with the above-mentioned harmonic standards, a simple but less effective approach is to insert a passive filter between the harmonic generating sources, namely, the diode bridge rectifier or the phase-controlled thyristor rectifier and the utility. The other approach, referred to as the active solution, is to use a high-frequency power electronic converter as a pre-regulator that will shape the current drawn from the utility to be sinusoidal. The passive filters are reliable, less complex, and usually less expensive compared with the active solutions, but are bulky, heavy, and cannot be used for applications that have to meet stringent harmonic distortion standards. Another drawback of passive filters as compared with active solutions is their inability to regulate the DC-link voltage.

Passive filters can broadly be classified into series filters, shunt filters, and a hybrid combination of the two. Series filters introduce an impedance in series with the utility to reduce harmonic currents. Shunt filters provide a low-impedance path for the harmonic currents generated by the rectifiers so that they are not reflected in the current drawn from the utility. Figure 17.55 shows a simple series-type filter, where a suitably designed inductor is inserted between the diode bridge and the mains. The waveform of the corresponding line current is shown in Fig. 17.55b. Comparing this current to that without a series filter as shown in Fig. 17.51a, it can be seen that the rms value for a given power is lower in the present case. Tuned filters to attenuate selected current harmonics, usually 3rd and 5th, are also employed to improve the power factor. The main drawback of such an approach is that if the input voltage has even a small component at the tuned frequency, the filter can resonate with the mains and draw large harmonic currents.

Active Solutions

An active power factor correction (PFC) circuit refers to the use of a power electronic converter, switching at high frequencies, to shape the input current to be sinusoidal and in phase with the input utility voltage. Using active PFC techniques, it is possible to achieve a power factor greater than 0.99 and a THD less

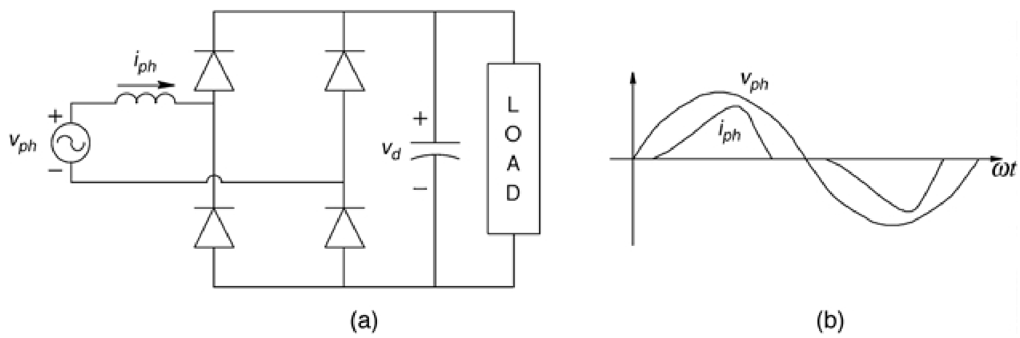

FIGURE 17.55 Passive solution for single-phase rectifier.

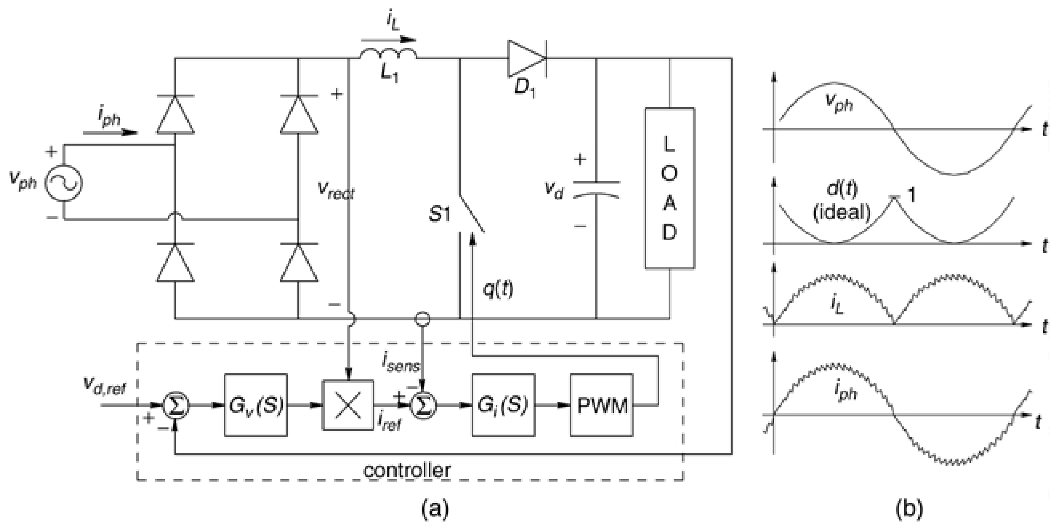

FIGURE 17.56 Single-phase boost power factor correction.

than 5%. Compared with the passive solutions, they are less bulky and can easily meet the standards on harmonic distortion. Another major advantage of the active approach is that the DC-link voltage can be regulated (although usually with slow dynamics), even when the input voltage varies over a wide range. Commonly used PFC circuits for single-phase and three-phase power factor correction applications are discussed below.

Single-Phase Active PFC Circuits

Boost Converter–Based PFC

The boost converter with a front-end diode bridge rectifier (without the large input filter capacitor) is perhaps the most popular single-phase PFC topology at present [6, 7]. Figure 17.56 shows the schematic diagram of the single-phase boost PFC circuit along with the controllers. The duty ratio of S_1 is modulated such that the waveshape of the inductor current closely tracks the rectified sinusoidal voltage obtained at the output of the diode bridge rectifier. The inductor L_1 is designed such that the operation of the converter is always in continuous conduction mode (CCM).

The controller involves a slow outer voltage loop and an inner fast current loop. The outer voltage loop produces a signal that depends on the error between the actual and the desired output voltages. The reference to the inner current loop is derived from the rectified sinusoidal input voltage, suitably scaled by multiplying it with the output signal from the voltage loop. For the current loop, either peak current mode control or average current control can be used. In average current mode control, the current controller amplifies the error between the actual inductor current and the current reference. This amplified error signal is then compared with a fixed frequency ramp to obtain the gate control signal $q(t)$ for the switch. In peak current mode control, the actual inductor current $i_L(t)$ is compared with the current reference, $i_{\text{ref}}(t)$. When $i_L(t)$ just exceeds the reference the switch is turned off, and is turned on again at start of the next switching cycle by a fixed-frequency clock signal. It is also possible to control the current using a hysteresis controller.

As in DC-DC converter applications, the output voltage of the boost converter needs to be above the maximum input voltage. Hence, for single-phase PFC applications using the boost converter, the DC-link voltage has to be higher than the peak of the maximum input utility voltage. For a nominal utility voltage of 110 V rms, the DC-link voltage needs to be close to 200 V DC. In a single-phase rectifier with unity power factor, the input power fluctuates at twice the fundamental frequency (for example, 120 Hz), whereas the output power is constant (within the 120-Hz cycle). Hence, energy storage elements are needed to account for the difference between the instantaneous values of the input power and the output power. The output capacitor of the boost converter thus needs to be large in single-phase PFC applications.

Figure 17.56b shows the waveforms corresponding to the input utility voltage, the idealized instantaneous switch duty ratio, the inductor current, and the current drawn from the utility. Usually small, low-pass filters are used to reduce the high-frequency switching component in the input current, and to meet EMI standards.

The design of the feedback controller for both the outer voltage loop and the inner current loop is discussed in detail in Refs. 6 and 7. The reference signal to the current loop, which is mainly a rectified sinusoidal voltage, contains a large second-order harmonic term, and substantial higher-order harmonics. Therefore, to obtain low THD for the input current, the current loop should be able to track the higher-order harmonics. Thus, the current loop should have a high bandwidth. This is especially important if the input frequency is higher, as, for example, 400 Hz in avionic applications.

The current through the filter capacitor has a large second harmonic component apart from the switching frequency component. This is reflected in the DC-link voltage as large second harmonic ripple voltage. The outer voltage loop therefore should be designed for low bandwidth (much less than 120 Hz for utility applications) so that the above-mentioned ripple voltage does not distort the reference signal applied to the current loop. Since voltage control is required to have a low bandwidth it does not compensate for 120 Hz and the dynamic response for step-load changes is very poor. Usually the boost PFC is used as a pre-regulator with main emphasis on drawing sinusoidal currents at unity power factor. The pre-regulator is followed by a second-stage DC-DC converter, which provides electrical isolation as well as good dynamic regulation of load voltage at any desired voltage levels.

Other CCM Topologies

Similar to the boost converter, several other basic DC-DC converter topologies can be used with a diode bridge rectifier for PFC applications. SEPIC and Cúk converters operating in CCM have the same advantage as the boost converter in that they draw continuous current from the utility with small, high-frequency ripple [8]. An additional advantage of SEPIC or Cúk converter over the boost converter for PFC applications is that the output voltage can be less than the peak of the input voltage. The trade-off is in the complex controller design.

For single-phase PFC applications, buck-derived converters are not practical because such a converter will not be able to draw any current from the input when the input voltage is instantaneously less than the output voltage. Figure 17.57 shows an interesting combination of a buck and a boost converter suitable for PFC applications that require the output voltage to be less than the peak input voltage [9]. When the main voltage is instantaneously less than the desired output voltage, S_1 is fully on, and the duty ratio of S_2 is controlled in response to the current reference. The operation is now in boost mode. When the input voltage is higher than the output, S_2 is fully off and the duty ratio of S_1 is controlled, and the operation is now in the buck mode. In the buck mode the input current is discontinuous, but its average value tracks the input voltage.

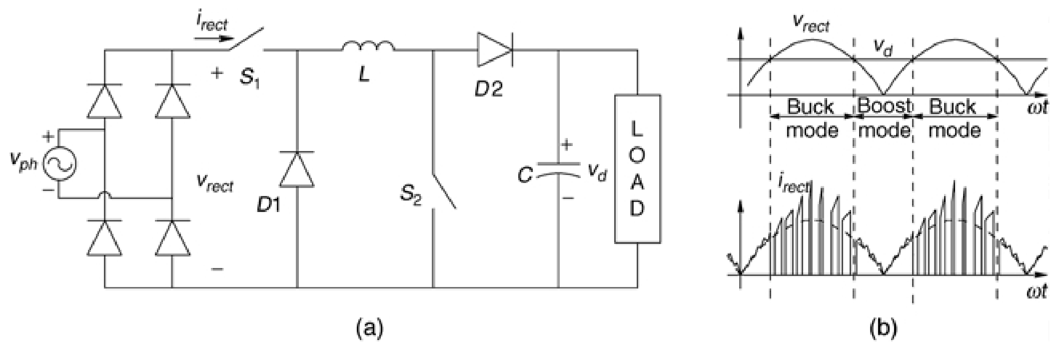

FIGURE 17.57 Buck–boost single-phase power factor correction.

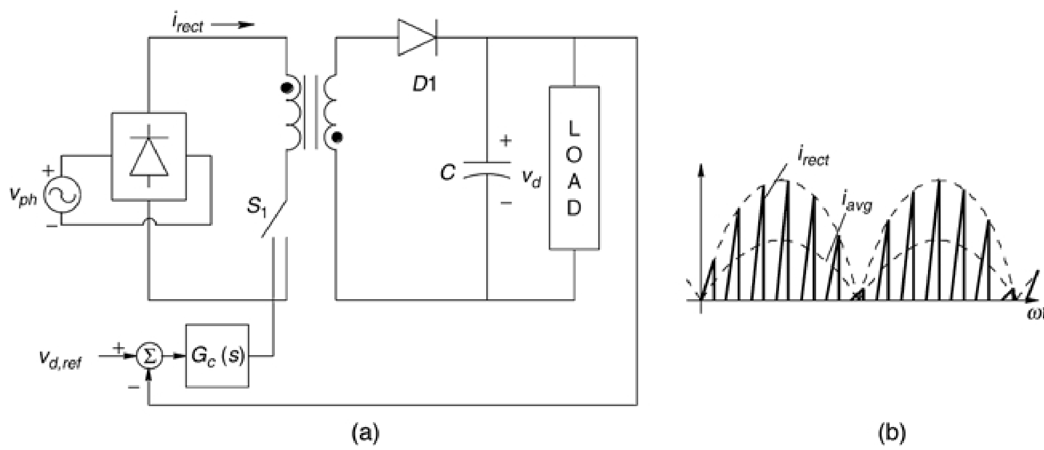

FIGURE 17.58 Single-phase PFC: flyback operation in discontinuous mode.

PFC Converters Utilizing DCM Operation

All the converters discussed so far for single-phase PFC applications operate in CCM. In these converters the duty ratio of the switch is modulated by a current loop to shape the input current within a line-frequency cycle, and a slow outer voltage loop is used to regulate the output DC voltage. There is a family of converters that, while operating in discontinuous conduction mode (DCM), inherently behave as a resistive load and thus draw sinusoidal current at unity power factor (UPF). Flyback, SEPIC, and Cúk converters belong to this family of converters. Their operation involves maintaining a constant duty ratio within a line-frequency cycle, without the need for a current loop. A slow output voltage loop adjusts the duty ratio to regulate the DC-link voltage.

Figure 17.58a shows the schematic diagram of a flyback power factor correction circuit operating in DCM. The design details, including the output voltage control loop, are discussed in detail in Ref. 10. The expression for the input current, $i_g(t)$ (average over a switching period), is as shown in Eq. (17.64) and is derived in Ref. 10.

$$i_g(t) = \frac{V_{pk}}{R_e}|\sin(\omega t)| \tag{17.64}$$

where $R_e = 2n^2L/D^2T_s$, ω is the fundamental frequency in rad/s, n is the turns ratio of the flyback transformer, D is the duty ratio of the switch, and T_s is the switching period. The corresponding input current is shown in Fig. 17.58b, and as seen its average follows the input main voltage. Boost converters

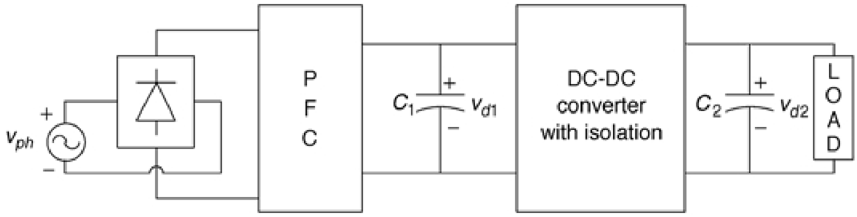

FIGURE 17.59 Two-stage solution for single-phase AC-to-DC converters with PFC.

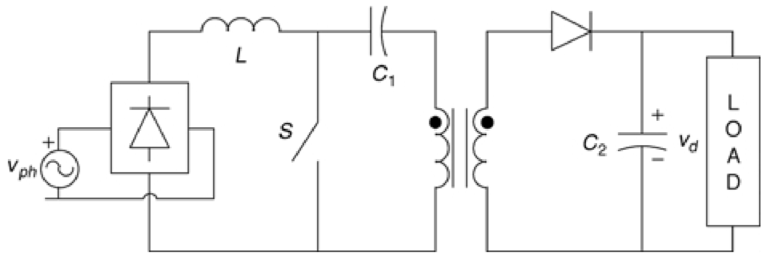

FIGURE 17.60 Single-phase AC-DC converter with PFC: isolated SEPIC.

can also be operated in DCM mode, but unlike the flyback converter it introduces harmonic distortion. THD in a boost DCM converter reduces with increasing output voltage.

The operation of Cúk and SEPIC converters in DCM mode as natural sinusoidal rectifiers is discussed in Refs. 11 and 12, respectively. The advantage of Cúk or SEPIC converters in DCM compared with the flyback or boost is that the input current does not return to zero in each switching period. DCM mode for Cúk and SEPIC is defined as operating condition when the sum of the two inductor currents becomes discontinuous. Also, the two inductors can be coupled, resulting in significant reduction in the high-frequency ripple component in the input current.

Single-Stage Solutions

The PFC circuits described above are used as pre-regulators with a UPF feature. They are usually followed by a DC-DC converter stage as shown in Fig. 17.59, which provides isolation and achieves good dynamic regulation for the load voltage. The voltage levels required by the modern loads are very low—for example, 2 V and below for some integrated circuits. The DC-link voltage at the output of the PFC stage is around 200 V. Conversion of this high DC voltage directly to 2 V as required by the loads is highly inefficient, mainly because of the large turns ratios required. There are two options. The first is to use a three-stage approach with a PFC pre-regulator, a DC-DC converter that converts the 200-V DC-link voltage to isolated distribution level voltage (around 28 V, for example), and another non-isolated DC-DC converter to provide the final well regulated DC voltage at 2 V. The second option is to use what are referred to as single-stage PFC circuits, which directly convert the main voltage to an isolated distribution-level DC voltage (28 V). This is then followed by another DC-DC converter to obtain the well-regulated low DC voltage. The second approach is more efficient and also results in higher overall power density.

Although the boost converter in CCM is a popular topology for non-isolated PFC solutions, it is not very attractive for the single-stage applications, mainly because it is difficult to achieve isolation using the boost configuration in CCM mode. Many DCM solutions have been suggested [13] to realize single-stage PFC circuits, and are useful for lower power applications. For applications that have strict restrictions on the high-frequency component in the input current, the isolated version of SEPIC converter as shown in Fig. 17.60 and operating in CCM is an attractive solution [8]. By suitable choice of turns ratio any desired output voltage can be obtained. Similar to DC-DC converters, interleaving of two or more converters can result in ripple reduction [14].

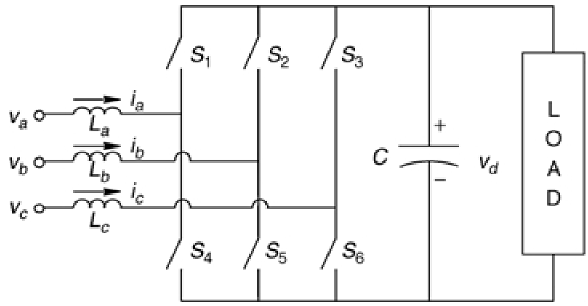

FIGURE 17.61 Three-phase six-switch boost PFC.

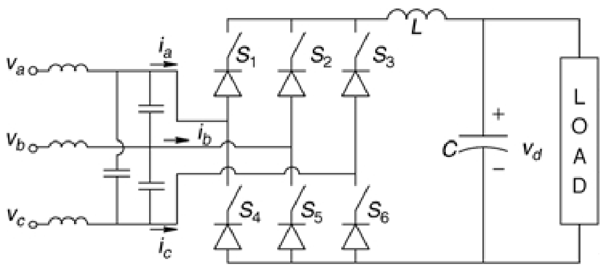

FIGURE 17.62 Three-phase six-switch buck PFC.

Three-Phase PFC Circuits

At higher power levels, for example, above a few kilowatts, three-phase rectifiers are normally preferred. As in the case of single-phase rectifiers, passive PFC solutions are acceptable for less-demanding applications. The passive solutions are typically bulkier. For applications with stringent restrictions on the allowable harmonic distortion active current shaping using high-frequency power electronic circuits is necessary.

Six-Switch PWM Converters

Perhaps the most effective, but expensive, solution is the six-switch boost PWM (pulse-width-modulated) converter [15] shown in Fig. 17.61. The design of the converter ensures that the operation is always in CCM. The control involves an outer voltage loop, which gives the reference to the inner current loop. A preferred method of current control for the inner loop is to use space-vector modulation (SVM) techniques, with digital implementation. The output voltage needs to be higher than the peak of the input line-to-line voltage [16]. A major advantage of the six-switch PWM boost rectifier is its bidirectional power flow capability. However, because of the higher cost and complex control requirement, its use is limited to higher power applications and applications requiring bidirectional power flow. Unlike single-phase converters, in all the three-phase PFC circuits, the power drawn from the input is not pulsating (at two or six times the fundamental frequency). Hence, the output capacitor can be designed based on just the hold-up time requirement. Figure 17.62 shows the schematic diagram of a six-switch buck-derived PWM PFC circuit. The advantage of the buck-derived converter is that the output voltage can be lower than the peak value of the input voltage. However, the input current is discontinuous, and hence large filters are needed to reduce the switching frequency component in the input current.

Single-Switch Three-Phase PFCs

The simplest active three-phase PFC circuit is the single-switch DCM boost converter shown in Fig. 17.63a [17]. The duty ratio S is kept constant within a line-frequency cycle, and is controlled by a slow, outer voltage loop. Because of DCM operation, all the three-phase currents are zero at the beginning of the

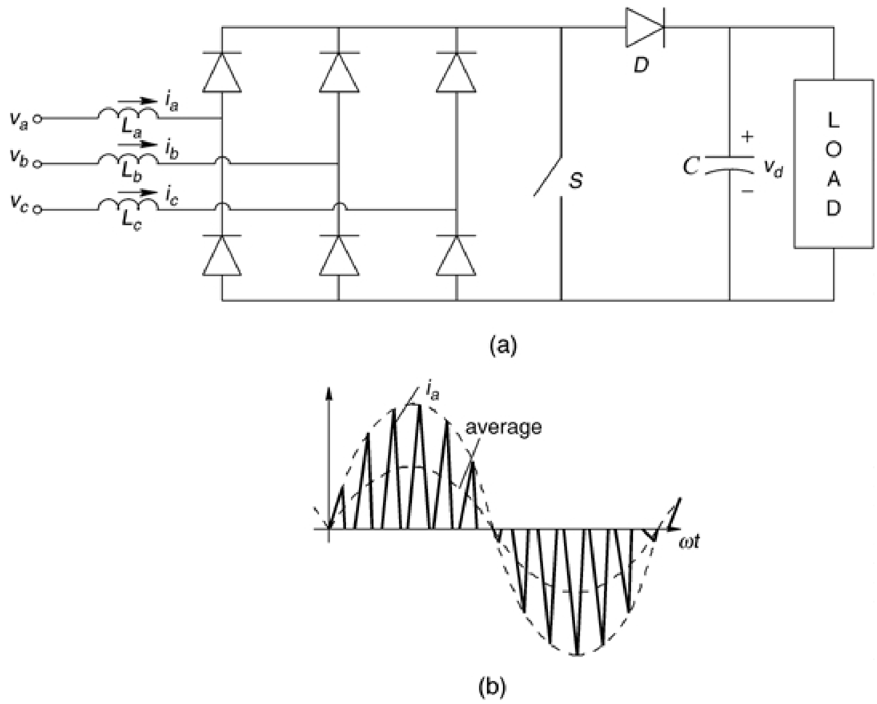

FIGURE 17.63 Single-switch three-phase DCM boost PFC.

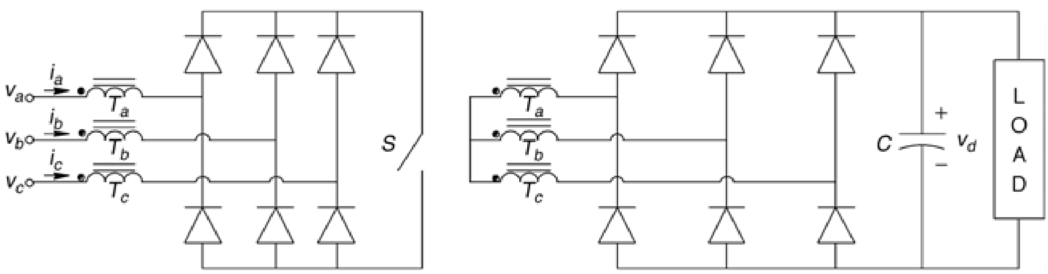

FIGURE 17.64 Single-switch three-phase DCM flyback PFC.

switching period, and change linearly at a rate proportional to the corresponding instantaneous phase voltages when S is turned on. Therefore, the input current peaks are proportional to the input voltages. However, during the off period, this linear relation does not hold, and therefore the current is distorted. Harmonic analysis shows that the input current exhibits a large fifth-order harmonic. The THD depends on the ratio of the output voltage to the peak of the input main voltage and reduces with increasing output voltage. The output voltage must be very high to achieve low THD. High peak currents and high filter rating at the input are other major disadvantages. Techniques to reduce the distortion with fairly low output voltage levels are discussed in Ref. 18.

Figure 17.64 shows a scheme where the three inductors of the earlier scheme (Fig. 17.63) are replaced by three flyback transformers [19]. The main advantage of this scheme is that during the on-interval the primary current is proportional to the input voltage and during the off-interval it is zero. Hence, on an average, the input current accurately follows the input voltage resulting in lower THD as compared with the boost scheme. Also, the turns ratio can be chosen appropriately to obtain any desired output voltage. Hence, this scheme is an attractive option especially at low power levels, around 1 kW.

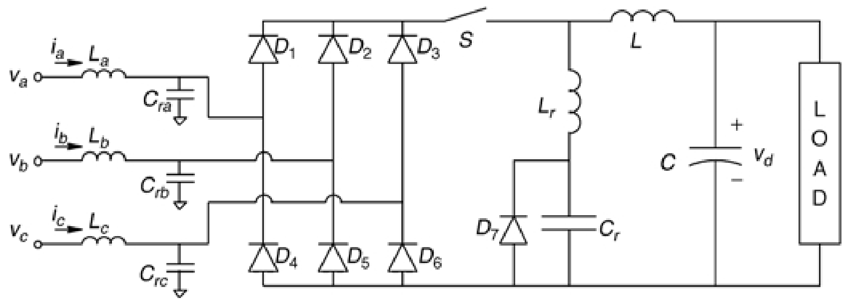

FIGURE 17.65 Single-switch three-phase buck-derived multiresonant ZSC PFC.

Figure 17.65 shows a representative topology from a class of multiresonant single-switch three-phase PFC circuits [20]. Low THD can be achieved with the additional advantage of soft-switching for the main switch. However, the VA rating of the switch is high for a given output power, and hence its application is limited to lower power levels. Control of output voltage is achieved by varying the switching frequency. The operation is designed to be in discontinuous voltage mode (DVM) with respect to the input capacitors, and the average voltage across the capacitors is proportional to the corresponding instantaneous phase voltages.

Use of Single-Phase PFCs for Three-Phase Applications

An attractive solution to achieve high-performance three-phase rectification is to use three single-phase PFC circuits connected in star or delta at the input side and in parallel at the output side. When non-isolated single-phase converters are used in such a scheme, precaution should be taken to avoid interaction among the three phases. Ref. 21 discusses in detail the operation of three nonisolated boost PFC circuits with extra circuitry to avoid phase interactions. The advantages include low overall switch ratings and low passive component ratings compared with other solutions of similar performance. Also, well-known single-phase PFC methods, as well as control ICs can be directly applied for the three-phase extension.

Figure 17.66 shows three isolated single-phase SEPIC converters operating in CCM, with delta connection at the input and parallel connection at the output [22]. Since the three single-phase PFCs are isolated, there is no problem of adverse interaction among the three phases, and hence it is an attractive single-stage solution.

Vienna Rectifier

A three-switch, three-level, three-phase PFC circuit referred to as Vienna rectifier [23] is shown in Fig. 17.67. The three switches are controlled usually by space-vector PWM such that the input currents accurately track the phase voltages. When a switch is turned on, the corresponding phase is connected to the midpoint *O*, resulting in an increase in the phase current. Turning off the switch results in the conduction of the associated diodes in the upper or lower bridge half, causing the phase currents to reduce. Because of the inclusion of the midpoint of the output voltage into the system function, the circuit has three-level characteristics. Hence, the high-frequency content in the mains current is reduced significantly as compared with that in the two-level converters. Another advantage is the relatively lower combined VA ratings for the controllable switches. The disadvantage is the slightly more complex control circuitry.

Buck-Derived Three-Phase PFC

Figure 17.68 shows the schematic diagram of two interleaved buck-derived three-phase PFC circuits with three switches each. The control is very simple, as it is based on conventional sinusoidal PWM, and retains good harmonic performance [24]. The output voltage can be lower than the peak of the input voltage. However, the input currents are discontinuous, making it necessary to use large low-pass filters at the input to attenuate the switching frequency content. As suggested in Fig. 17.68, two or more such

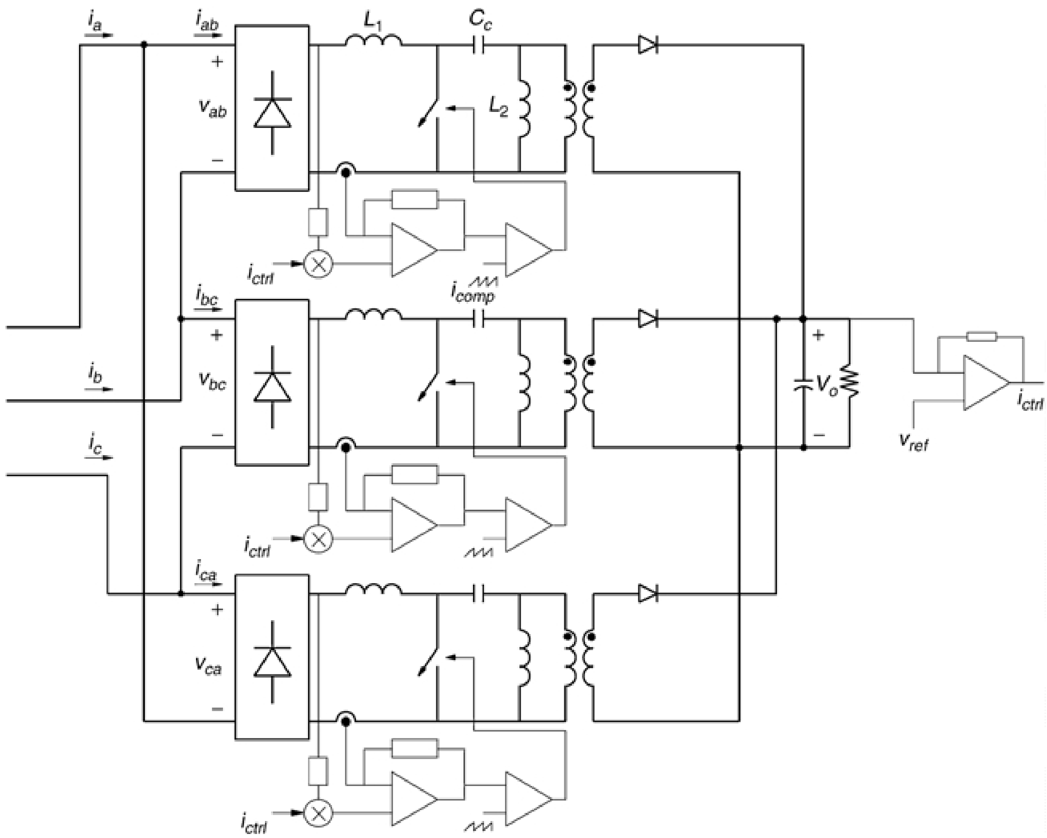

FIGURE 17.66 Three single-phase SEPIC converters delta-connected at input and paralled at the ouptput.

buck-derived PFC circuits can be interleaved, i.e., connected in parallel with their switching signals suitably sequenced. Interleaving leads to ripple cancellation, resulting in lower filter requirement.

Active Filters

Use of passive filters between the harmonic generating rectifier load and the utility is discussed briefly in the section on passive solutions. The main drawbacks mentioned were that passive filters are bulky, can result in over/undercompensation as the load varies, and can result in a low-impedance current path for the harmonic components present in the utility voltage. In this section, another approach to reduce the harmonic distortion, namely, the active filters, is discussed [25, 26]. Figure 17.69 shows a block diagram representation of the concept of active filters to eliminate the harmonic currents from entering the utility. The current drawn by the rectifier system (single-phase or three-phase) has a fundamental component i_{L_1} and a distortion component $i_{L_{\text{distortion1}}}$. The active filter senses the current drawn by the rectifier system and, by high-frequency current-mode control, delivers the distortion component $i_{L_{\text{distortion1}}}$. Therefore, the utility needs to deliver only the fundamental component of the current drawn by the rectifier system. Since the active filter needs to supply only the harmonic component of the current, no active power needs to be delivered to the load by the active filter. Hence, the active filter does not require a separate energy source. Figure 17.70 shows a typical implementation of an active filter for three-phase PFC applications. For high-power systems, where the use of active filter may be limited by the availability of high-voltage and high-current devices, hybrid filters, which are a combination of active filters and passive LC filters, are an effective approach to meet harmonic standards.

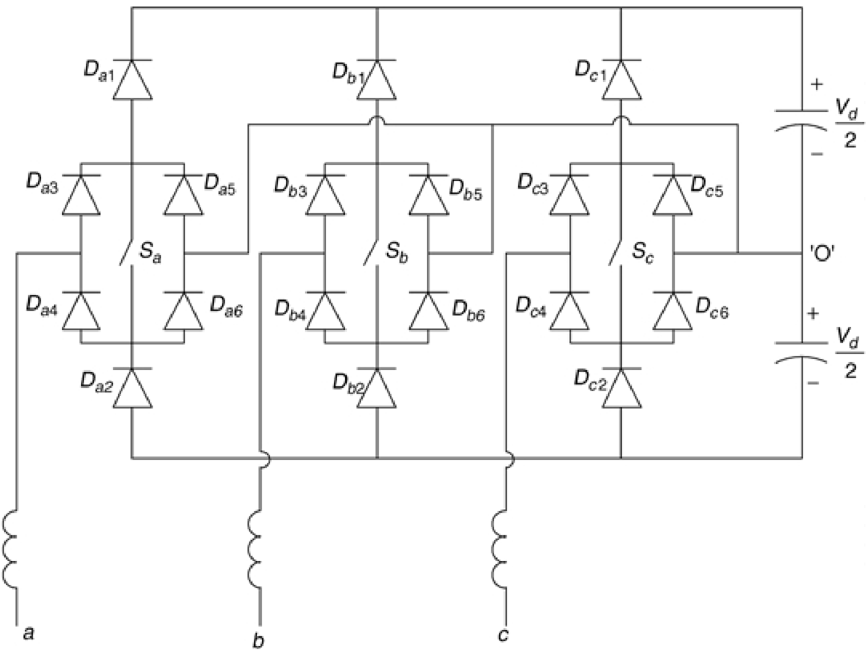

FIGURE 17.67 Vienna rectifier–I.

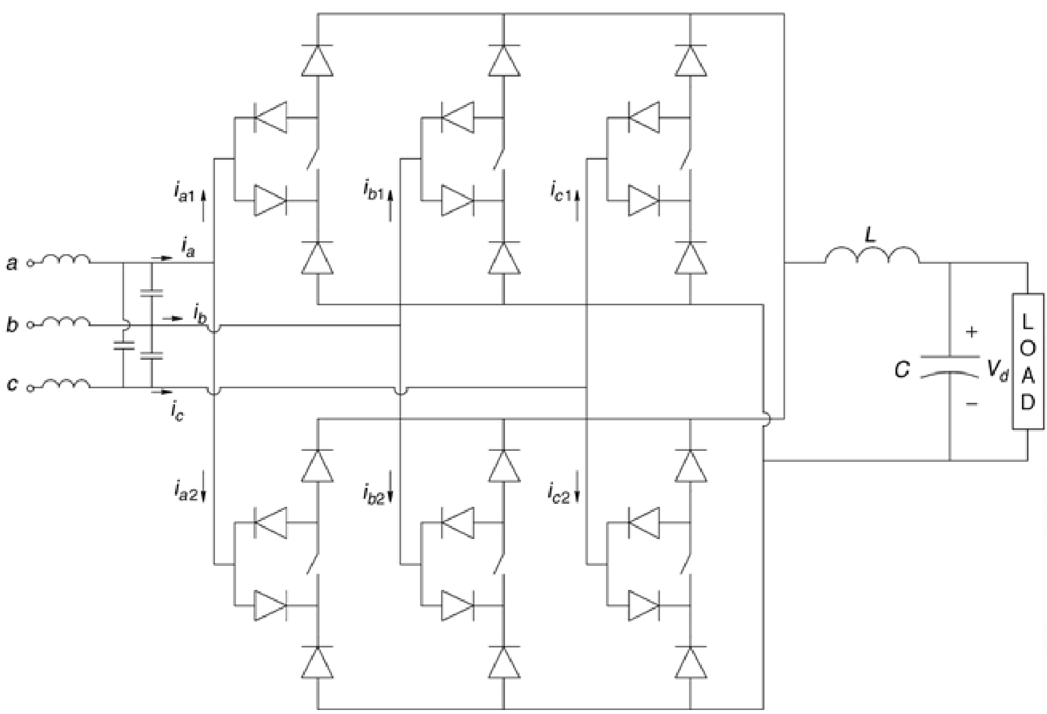

FIGURE 17.68 Three-phase interleaved PFC.

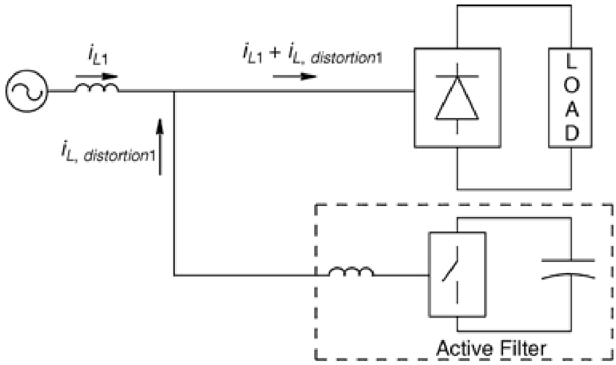

FIGURE 17.69 Active filter.

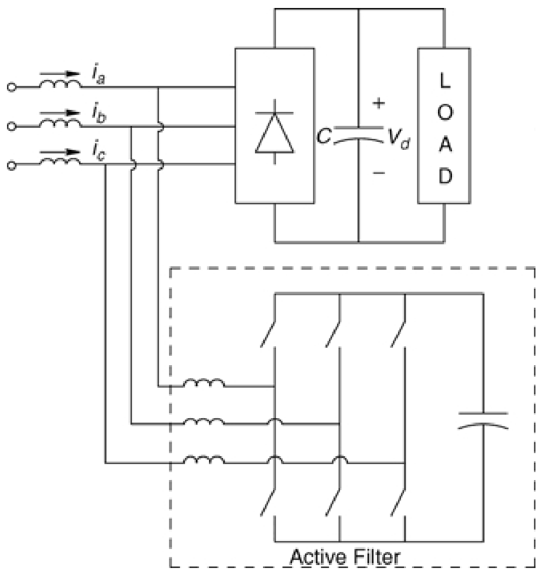

FIGURE 17.70 A three-phase active filter for power factor correction.

Summary

The input currents drawn by conventional diode bridge and phase-controlled thyristor rectifiers are far from sinusoidal and are rich in harmonics. The harmonic currents pollute the utility system and interfere with other loads connected to the system. Several standards that place restrictions on the amount of distortion in the line current were discussed briefly. Several techniques, such as passive filters, active current shaping using high-frequency power electronic converters, and the active filter approach were introduced. Various topologies were discussed for the active current-shaping technique, for both single-phase and three-phase applications. The relative merits of these topologies and the applications for which they are best suited were pointed out.

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