14. High-Voltage AC/DC Conversion, or Rectification

Vacuum tubes, or electron devices, are unidirectional conductors. Electron charge carriers will only go in one direction. Therefore, at some point in the realization of a microwave-tube transmitter a source of direct current (dc) will have to be provided, and often at very high voltages consisting of tens or even hundreds of kilovolts. The storage of electrical energy in large capacitor banks is also a unidirectional process, even though dc cannot flow through a capacitor. (Charge is stored in a capacitor by the flow of current into it in one direction and removed by a current out of it in the opposite direction. But these directions, once established, do not change.) Almost no microwave-tube transmitters of significant power output are directly powered from a primary source of dc, such as a battery, thermoelectric device, or dc generator. (Some very exotic, single-shot, explosive-energy electrical converters do find uses, though.) Many systems, however, are indirectly powered from such sources, but only after voltage-level conversion that involves the internal generation of alternating current by means of an inverter. Other transmitters are indirectly powered from external sources of alternating current (ac), such as commercial power grids or dedicated alternators. In all of these cases conversion from ac to dc, which is called rectification, is required before it can be used to power our microwave-tube transmitter electron beam (or beams).

14.1 Polyphase ac concepts

Every engineer knows something about ac-dc rectification, but a few still don't know much about polyphase rectification—or even polyphase ac circuits, for that matter. As power levels increase, polyphase circuits become more and more efficacious, and at the higher power levels already discussed they are all but mandatory.

Commercial power is generated and distributed in the form of three balanced phases comprising voltage vectors that are displaced from one another by 120°, as shown in Fig. 14-1. A number of advantages accrue from this arrangement.

- The current in the conductor connecting the neutral, or common point, of a wye-connected source and the common point of a balanced, wye-connected load has a vector sum that is zero.
- The vector sum of voltages around a properly phased, delta-connected source is also zero, allowing the delta to be "closed" without circulating current.
- The vector sum of the mechanical forces on the windings of a three-phase alternator is also zero, so that there are no unbalanced rotational forces.
- The mechanical forces on the rotor of a three-phase motor are purely rotational, which is most desirable.

As we will see, there are additional advantages to three-phase operation when it

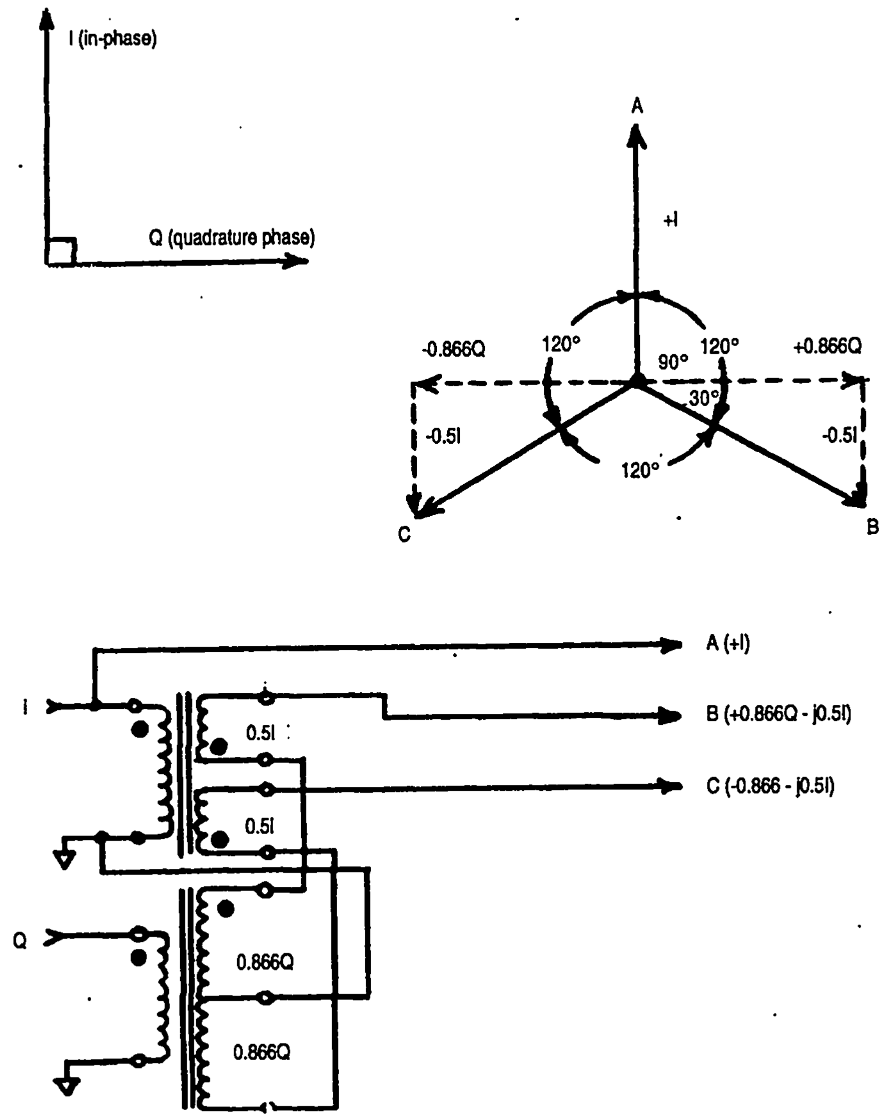

Figure 14-1. If you start with two phases, you can make the rest.

comes to rectification.

A balanced, rotating electrical machine is not the only source of a three-phase electrical system, however; nor is there anything magical about just three phases. To create a system having a theoretically unlimited number of discrete phases, all we need to start with is two. And it theoretically does not matter how infinitesimal the phase difference is between them. Even if the difference is a microradian, one vector can be resolved into components that are in-phase and quadrature-phase (I and Q) with respect to the other. Indeed, modern digital signal processing is based on transforming a repetitive waveform into its Fourier series of harmonically related components and evaluating two samples of each, the I and Q components.

Given two voltage vectors of the same amplitude that are in quadrature (which is sometimes called a bi-phase system) and two transformers with identical primary windings and secondary winding ratios and polarities as shown in Fig. 14-

1, a balanced three-phase system can be synthesized. Different transformer winding ratios will accomplish the same thing for different ratios of quadrature-phased vector amplitudes. It also should be obvious that there is no reason to be limited to just three phases. More transformers will produce more phases.

14.2 The three-phase, half-wave rectifier

Single-phase ac rectifiers require external hold-up for either load current (the inductor-input filter) or load voltage (the capacitor-input filter) if the load requires non-varying direct current. Without some form of energy storage between the rectifier and a resistive load, the closest that load voltage and current can ever come to non-varying dc is a succession of half-sine-wave segments, all of the same polarity, all with infinite peak-to-valley ratio. It is direct current, to be sure, but a long way from non-varying. Polyphase ac rectification overcomes this problem.

The simplest form of three-phase rectifier is the half-wave, or three-pulse rectifier, as shown in Fig. 14-2. The transformer secondary is wye-connected. A rectifier diode is placed in series with each phase, and the return from the load is to the neutral point. The rectifier shown is connected for positive-polarity dc output. Whichever phase—A, B, or C—is instantaneously more positive than the other two will be the conducting phase, and conduction will continue for 120°, or 1/3 cycle, with the peak voltage at the center of the conducting interval. The "natural" commutation point, or transfer point between the phases, occurs when

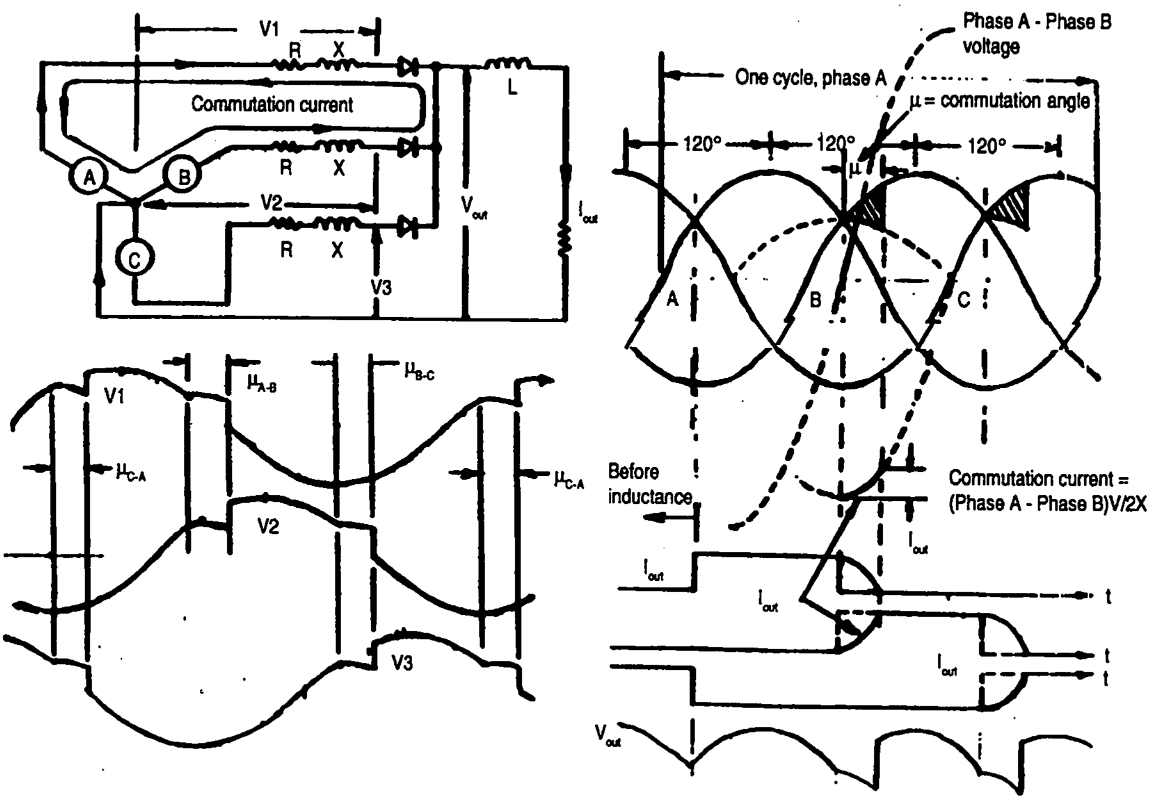

Figure 14-2. The effect of ac line reactance on rectifier current commutation.

the instantaneous voltage of the next sequential phase becomes more positive than the previously conducting phase. At this point for either zero per-phase source inductance or zero load current, the load voltage will simply follow that of the newly conducting phase. This pattern repeats itself throughout successive phase rotations. The load voltage, therefore, will be a series of sinusoid tops spanning each wave's -60° point to its +60° point. There will be three such tops per cycle, hence the term "three-pulse," or "three-bump," rectifier. If the peak voltage is unity, the instantaneous voltage at the commutation point is 0.5, so the peak-to-valley ratio is 2. The theoretical no-load (or no-source-inductance) Fourier series for the frequency spectrum of the load voltage is

$$\frac{3}{\pi}\times\sin\frac{\pi}{3}\left[1+\frac{2}{8}\cos 3\omega t-\frac{2}{35}\cos 6\omega t+\frac{2}{80}\cos 9\omega t K+\frac{2}{(3n)^2-1}\cos 3n\omega t\right].$$

The first term has a value of 0.907, which is the ratio of the zero-frequency, or average, "dc" component of output voltage to the peak value of input voltage, even though the valleys of the waveform are only 1/2 as great as the peaks. (Note that the comparable first term for single-phase, full-wave [or two-pulse] rectification is $2/\pi \times \sin(\pi/2)$, which is 0.636. This is the case, even though its order of rectification, or pulse number, is 2/3 that of the simplest three-phase rectifier.) The other terms in the series are the ripple components, the first of which, theoretically, is the third harmonic.

If, however, the three-phase source inductance is not zero—and it never is—and load current is not zero either, the output voltage will not simply follow the tops of the phase voltages. The three-phase source impedance will consist primarily of transformer winding resistance, *R*, and the reactance of the leakage inductance, *X*. The resistance is usually no more than 1/5 as great as the reactance and can be ignored as a contributor to total ac impedance. Assuming that inductor *L* in series with the load is sufficiently large, the output current will be non-varying dc. However, the inductance in series with each phase—inductance through which load current is also flowing—will resist the transfer, or commutation, of the load current to the incoming phase. It will develop whatever voltage across itself that is required to keep its diode forward biased and current going through it. Fortunately, there is a countervailing current called commutation current that opposes the inductor current, eventually forcing it to zero and permitting the commutation of load current to the next phase in line. The commutation current is driven by the vector sum of the outgoing and incoming phase voltages (shown as A and B in the Fig. 14-2) and limited by the reactances of the two phases in series. Heading positive, the driving voltage for commutation current passes through zero at the natural commutation point. Being limited by inductive reactance, however, the commutation current lags the voltage by 90°, hence it is passing through a negative peak at that time. During the time that the commutating current is driving the load current from one phase to the next, which is known as the commutation angle, the output voltage follows neither phase voltage but rather the average of the two. Therefore, the output average voltage is less positive than the incoming phase voltage but more positive than the outgoing phase voltage. At the end of the commutation event, current has transferred to the new phase, and the output voltage abruptly jumps up to coincide with the new phase voltage at that instant. The voltage follows it along until the next commutation point is reached, where the process is again repeated. The relative difficulty of the commutation and the length of the commutation angle are proportional to what is called the commutation factor, which is the product of load current and phase reactance divided by phase voltage.

14.3 The three-phase, full-wave (six-pulse) rectifier

The three-phase, half-wave (or three-pulse) rectifier described above is used primarily in very-low-voltage, high-current applications, where rectifier voltage drop is a primary consideration. Far more popular is the three-phase, full-wave (or six-pulse) rectifier, as shown in Fig. 14-3. Forward conduction of load current occurs for both positive and negative half-cycles of each phase. The no-load output voltage is a succession of 60° conduction intervals, which occur between the natural commutation points from -30° to +30°, with respect to each successive peak. There are six such conduction intervals per cycle, hence it is called "six-pulse," or "six-bump," rectification. The valleys are determined by multiplying the cosine of 30° by the voltage peaks, or 0.866 x V_{pk}, for a peak-to-valley ratio of 1.15.

The Fourier series describing the theoretical no-load spectrum of the load voltage is

$$\frac{6}{\pi}\times\sin\frac{\pi}{6}\left[1+\frac{2}{35}\cos 6\omega t-\frac{2}{143}\cos 12\omega t+\frac{2}{323}\cos 18\omega t K\ +\frac{2}{(6n)^2-1}\cos 6n\omega t\right],$$

where n is the harmonic number. Ignoring load current and commutation, whose effects on voltage and current waveforms are shown in Fig. 14-3, the no-load spectrum can be broken down into an average dc or zero-frequency term and a series of ripple-frequency components that diminish in amplitude as the harmonic number increases. The value of the first term is 0.995, which is the ratio of the average value of the zero-frequency term to the peak value of the alternating input voltage. The first theoretical ripple component is the sixth harmonic of the input frequency, and its amplitude is 2/35 of the dc term. Higher-order ripple components are multiples of the sixth harmonic and are diminished by the factor $2/[(6n)^2 -1]$ from the dc term as the harmonic number increases.

This waveform comes pretty close to what most engineers consider dc, even without additional filtering. In fact, the need for filtering at all is mostly to attenuate the second and fourth harmonic ripple components, which arise from voltage imbalance between the input phase-to-phase voltages. Theoretically—and sometimes in practice—this imbalance can be compensated for directly by separately adjusting phase voltages for greater equality. This procedure will null components that are not multiples of the sixth harmonic in the ripple spectrum. Many high-voltage power-supply vendors will automatically assume there is a three-phase line imbalance of some value—typically 5%—and design filtering to bring the second and fourth harmonic ripple components in line with the overall

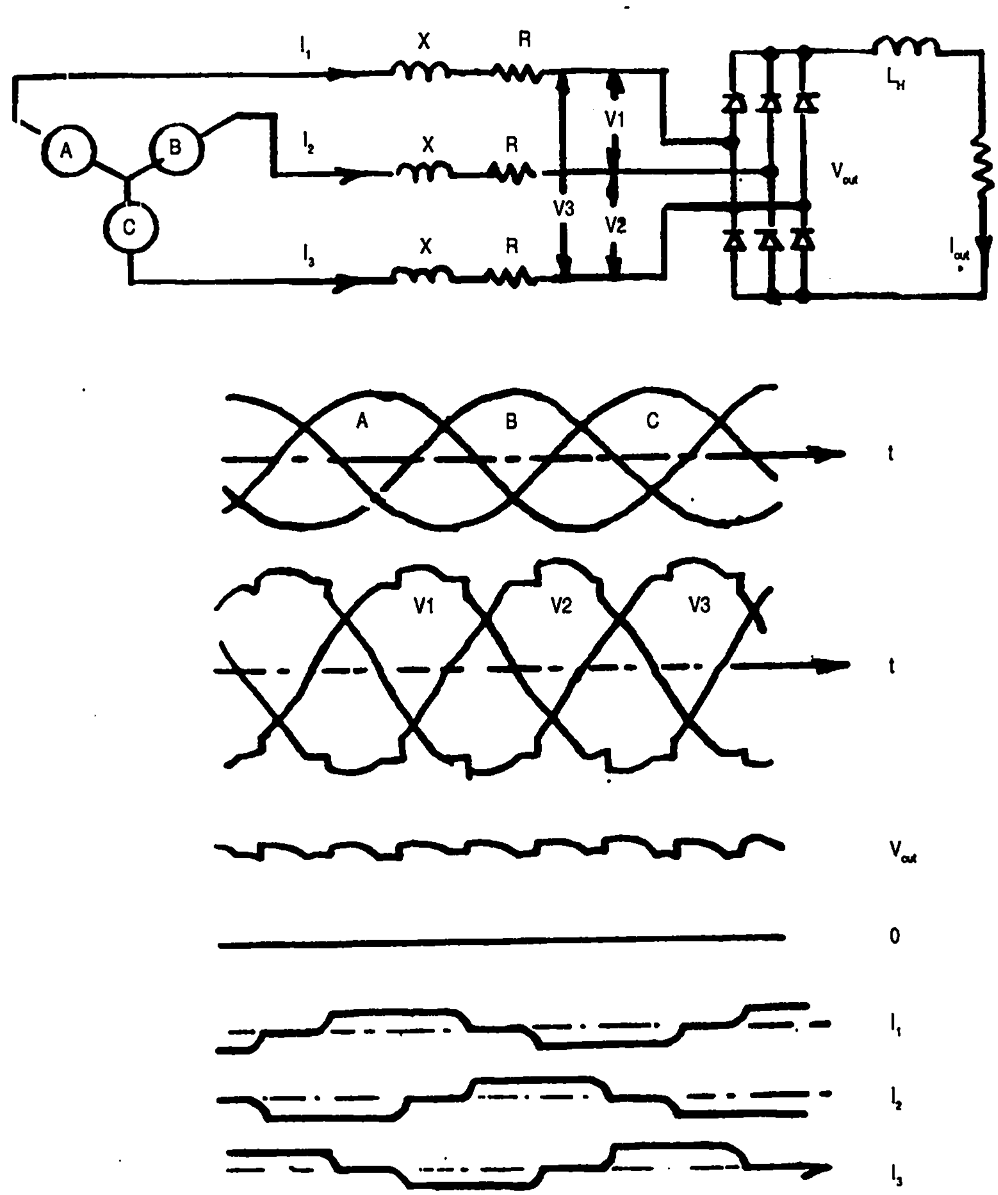

Figure 14-3. The three-phase, full-wave, or six-pulse rectifier

ripple requirement.

14.4 The six-phase, full-wave (12-pulse) rectifier

As popular as the six-pulse rectifier has become for high-power, high-voltage applications, the 12-pulse rectifier is rapidly supplanting it as the configuration of choice. Even though it requires two full-wave bridge rectifiers instead of one and two separate secondary windings to feed them, the total number of rectifier junctions is usually the same. This is because series-connected individual rectifier junctions are required to provide the total reverse-voltage capability and it does not matter whether they are connected as one bridge or two. The transformer core and coil combination is also much the same, except the secondary is divided two halves from a total volt-ampere perspective.

The six-phase input required for a 12-pulse rectifier can be derived in several ways. A "natural" way is by the use of a dual secondary, one winding of which is connected in delta, and the other in a wye. Not surprisingly this design is referred to as a "delta-wye" configuration and is shown in Fig. 14-4. A disadvantage of this arrangement is that the two secondaries are not identical, except in volt-ampere product. The legs of the delta winding must have √3 times the voltage of the wye-connected legs. To even things out, the wye-connected legs must handle √3 times the current. The wye windings have fewer turns but are of heavier gauge wire than the delta-connected windings. This makes things a bit more complicated for the winding shop (and therefore makes it easier to get something wrong). The "natural" aspect of the six-phase output is actually illustrated in the figure. Note that if the windings were drawn to a scale proportional to their voltages and aligned so that they were parallel to their corresponding primary winding, the vectors representing the resultant, or line-line, voltages of the wye-connected winding would be tilted by 30° with respect to the corresponding delta-connected output vectors. This 30° "interphase" displacement interleaves the outputs of the two six-pulse rectifiers in series, yielding a 12-pulse output with conduction angles of 30° for each output pulse, or "bump." The phase progression is shown in Fig. 14-5, where each phase—A, B, and C—has a "shadow" displaced by 30°—A′, B′, and C′.

Except for the effects of any three-phase line-voltage amplitude imbalance,

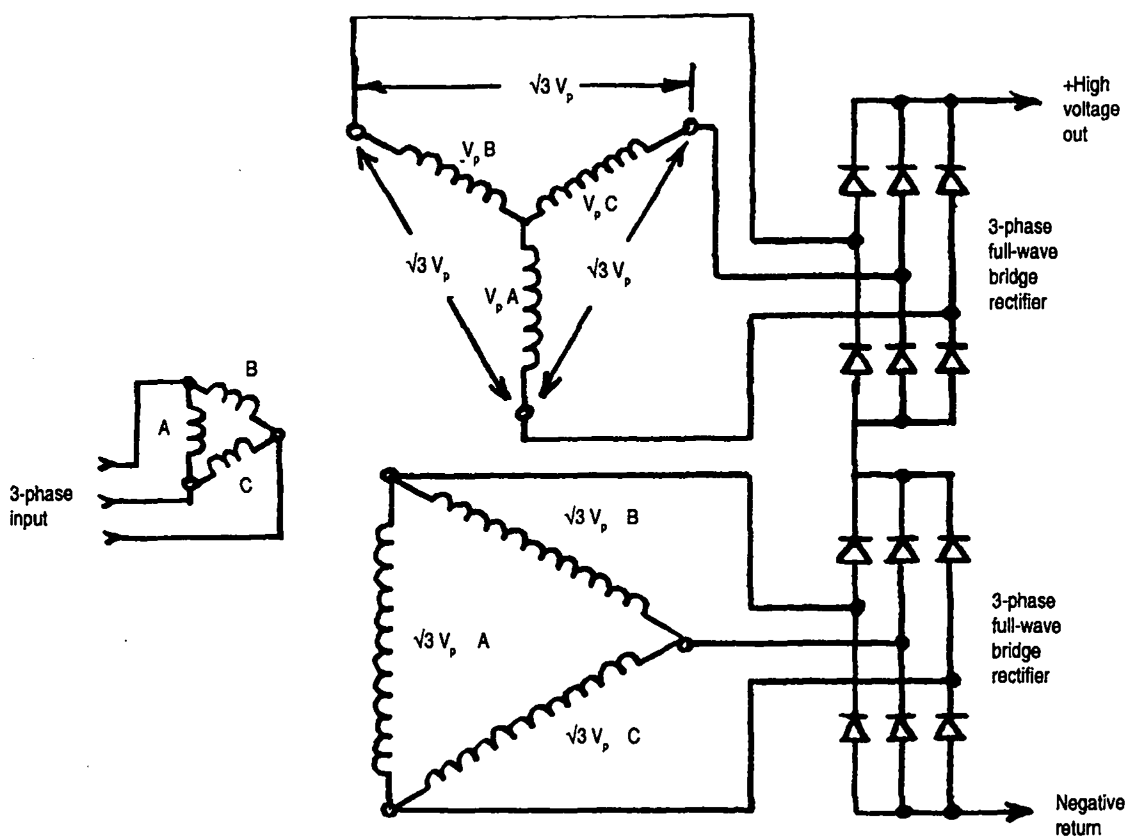

Figure 14-4. The "delta-wye," six-phase, full wave, or 12-pulse rectifier.

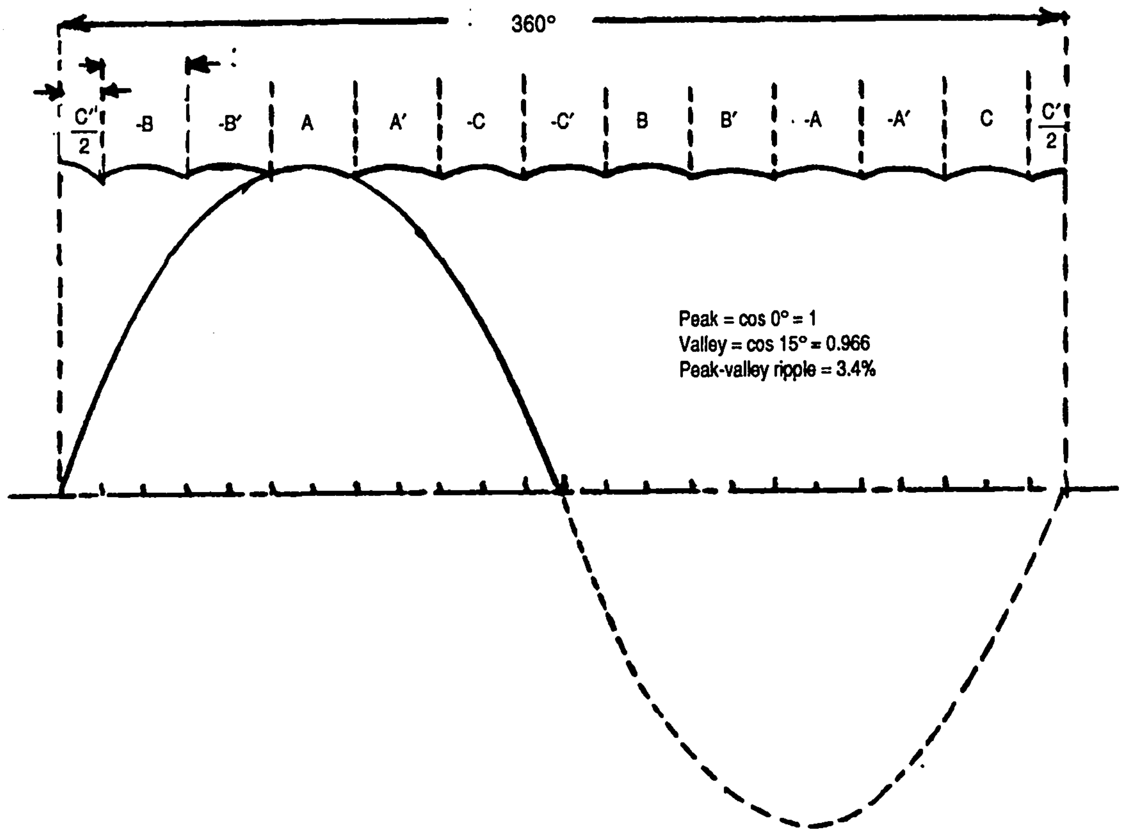

Figure 14-5. The phase progression for 12-pulse rectification.

the six-phase, 12-pulse rectifier produces nearly pure non-varying dc. Disregarding the commutation effect of source inductance, the valleys in the time-domain output voltage are 97% of the peaks, for a peak-to-valley ratio of 1.035. The Fourier series describing the theoretical no-load output spectrum is

$$\frac{12}{\pi}\times\sin\frac{\pi}{12}\left[1+\frac{2}{143}\cos 12\omega t-\frac{2}{575}\cos 24\omega t+\frac{2}{1295}\cos 36\omega t K+\frac{2}{(12n)^2-1}\cos 12n\omega t\right].$$

The first term is 0.9886. The first unavoidable ripple component is the 12th harmonic, which is already attenuated by the factor 2/143, making it 37 dB below the "dc" term. It should be noted that high-voltage power supplies for use in applications that require extremely low ripple amplitude (such as -120 dB) at relatively high ripple frequencies (such as 10 kHz or above) often use rectifiers like this operated from 60 Hz. A component at 10.08 kHz, for instance, would be the 168th harmonic of 60 Hz, having a harmonic number, n, of 14. It would be attenuated by a factor of $2/[(12 \times 14)^2 - 1]$, or 1/28223, which is -89 dB without filtering. The residue left after filtering is often the noise produced by corona, or partial-discharge, resulting from imperfect high-voltage design. Such a corona often becomes the irreducible noise floor.

The disadvantage of the "delta-wye" transformer secondary arrangement is

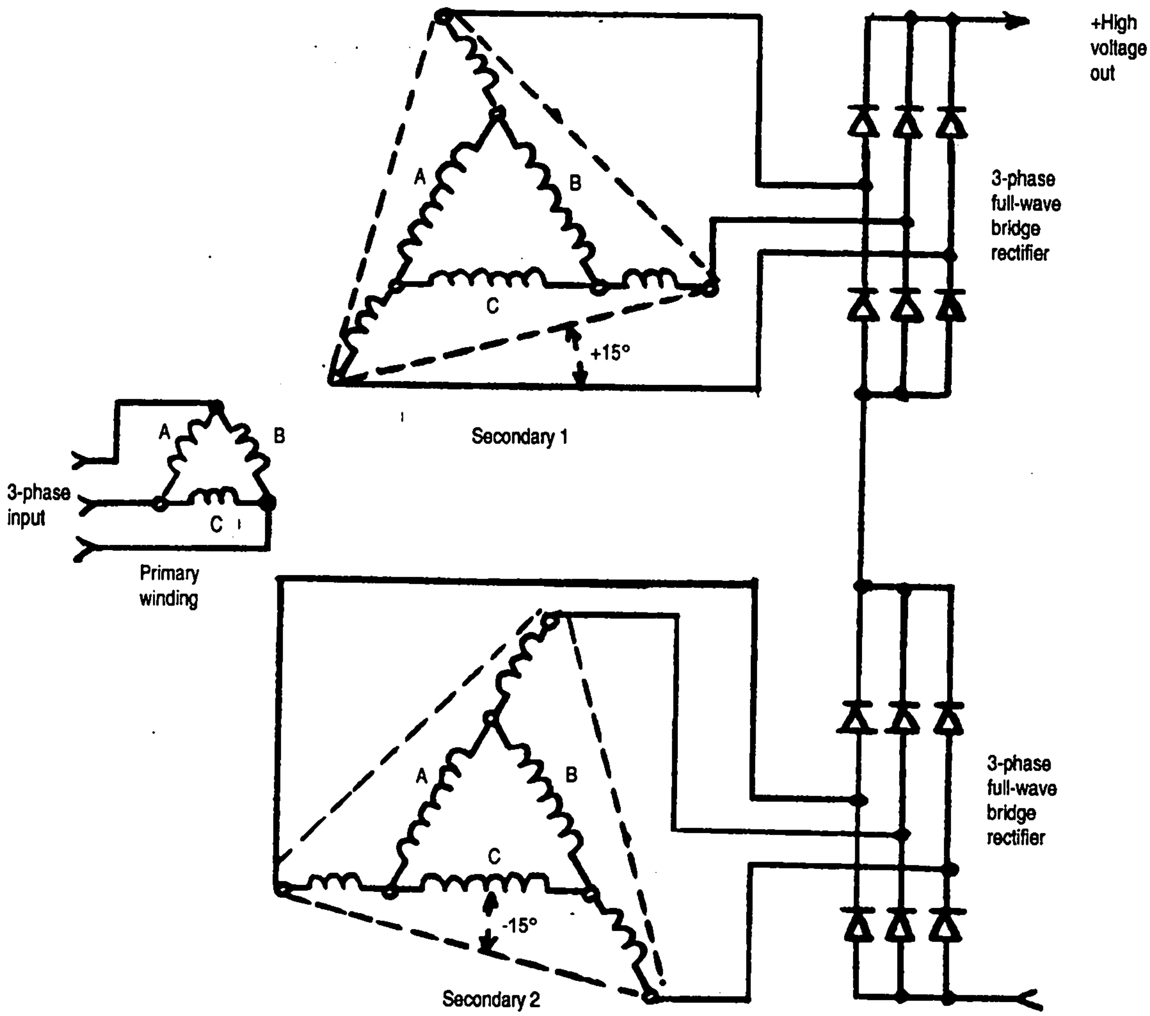

Figure 14-6. The "extended-delta" six-phase, full-wave, or 12-pulse rectifier.

largely overcome by the "extended-delta" pair of secondary windings, as shown in Fig. 14-6. In this case, identical secondary windings can be used. The windings for each phase have taps, or "extensions," at both ends. If the taps are properly placed, making the connections shown will produce a +15° "tilt" to the line-line resultant voltage vectors of the upper secondary and a -15° tilt to those of the lower one, thus generating the 30° interphase angle between the outputs of the series-connected full-wave rectifiers.

14.5 The 12-phase, full-wave, 24-pulse rectifier

Figure 14-7 shows an extension of the 12-pulse rectifier into a 24-pulse rectifier. This is an actual commercial design, comprising a pair of complete "delta-wye"-type 12-pulse rectifiers, each with its own complete transformer. The primary windings of the two transformers are identically wound and connected in a configuration called "almost delta," or "polygon." For both primary windings, a small C-phase winding segment is connected between the A and B junction, a small A-phase winding segment is connected between B and C, and a small B-phase winding segment is connected between C and A. Therefore, all phases are identical. However, the feed lines from the incoming three-phase source are not

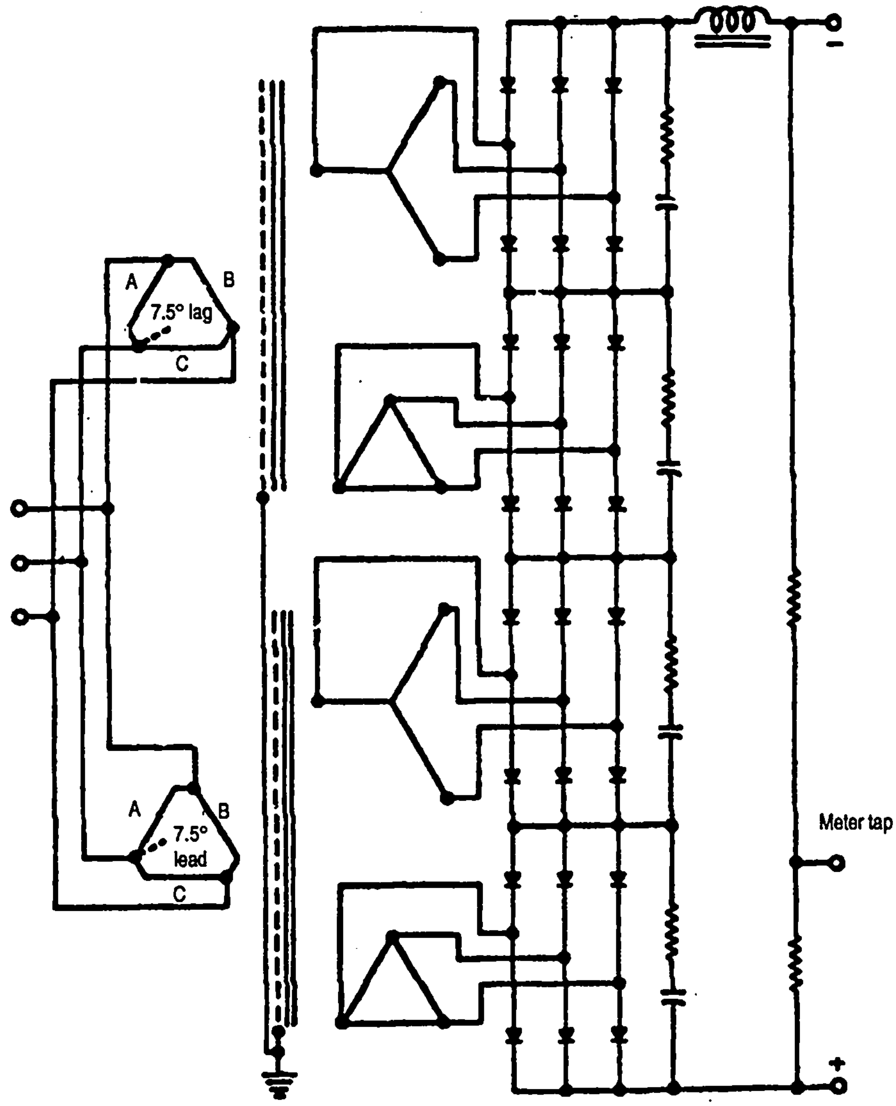

Figure 14-7. A 24-pulse rectifier system that uses polygon primaries.

connected to identical points. The difference imparts a 7.5° lead to one resultant voltage vector and 7.5° lag to the other for a total difference of 15° "inter-inter-phase." In other words, that the two 12-pulse rectifier outputs are interleaved to give 15° conduction angle for each sequential phase segment.

As can well be imagined, there is no theoretical upper limit to how far such polyphase upscaling can be carried, but there is certainly a point of diminishing returns. Although more than a few 24-pulse-rectifier systems are currently operating, this configuration is considered the practical upper limit.

14.6 Polyphase-rectifier line current

So far, the discussion of polyphase rectification has dwelt upon design and user-friendly aspects, such as how close can the outputs come to true non-varying dc, and how tenuous is the ripple spectrum. There is another reason to admire polyphase rectification that is more selfless and environmentally friendly: it inherently ameliorates line-current harmonic pollution.

Consider the typical front end of an "off-line" switch-mode ac-dc converter operating from single-phase ac input, as shown in Fig. 14-8. It is designed to generate a 280-Vdc "rail" from either 115-Vac or 230-Vac input by using the rectifiers as either a full-wave bridge (230 V) or as a full-wave voltage doubler (115 V), depending upon the switch position. The output voltage from the rectifiers into a resistive load bears little resemblance to non-varying dc. For this reason, huge hold-up storage capacitors, C_H, are charged to the peak value of the incoming alternating voltage and hold the value nearly constant until the next positive peak of full-wave rectifier output. The rectifier conduction angle, however, tends toward zero because the rectifiers are back-biased for most of the time.

All of the power consumed is contained in the product of the line voltage and the fundamental-frequency component of line current, which is in phase with the voltage. The total RMS value of the spikelike input current waveform can be two to three times as great as the in-phase, fundamental-frequency component, and the non-useful contributors to the total are virtually all harmonic components because the fundamental-frequency component is very nearly in phase with the line voltage. (When the fundamental-frequency component of line current is not in phase with the voltage, the effect on power factor is called "displacement,"

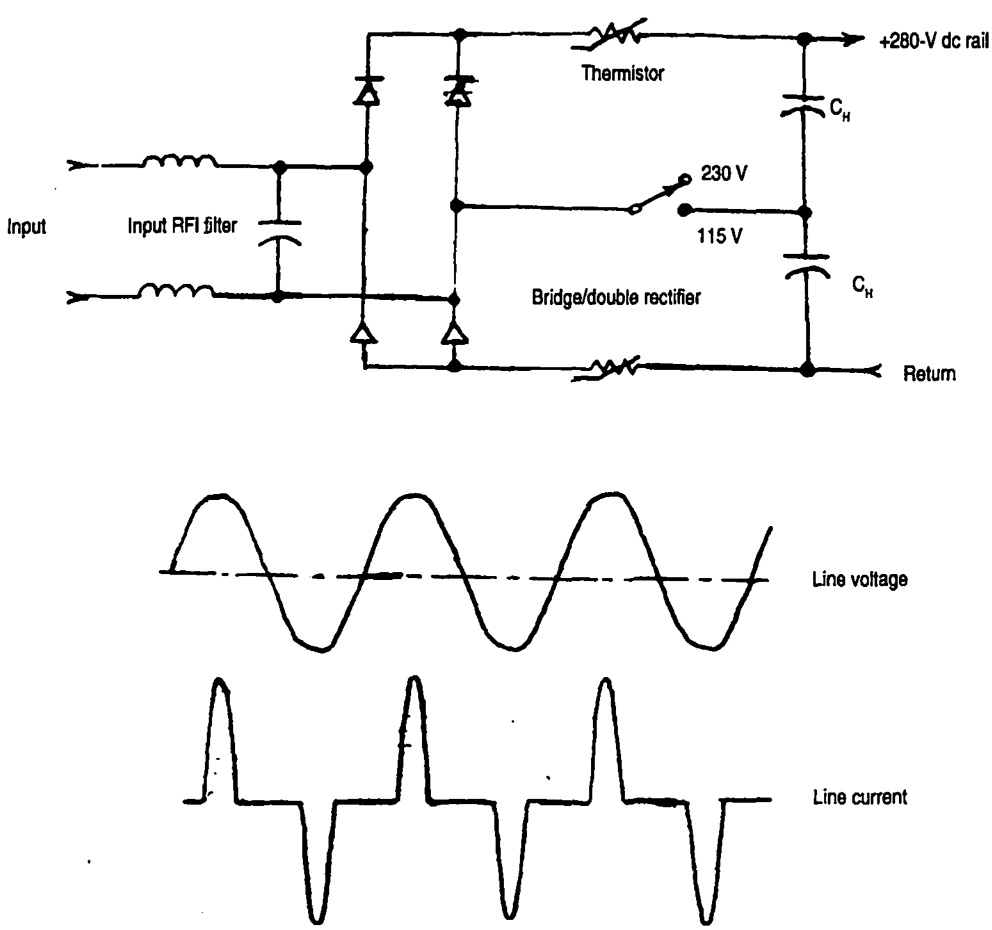

Figure 14-8. Typical front end of off-line, single-phase power supply and typical current wave shape.

and it is proportional to the cosine of the relative phase. However, when the decreased power factor results from harmonic components of line current, the effect is described as "distortion.")

Even though the conversion efficiency of such a supply can be over 95%—the only losses being diode-voltage drops—the power factor, which is the ratio of the input power to the total input volt-amperes (including harmonic components), can easily be less than 0.5. This means that the total RMS input current can be twice as great as the useful component, thus heating up feed lines and circuit breakers four times as much as the useful component would by itself.

The power factor of such a supply would be improved if the dc filter input were inductive instead of capacitive. If the inductor were large enough, the load current would be non-varying dc, and the line current would then be a square wave that has a peak amplitude equal to the dc output current. The peak value of fundamental-frequency component that would be $4/\pi$ times the square-wave peak amplitude and it would be in phase with the voltage. The RMS value of this current is

$$\frac{4}{\pi}\times\frac{1}{\sqrt{2}}\times I_{Pk},$$

and the RMS value of total input current is I_{Pk}. Assuming the line voltage has a peak value of V_p, the average dc output voltage would be $2/\pi \times V_p$, and the RMS value of input voltage would be

$$\frac{V_p}{\sqrt{2}}.$$

The dc output power would be

$$V_{dc}\times I_{dc}=\frac{2}{\pi}V_p\times I_{Pk}.$$

With no rectifier or inductor losses, the input and output power would be equal. The input power is the product of the in-phase fundamental-frequency components of line voltage and current, which is

$$\frac{V_p}{\sqrt{2}}\times I_{Pk}\times\frac{4}{\pi\sqrt{2}}=\frac{2}{\pi}V_p\times I_{Pk},$$

as shown above. The total input volt-ampere rating is

$$\frac{V_p}{\sqrt{2}}\times I_{Pk}.$$

The ratio of the two is the power factor, which in this case is 0.9—a big improvement over the capacitive input filter. However, there is a problem with inductor

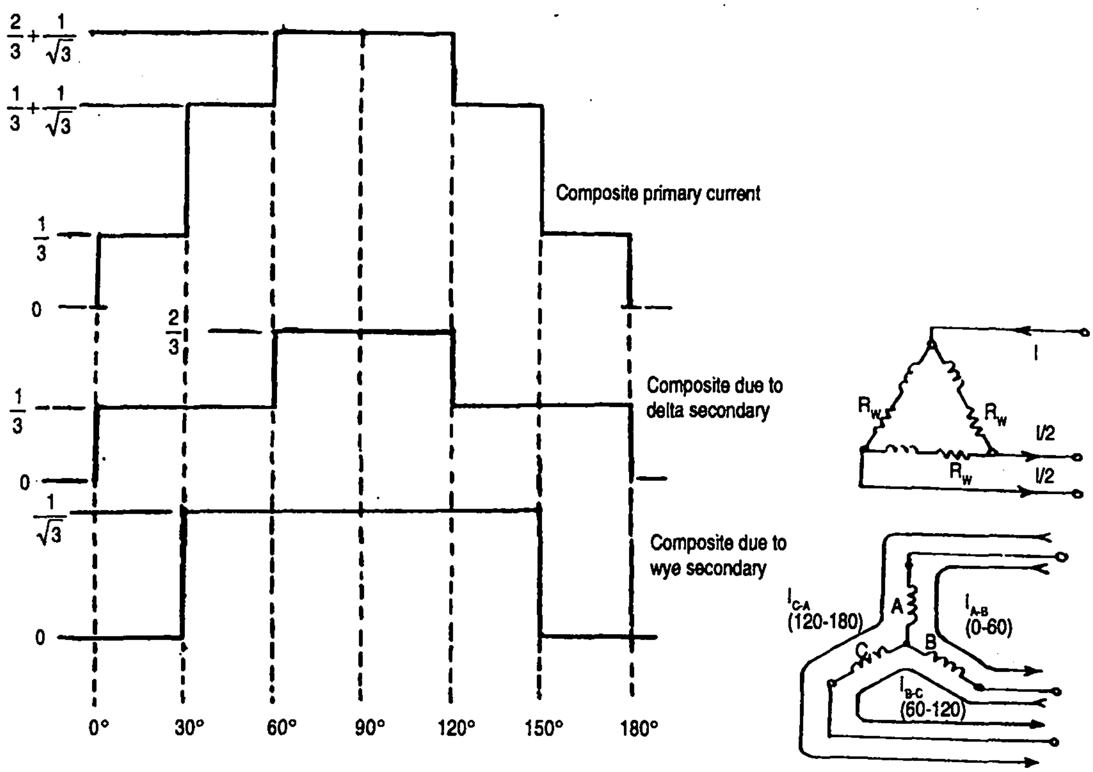

Figure 14-9. Theoretical input line current for delta-wye 12-pulse rectifier.

inputs.

Inductors of the type that will do the job are large, heavy, and expensive, and they subvert the purpose of the high-frequency switch-mode converter that follows the off-line rectifier. In addition, the rail voltage will be lower in the bridge-rectifier mode, and the voltage-doubler will not work at all.

Consider, instead, the waveforms for the line current resulting from the delta and wye windings of a 12-pulse rectifier, shown in Fig. 14-9. With a resistive load on the rectifier, the load current will be substantially constant even without a large input inductor filter, because the rectifier output is so close to non-varying dc to begin with.

The component due to the delta winding will have a peak value of 2/3 of the output current, which flows for the 60° interval centered on the peak. For the other 120° of each half-cycle, the current will be 1/3 of the peak. This is because the current, *I*, will split at the delta junctions in accordance with the dc resistance of the windings. Each phase winding is shunted by the other two in series, thus this configuration has twice the dc resistance of the single-phase winding. Therefore, the current will split with the ratio of 2:1, so 2/3 will flow in the phase winding having the highest instantaneous voltage and 1/3 will flow in the series combination of the other two.

The component due to the wye-connected winding will have a conduction angle of the center 120° of each half-cycle because, at any given time, current is flowing in 2 of the 3 phases. Each phase is idle only 1/3 of the time. The amplitude of the current will be $1/\sqrt{3}$ as great as the dc value because of the $\sqrt{3}$ ratio of the delta and wye turn ratios to ensure equal rectifier input voltages.

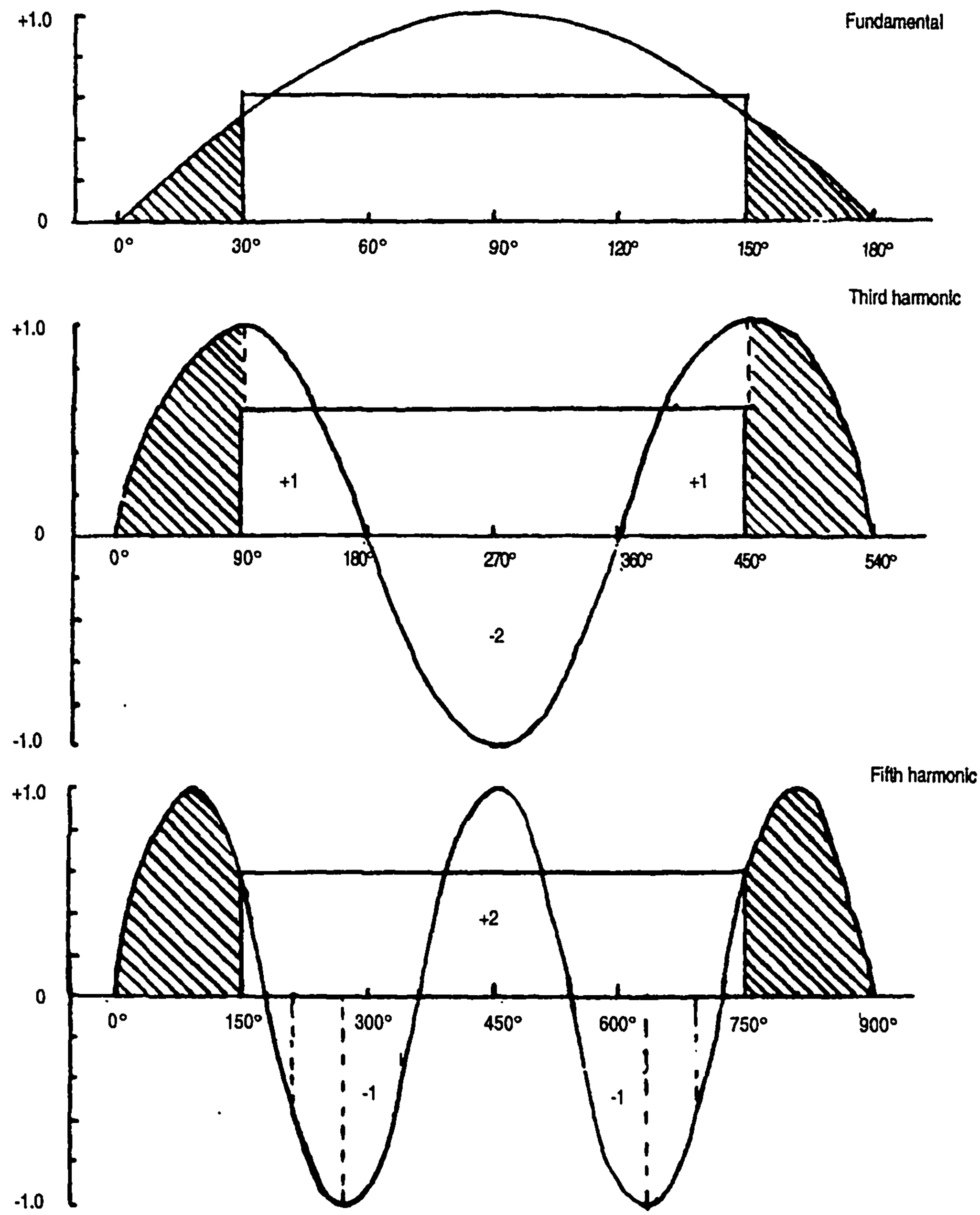

Figure 14-10. Graphical-analysis Fourier series through the fifth harmonic of a six-pulse rectifier line current.

The theoretical Fourier series for the line current of a rectifier system having a pulse number of P is

$$I = \sin \omega t + \frac{1}{P-1}\sin(P-1)\omega t + \frac{1}{P+1}\sin(P+1)\omega t$$
$$+ \frac{1}{2P-1}\sin(2P-1)\omega t + \frac{1}{2P+1}\sin(2P+1)\omega t K$$
$$+ \frac{1}{nP-1}\sin(nP-1)\omega t + \frac{1}{nP+1}\sin(nP+1)\omega t .$$

Note that the line-current harmonic components are the odd values just below and above the rectifier-output harmonic components. For instance, a six-pulse rectifier would have a theoretical first ripple component of the sixth harmonic, while the first line-current harmonics would be the fifth and seventh. For those who might wish to see a very simple-minded proof that requires no mathematics at all, Fig. 14-10 illustrates this phenomenon for the line current due to the wye-connected secondary winding. If we disregard the coefficients and concentrate on the harmonic-component ratios, a Fourier series is evaluated by multiplying the subject waveform by all of the harmonically related sine and cosine waveforms (in-phase and quadrature). Assigning a value of unity to the line current, the fundamental can be seen to be a sine-wave segment from 30° to 150°. The third harmonic sine wave is multiplied by zero between 0° and 90° and again between 450° and 540°. For the remainder, just using the similar shapes, there are two positive quarter-cycles (+1) and one negative half-cycle (-2), which cancel, so there is no 3rd harmonic component. (If we trace out the cosine waveforms, we can easily see that they all cancel to zero.) The fifth harmonic sine wave, however, has non-zero product between 150° and 750°. In the center, two negative quarter-cycles cancel the one positive half-cycle and the two positive and negative 30° segments at the two ends cancel each other. What is left is two negative segments, one from 210° to 270° and the other from 630° to 690°. If they are pushed together, they have the same wave shape as the fundamental component, except they take up 1/5 of the area.

If one does the same "mathless" analysis of the primary current due to the delta-connected secondary, the same wave shape emerges for the fifth harmonic. But it is positive instead of negative, so it would cancel the fifth harmonic component due to the wye-connected secondary in the 12-pulse rectifier. As the pulse number of the rectifier is increased, it can be seen that the line current becomes less and less polluted by harmonics and increasingly resembles in shape a fundamental-frequency sine wave.

14.7 Who cares about harmonic pollution?

The electrical power utilities certainly care about harmonic pollution. In the first place, they do not get paid for the current it represents, but they do have to suffer the power loss it causes. Furthermore, "trapped" resonances at the harmonic frequencies can cause all manner of electrical mischief in terms of resonant overvoltage. And from almost from the beginning of the electrical age, the ability of power-line harmonics to affect other electrical services has been noted.

The telephone company, whose wires parallel those of the power company in

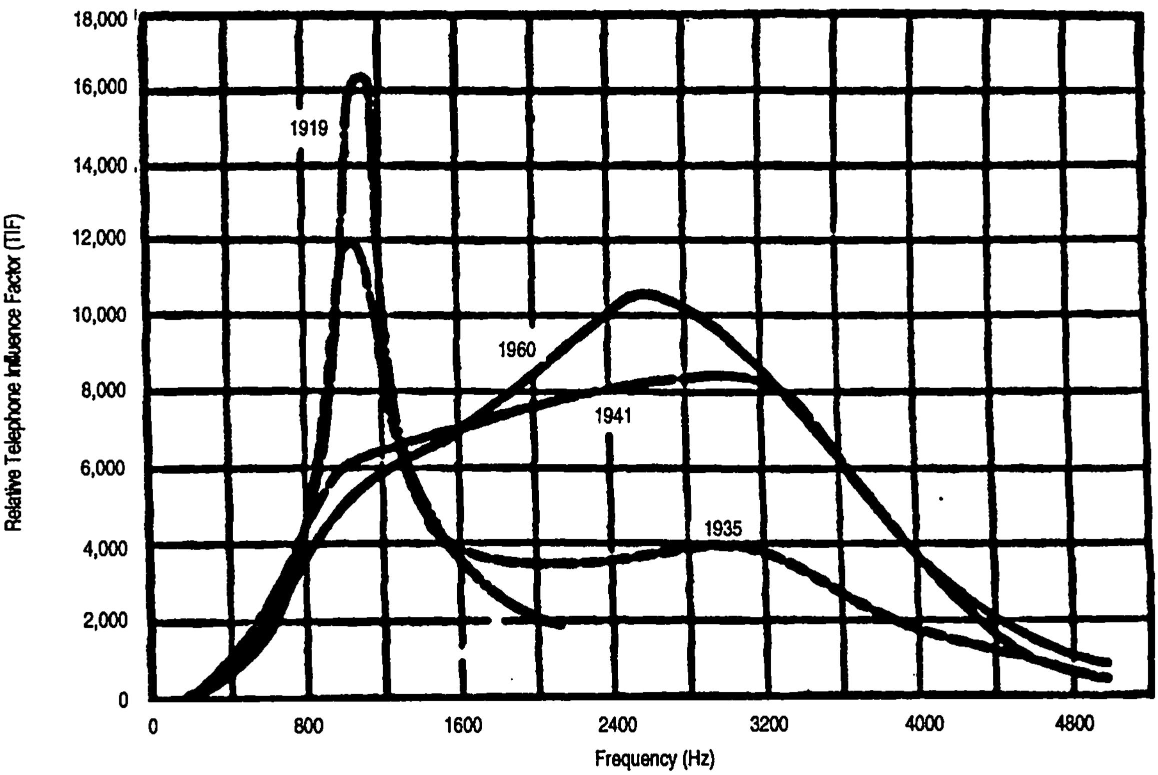

Figure 14-11. Graphs showing changing sensitivity of telephone equipment to coupled harmonic signals.

many situations, has been aware for a long time of the "metallic" noise that power-frequency harmonics can induce. Figure 14-11 shows some graphs that trace the evolution of the most troublesome frequencies the telephone company has had to endure as its equipment has improved in frequency response and fidelity. The graph plots the disturbance frequency against the Telephone Influence Factor (TIF), which is a number that defines relative interference. In 1919, for instance, the most offensive frequency component was barely 1 kHz. This did not mean that the phone equipment itself was the most sensitive at that frequency, because the TIF includes a mechanism that takes into account the coupling between power and phone wires. This coupling increases almost linearly with frequency. By 1960, a broader range of frequencies, peaking at 2800 Hz, became troublesome. Today, harmonic pollution can also disrupt digital data sent from modem to modem. (But the TIF concerns will soon fall by the wayside when all telephone signals are transmitted by fiber-optic cables.) To reduce this pollution problem, some governments and organizations have promulgated strict standards. In fact, present European standards for line-current harmonic content, as well as those standards prescribed by the U.S. Navy, are far more stringent than even the phone company could wish for.

Table 14-1, which tabulates the odd harmonics that would appear in the line current of a six-pulse rectifier ($n \times 6 \pm 1$) through the 73rd harmonic, shows the theoretical and actually measured values of harmonic current for a six-pulse rectifier. It then compares them with a 24-pulse rectifier that the six-pulse rectifier is part of. The values of current are then multiplied by the TIF weighting factor (T) for that particular frequency, which is shown in the third column from the left. The composite TIF, which is only a relative term, is obtained by taking the root of the sum of the squares of each $I \times T$ product and dividing that by the fundamental-frequency current. Note that the six-pulse rectifier has a TIF of 277, whereas the 24-pulse-rectifier TIF is only 60, which is a marked improvement.

14.8 Other forms of line pollution

Another form of objectionable line-current disruption is "flicker" modulation, which is caused by load changes that have a relatively low recurrence rate. Although this type of modulation is by no means peculiar to rectifier loads, the electronic agility of most systems fed by high-power rectifiers make them the most likely to produce this sort of pollution.

The "flicker" referred to is the modulation in the light output of an electrical lighting system. It is most often seen in incandescent light bulbs. Often, large electronic systems will actually have a flicker-modulation specification applied to them that defines a maximum change in line voltage they can produce. Equally often, the specification will be misapplied. How much the line voltage changes depends not only on the variational current demand of the system but on the internal impedance of the power source. An extremely stiff, or low-impedance, source will not permit voltage variation, no matter what the line current does.

Table 14-1. Weighted Telephone Influence Factor calculations for six-pulse and 24-pulse rectifier systems.

			Six-Pulse Rectifier			24-Pulse Rectifier		
Harmonic number	Frequency	TIF Value	Measured line current	Theoretical line current	IxT product	Measured line current	Theoretical line current	IxT product
Fund	60 Hz	0.5	74 A	74 A	37	604 A	604 A	302
5	300	225	16.7	14.8	3760	2.42	0	540
7	420	650	6.47	10.6	4210	1.1	0	720
11	660	2260	4.6	6.7	10,400	0.5	0	1200
13	780	3360	2.98	5.7	10,000	0.33	0	1100
17	1020	5100	1.56	4.3	7960	0.06	0	260
19	1140	5630	1.17	3.9	6590	0.22	0	1220
23	1380	6370	0.5	3.2	3100	4.1	26.3	26,100
25	1500	6680	0.34	3.0	2800	3.2	24.2	21,300
29	1740	7320	0.46	2.6	3400	0.17	0	1240
31	1860	7820	0.31	2.4	2400	0.07	0	560
36	2100	8830	0.44	2.1	3900	0.14	0	1210
37	2220	9330	0.33	2.0	3030	0.15	0	1380
41	2460	10340	0.24	1.8	2480	0.04	0	450
43	2580	10600	0.19	1.7	2050	0.09	0	960
47	2820	10210	0.12	1.6	1230	0.95	13	9660
49	2940	9820	0.08	1.5	790	0.76	12.3	7500
53	3180	8740	0.14	1.4	1190	0.06	0	550
56	3300	8090	0.8	1.35	660	0.08	0	650
59	3540	6370	0.14	1.25	930	0.14	0	930
61	3660	6130	0.1	1.2	630	0.06	0	360
65	3900	4400	--	1.1	--	0.03	0	130
67	4020	3700	--	1.1	--	0.05	0	190
71	4260	2750	0.07	1.0	180	0.41	8.5	1100
73	4380	2190	0.03	1.0	70	0.37	8.3	810

Root sum square of 6-pulse I x T product = 20,500
Weighted Telephone Influence Factor = 20,500/74 A = 277

Root sum square of 24-pulse I x T product = 36,100
Weighted Telephone Inluence Factor = 36,100/604 A = 60

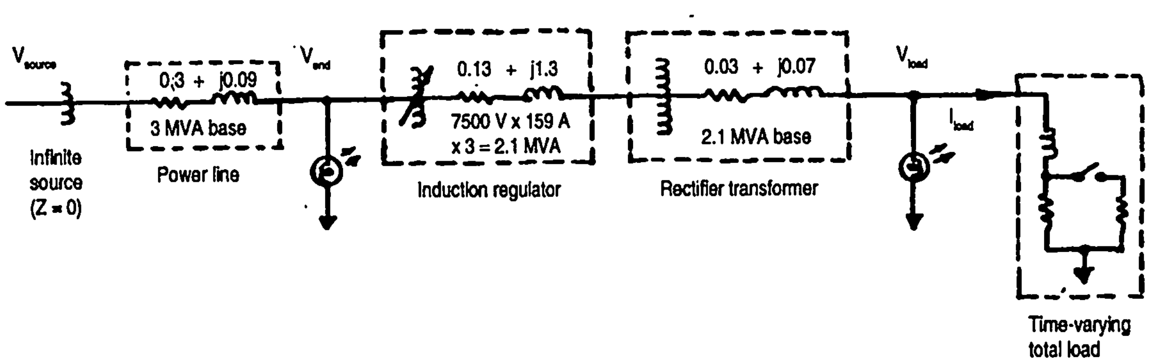

Figure 14-12. Typical line impedance for large transmitter system.

Not all sources are stiff enough, however, to keep flicker within prescribed limits. Therefore, they may need to be investigated.

Figure 14-12 shows the impedance diagram of an actual high-power feed, in this case the one that supplies the Haystack Hill complex. The reason that it was of interest is that the LRIR had as one of its modes of transmitter operation a 50% duty factor at a 10 pps repetition rate. The peak demand during the 50-ms "on" time, was more than 2 MW. It was assumed that picking up and dropping a 2-MW load 10 times per second might not go unnoticed. Reviewing Fig. 14-12, therefore, will provide a little refresher in normalized power-source impedance and the concept of VA base, or, in this case, MVA base.

The eventual MVA base, 2.1 MVA, is that of the transmitter dc power supply. To determine the total source impedance in per-unit or percent, it is necessary to normalize the other components to the same base. The resistive components of source impedance can be ignored because their contribution to total impedance is negligible. The significant impedance of the rectifier transformer, *j* 0.07, is almost entirely leakage reactance and is already normalized to its own load-impedance base of 2.1 MVA. The reactance of the variable transformer ahead of it, the induction regulator, is given in absolute terms, as *j* 1.3 ohms. To normalize it we must find its load impedance, which is the square of its maximum output voltage, 7500 V, divided by its MVA base, 2.1 MVA, or 26.8 ohms. Therefore, the per-unit value is therefore *j* 1.3 ohms divided by 26.8 ohms, or *j* 0.05. What remains to consider is the power line that feeds the system from the virtually infinite source. Its per-unit reactance is *j* 0.09, but its VA base is 3 MVA instead of 2.1 MVA. To convert this value it is necessary to multiply the reactance by 2.1/3, giving *j* 0.063 as the per-unit reactance on a 2.1 MVA base. The total reactance

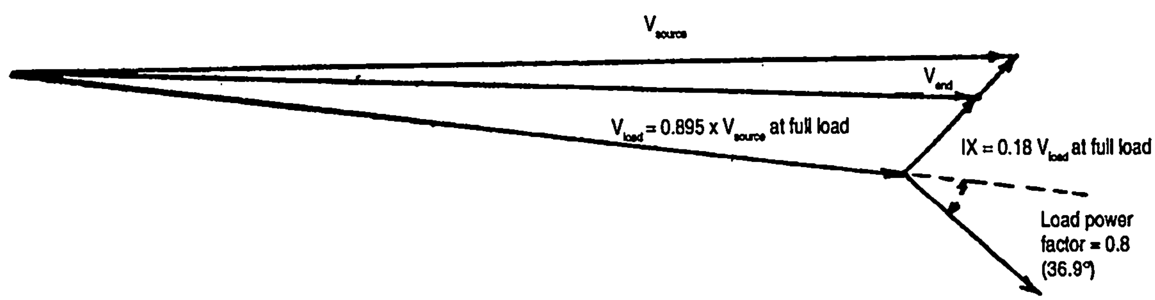

Figure 14-13. Voltage regulation of power system shown in Fig. 14-12.

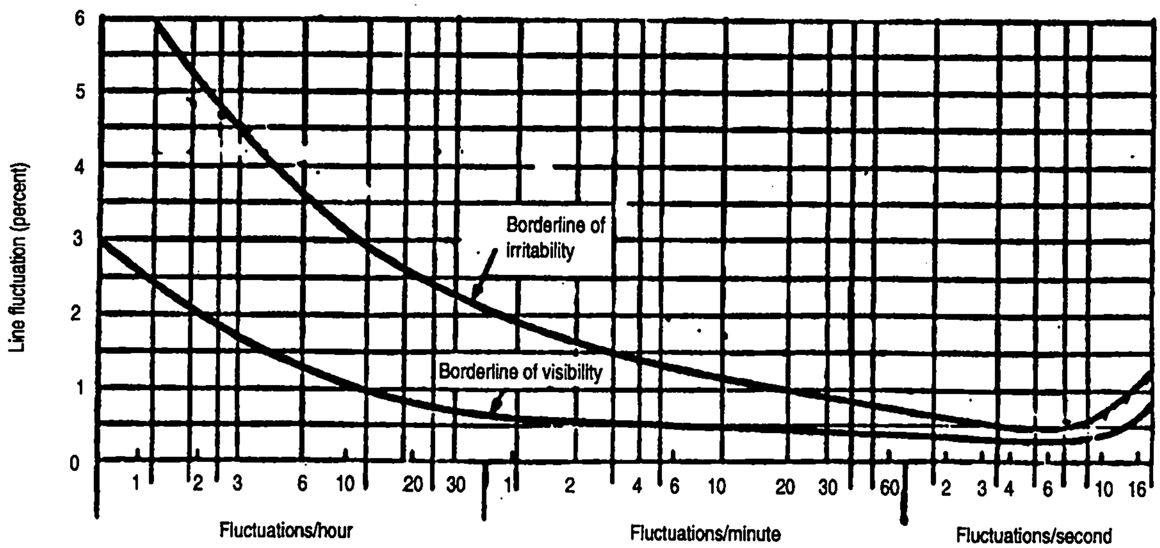

Figure 14-14. Subjective properties of "flicker" modulation.

then is $j(0.063 + 0.05 + 0.07) = j\ 0.18$, or 18% reactance, which is a long way from being "stiff."

Figure 14-13 shows how the load voltage at full-load would differ from the source or no-load voltage. The load power factor is approximately 0.8, so the line current will lag the load voltage by 37°. The voltage drop across the source reactance will be in quadrature with the current. The vector sum of source voltage and reactance voltage drop is 0.895 times the source voltage, indicating about 10% no-load-to-full-load voltage change. Is this a little or a lot?

Figure 14-14 shows two curves that are the results of subjective responses of people to the lighting flicker caused by different voltage changes at different flicker rates. One curve plots the borderline of irritability, the other the borderline of visibility. The total area above each curve represents an area where the values of voltage change produce visible flicker or induce irritability. Note that humans beings are most sensitive, both in terms of visual perception and emotional irritability, when the flicker rate falls between 8 and 10 fluctuations per second (sound familiar?). A voltage fluctuation of about 0.25% produces observable lighting flicker, whereas voltage fluctuation of less than 0.5% brings on irritation. Needless to say, the high-voltage dc-power-conditioning system for the Haystack Hill LRIR transmitter required serious load-transient mitigation in the form of a large energy-storage capacitor bank, electronic voltage regulation, and current regulation to keep the load on the power line constant regardless of operating mode on the dc output side.