4

Effects of Harmonic Distortion

4.1 Introduction

Once the harmonic sources are clearly defined, they must be interpreted in terms of their effects on the rest of the system and on personnel and equipment external to the power system.

Each element of the power system must be examined for its sensitivity to harmonics as a basis for recommendations on the allowable levels. The main effects of voltage and current harmonics within the power system are:

- The possibility of amplification of harmonic levels resulting from series and parallel resonances.
- A reduction in the efficiency of the generation, transmission and utilisation of electric energy.
- Ageing of the insulation of electrical plant components with consequent shortening of their useful life.
- Malfunctioning of system or plant components.

Among the possible external effects of harmonics are a degradation in communication systems performance, excessive audible noise and harmonic-induced voltage and currents.

4.2 Resonances

The presence of capacitors, such as those used for power factor correction, can result in local system resonances, which lead in turn to excessive currents and possibly subsequent damage to the capacitors [1].

4.2.1 Parallel Resonance

Parallel resonance results in a high impedance at the resonant frequency being presented to the harmonic source. Since the majority of harmonic sources can be considered as

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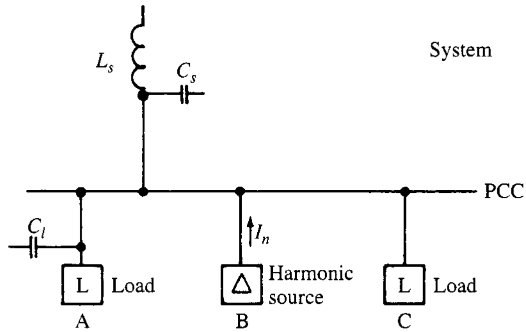

Figure 4.1 Parallel resonance at a point of common coupling (PCC)

current sources, this results in increased harmonic voltages and high harmonic currents in each leg of the parallel impedance.

Parallel resonances can occur in a variety of ways, the simplest perhaps being that where a capacitor is connected to the same busbar as the harmonic source. A parallel resonance can then occur between the system impedance and the capacitor.

Assuming the system impedance to be entirely inductive, the resonant frequency is

$$f_p = f\sqrt{\left(\frac{S_s}{S_c}\right)} \tag{4.1}$$

where f is the fundamental frequency (Hz), f_p is the parallel resonant frequency (Hz), S_s is the short-circuit rating (VAr) and S_c is the capacitor rating (VAr).

Further opportunities for parallel resonance can occur with a more detailed representation of the system. For instance in Figure 4.1 the harmonic current from consumer B encounters a high harmonic impedance at the busbar. This may be due to a resonance between the system inductance (L_s) and the system (C_s) and/or load capacitance (C_l).

To determine which resonance condition exists it is necessary to measure the harmonic currents in each consumer load and in the supply, together with the harmonic voltage at the busbar. In general, if the current flowing into the power system from the busbar is small, while the harmonic voltage is high, resonance within the power system is indicated. If instead a large harmonic current flows in consumer A's load and leads the harmonic voltage at the busbar, resonance between the system inductance and the load capacitor is indicated.

4.2.2 Series Resonance

Consider the system of Figure 4.2. At high frequencies the load can be ignored as the capacitive impedance reduces. Under these conditions a series resonant condition will exist when

$$f_s = f\sqrt{\left(\frac{S_t}{S_c Z_t} - \frac{S_l^2}{S_c^2}\right)}, \tag{4.2}$$

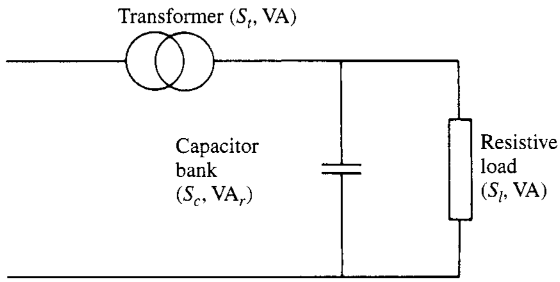

Figure 4.2 Series resonance circuit

where f_s is the series resonant frequency (Hz), S_t is the transformer rating, Z_t is the transformer per unit impedance, S_c is the capacitor rating and S_l is the load rating (resistive).

The concern with series resonance is that high capacitor currents can flow for relatively small harmonic voltages. The actual current that will flow will depend upon the quality factor Q of the circuit. This is typically of the order of 5 at 500 Hz.

4.2.3 Effects of Resonance on System Behaviour

Power Factor Correction Capacitors Harmonic resonances affect the design of power factor correction capacitors. The overload current capability of these capacitors is discussed in Section 4.4.3.

This problem can be illustrated with reference to a study carried out for an iron-sand mining plant consisting of six-pulse rectifiers and fed from a long-distance 11 kV line. The original rating of the scheme (8 MW at 0.85 power factor) was to be increased to 10 MW by the shunt connection of power factor correction capacitors at the plant terminals.

The minimum fault level at the point of connection was 35.4 MVA, which (on a 10 MW power base) is equivalent to a per unit maximum system impedance of

$$x_s = (V^2)/(\text{MVAF/MVAB}) = (1)/(35.4/10) = 0.28 \text{ p.u.}$$

The conversion from 8 to 10 MW required 3 MVAR (or 0.3 p.u. on the 10 MVA base), the corresponding capacitance being

$$x_c = V^2/\text{MVAR} = 1/(0.3) = 3.3 \text{ p.u.}$$

and this capacitance was divided into three independently switched banks.

The converter harmonic currents were now injected into the a.c. system impedance in parallel with the PF correction capacitors, the corresponding parallel resonant frequency resulting from the equation:

$$n\,x_s = (1/n)x_c \qquad \text{where } n = \omega_n/\omega$$

Therefore $n^2 = x_c/x_s = 3.3/0.28 = 11.8$, and $n = 3.4$ (when all the capacitance is connected), $n = 4.2$ (with one of the three banks disconnected) and $n = 5.9$ (with two banks disconnected).

However, measurements carried out at the plant indicated that the levels of third ($n = 3$) and fourth ($n = 4$) harmonic current content were insignificant, and the decision was made to eliminate the possibility of fifth harmonic resonance by placing 4% inductance in series with the capacitors.

Another area where resonance effects may lead to component failure is associated with the application of power line signalling (ripple control) for load management. In such systems tuned stoppers (filters) are often used to prevent the signalling frequency from being absorbed by low-impedance elements such as power factor correction capacitors. A typical installation is shown in Figure 4.3.

Where local resonances exist, excessive harmonic currents can flow, resulting in damage to the tuning capacitors. Figure 4.4 shows the harmonic currents recorded at one such installation where failure of this type occurred.

In another installation tuned stoppers (at 530 Hz) were fitted to 15 × 65 kVAr steps of power factor correction capacitance, each stopper rated at 100 A. Most of the stopper tuning capacitors failed within two days. The problem was eventually traced to a local

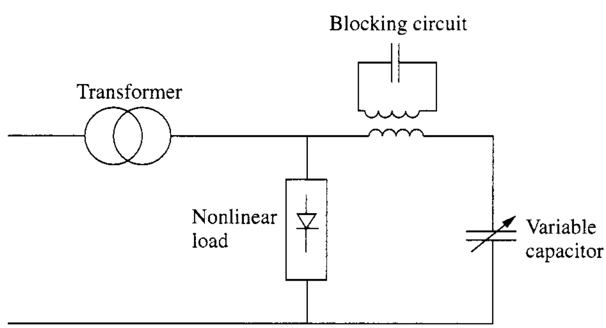

Figure 4.3 Tuned stopper circuit for ripple control signal

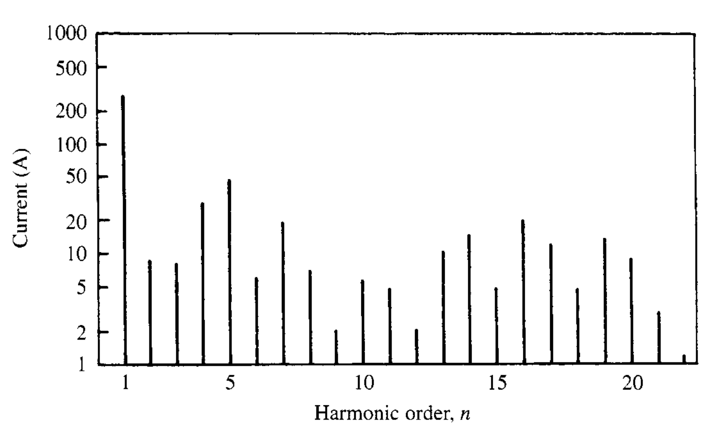

Figure 4.4 Harmonic currents measured through a blocking circuit

power system harmonic at 350 Hz, near to which frequency the tuned stoppers were found to series resonate with the power factor correction capacitors.

Magnification of Low-Order Harmonics The following simplified reasoning explains the harmonic magnification phenomena. Let us consider the case of a static converter (represented as a harmonic current injector) fed from an a.c. system of internal impedance Z_r at the h harmonic.

The a.c. system admittance at the fundamental and low-order harmonics is predominantly inductive, i.e.

$$Y_r = G_r - jB_r$$

In the absence of filters or compensation, the harmonic current (I_h) generates at the point of connection a harmonic voltage of amplitude

$$V_h = Z_r I_h = I_h / Y_r$$

When a capacitor bank or filters of admittance Y_f are present, the harmonic voltage at the point of connection becomes:

$$V_h' = I_h / (Y_r + Y_f)$$

and, since the admittance of a filter bank is predominantly capacitive ($Y_f = jB_f$),

$$V_h' = I_h / (G_r - jB_r + jB_f)$$

When $B_r = B_f$ the harmonic voltage is only limited by the system resistance, which is generally very small. Thus, when $Y_r + Y_f < Y_r$ the harmonic distortion is magnified, the magnification factor being:

$$V_h' / V_h = Y_r / (Y_r + Y_f)$$

Consequently, a low-order non-characteristic harmonic current, which has no practical adverse effect in the absence of the capacitor or filter banks, can be amplified to give a voltage greater that the filtered harmonics.

4.2.4 Complementary and Composite Resonances

The traditional definition of resonance as described above is used with reference to isolated parts of an overall system (e.g. the a.c. or d.c. sides of a static converter). This sort of resonance is well defined, being the frequency at which the capacitive and inductive reactances of the circuit impedance are equal. At the resonant frequency, a parallel resonance has a high impedance and a series resonance a low impedance.

This approach has led to the concept of *complementary resonance*, i.e. a high-impedance parallel resonance at a harmonic on the a.c. side closely coupled to a low-impedance series resonance at an associate frequency on the d.c. side. The first-order associate a.c. and d.c. side frequencies, derived from the general table of

a.c. side	d.c. side
Positive sequence harmonics $k+1$ $13 \pm k$ $25 \pm k$ $12n+1 \pm k, n = 3,4,\ldots$ $\vdots$	
	k $12 \pm k$ $24 \pm k$ $12n \pm k, n = 3,4,\ldots$ $\vdots$
Negative sequence harmonics $k-1$ $11 \pm k$ $23 \pm k$ $12n-1 \pm k, n = 3,4,\ldots$ $\vdots$	

Figure 4.5 Associate three-port harmonic orders

Figure 3.50, are shown in the three-port model of Figure 4.5. A reported experience of this condition, and used in the commonly used CIGRE benchmark HVd.c. test system [2], involves a second-harmonic parallel resonance on the a.c. side and a fundamental frequency series resonance on the d.c. side.

Moreover, when the a.c. and d.c. systems are interconnected by a static converter, the system impedances interact via the converter characteristics to create entirely different resonant frequencies. The term 'composite resonance' has been proposed [3] to describe this sort of resonance, emphasising its dependence on all the components of the system

A composite resonance may be excited by a relatively small distortion source in the system, or by an imbalance in the converter components or control. The resulting amplification of the small source by the resonant characteristics of the system can compromise the normal operation of the converter and even lead to instability. A true instability results when, at the composite resonant frequency, the resistance of the overall circuit is negative. This can occur at non-integer frequencies and is driven by conversion from the fundamental frequency and d.c. components to the composite resonance frequency via the converter control. A detailed representation of the control, firing angle and end-of-commutation angle modulation is needed to model this operating condition; this can be done using the advanced models described in Chapter 8.

A frequently reported case of this type in HVd.c. converters is the so-called core saturation instability [4]. This type of instability can be explained with reference to the block diagram of Figure 4.6. If a small level of positive-sequence second-harmonic voltage distortion exists on the a.c. side of the converter, a fundamental frequency distortion will appear on the d.c. side. Through the d.c. side impedance, a fundamental frequency current will flow, resulting in a positive-sequence second-harmonic current and a negative-sequence d.c. flowing on the a.c. side. The negative-sequence d.c. will begin to saturate the converter transformer, resulting in a multitude of harmonic currents being generated, including a positive-sequence second-harmonic current. Associated with this current will be an additional contribution to the positive-sequence second-harmonic voltage distortion and in this way the feedback loop is completed. The stability of the system is determined by this feedback loop.

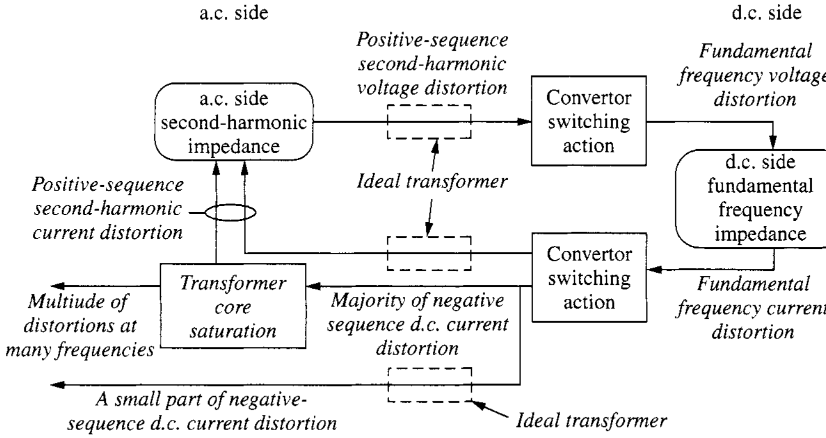

Figure 4.6 Mechanism of core saturation instability

Due to the dynamics of the instability, the d.c distortion is never exactly at the fundamental frequency, therefore the so-called negative sequence d.c. is not a true d.c. but is varying slowly, hence the use of the otherwise inappropriate term 'negative-sequence d.c.' The same mechanism could be triggered by the presence of some d.c. side fundamental frequency current induced by the proximity of an a.c. transmission system.

4.2.5 Poor Damping

Considerable power at the consumer's end of the system is controlled by power electronics to be constant power loads.

A constant power load, such as a motor variable speed drive or a switched mode power supply, presents a small-signal impedance or resistance to the power system that is negative; that is, any rise in voltage causes the current to fall. This removes the damping, or broadband energy absorption capability, from the power system. It can be expected that the continuous addition of such loads to the power system will cause instabilities or poor performance in the future

4.3 Effects of Harmonics on Rotating Machines

4.3.1 Harmonic Losses

Non-sinusoidal voltages applied to electrical machines may cause overheating. Motors are not normally derated so long as the harmonic distortion remains within the 5% normally recommended by the regulations. Above that limit they will often experience excessive heating problems. On the positive side, motors contribute to the damping of the system harmonic content by virtue of the relatively high X/R ratio of their blocked rotor circuit.

Harmonic voltages or currents give rise to additional losses in the stator windings, rotor circuits, and stator and rotor laminations. The losses in the stator and rotor conductors are greater than those associated with the d.c. resistances because of eddy currents and skin effect.

Leakage fields set up by harmonic currents in the stator and rotor end-windings produce extra losses. In the case of induction motors with skew rotors the flux changes in both stator and rotor and high frequency can produce substantial iron loss. The magnitude of this loss depends upon the amount of skew, and the iron-loss characteristics of the laminations.

As an illustration of the effect of supply waveform distortion on the power loss, reference [5] reports on the case of a 16 kW motor, operating at full output and rated fundamental voltage (at 60 Hz). With a sinusoidal voltage supply the total loss is 1303 W, whereas with a quasi-square voltage supply the total loss is 1600 W.

The following typical distribution of losses caused by supply harmonics has been reported [6] for the case of an inverter-fed machine: stator winding, 14.2%; rotor bars, 41.2%; end region, 18.8%; skew flux, 25.8%.

It would be inappropriate to apply the above loss breakdown to individual harmonics or to extend them to other machines, but it is clear that the major loss component is in the rotor. With the exception of the skew losses, the loss subdivision of a synchronous machine should follow a similar pattern.

When considering the harmonic heating losses in the rotor of synchronous machines it must be remembered that pairs of stator harmonics produce the same rotor frequency. For example the fifth and seventh harmonics both give induced rotor currents at frequency $6f_1$. Each of these currents takes the form of an approximately sinusoidal spatial distribution of damper bar currents travelling around the rotor at velocity $6\omega_1$, but in opposite directions. Thus for a linear system, the average rotor surface loss density around the periphery will be proportional to $(I_5^2 + I_7^2)$; however, because of their opposing rotations, at some point around the periphery the local surface loss density will be proportional to $(I_5 + I_7)^2$. If the fifth and seventh harmonic currents are of similar magnitude then the maximum local loss density would be about twice the average loss density caused by these two currents.

Extra power loss is probably the most serious effect of harmonics upon a.c. machines. An approximate assessment of the additional thermal stress of the coils can be achieved with the help of a weighted distortion factor adapted to inductance, i.e.

$$\mathrm{THD}_L = \frac{\sqrt{\sum_{n=2}^{N}\left(\frac{V_n^2}{n^\alpha}\right)}}{V_1} \tag{4.3}$$

where $\alpha = 1$ to 2, V_n is the single frequency r.m.s. voltage at harmonic n, N is the maximum order of harmonic to be considered and V_1 is the fundamental line to neutral r.m.s. voltage.

The capability of a machine to cope with extra harmonic currents will depend on the total additional loss and its effect on the overall machine temperature rise and local overheating (probably in the rotor). Cage-rotor induction motors tolerate higher rotor losses and temperatures provided that these do not result in unacceptable stator

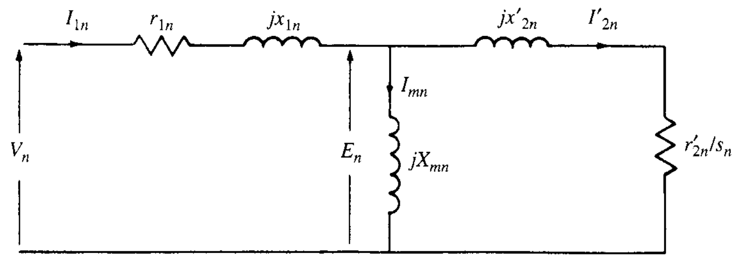

Figure 4.7 Equivalent circuit of induction machine per phase for harmonic n

temperatures, whereas machines with insulated rotor windings may be more limited. Some guidance as to the probably acceptable levels may be obtained from the fact that the level of continuous negative-sequence current is limited to about 10% for generators, and negative-sequence voltage to about 2% for induction motors. It is therefore reasonable to expect that if the harmonic content exceeds these negative-sequence limits, then problems will occur.

4.3.2 Harmonic Torques

The familiar equivalent circuit of an induction machine can be drawn for each harmonic as in Figure 4.7, where all the parameters correspond to actual frequencies of winding currents.

Harmonic currents present in the stator of an a.c. machine produce induction motoring action (i.e. positive harmonic slips S_n). This motoring action gives rise to shaft torques in the same direction as the harmonic field velocities so that all positive-sequence harmonics will develop shaft torques aiding shaft rotation whereas negative-sequence harmonics will have the opposite effect.

For a harmonic current I_n, the torque per phase is given by $I_n^2(r'_{2n}/s_n)$ watts at harmonic velocity. Referred to fundamental velocity this becomes

$$T_n = (I_n^2/n)(r'_{2n}/s_n) \text{ synchronous watts} \tag{4.4}$$

with the sign of n giving the torque direction.

Since s_n is approximately 1.0, equation (4.4) can be written as

$$T_n = (I_n^2/n)r'_{2n} \text{ per unit} \tag{4.5}$$

if I_n and r'_{2n} are per unit.

Using the relationship $V_n = I_n Z_n$ and $Zn \sim nZ_1$, the torque can be expressed in terms of the harmonic voltages, i.e.

$$T_n = (V_n^2/n^3)(r'_{2n}/X_1^2) \tag{4.6}$$

Because the slip to harmonic frequencies is almost unity, the torques produced by practical per unit values of harmonic currents is very small, and moreover the small

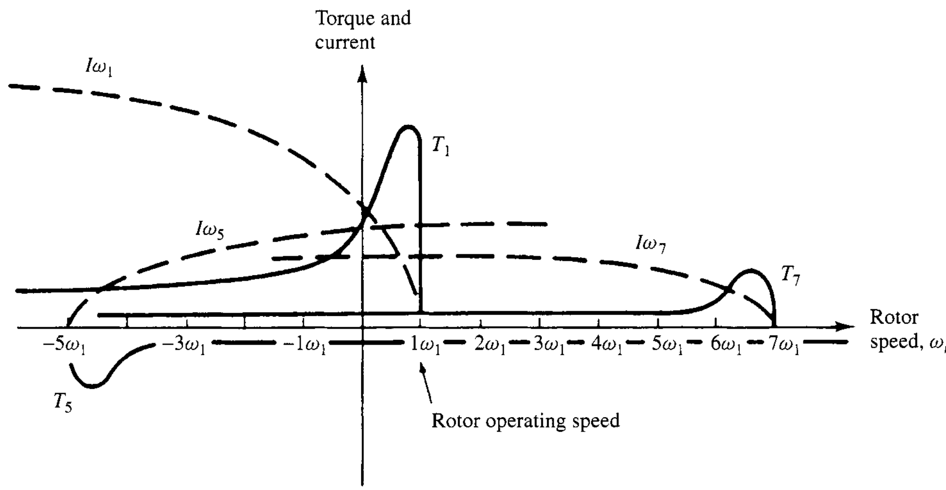

Figure 4.8 Rotor harmonic torques and currents

torques occur in pairs, which tend to cancel. This effect is illustrated in Figure 4.8. Therefore the effects of harmonics upon the mean torque may, in most cases, be neglected.

Although harmonics have little effect upon mean torque, they can produce significant torque pulsations.

Williamson [7] has developed the following approximate expression for the magnitudes of torque pulsations based on nominal voltage:

$$T_{3k} = [I_{n+}^2 + I_{n-}^2 - 2I_{n+}I_{n-}\cos(\phi_{n+} - \phi_{n-})]^{1/2} \text{ per unit}$$

where I_{n+} and I_{n-} are per unit values, $n+$ represents the $1 + 3k$ harmonic orders and $n-$ represents the $1 - 3k$ harmonic orders. This expression permits the preliminary assessment of possible shaft torsional vibration problems.

As an example, let us take the case of a supply voltage with total harmonic distortion of about 4%, resulting in machine currents of 0.03 and 0.02 per unit for the fifth and seventh harmonics, respectively. If both harmonics have the same phase angle, then for a 50 Hz machine on full voltage the torque will have a varying component at 300 Hz with an amplitude of 0.01 per unit. If the harmonics have the most adverse phase relationship, the amplitude will be 0.05 per unit.

4.3.3 Other Effects

The perturbation of the speed/torque characteristic by the presence of harmonics can cause clogging, a term used to describe the failure of an induction motor to run up to normal speed due to a stable operating point occurring at a lower frequency.

The stray capacitances in ASD-fed electric motors in the presence of harmonics cause capacitive currents to flow through the motor bearings and are often a source of their failure [8].

4.4 Effect of Harmonics on Static Power Plant

4.4.1 Transmission System

The flow of harmonic currents in the transmission network produces two main effects. One is the additional power loss caused by the increased r.m.s. value of the current waveform, i.e.

$$\sum_{n=2}^{\infty} I_n^2 R_n$$

where I_n is the nth harmonic current and R_n the system resistance at that harmonic frequency. Skin and proximity effects are functions of frequency and raise the value of the a.c. resistance of the cable, thus increasing the conductor I^2R losses.

The second effect of the harmonic current flow is the creation of harmonic voltage drops across the various circuit impedances. This means in effect that a 'weak' system (of large impedance and thus low fault level) will result in greater voltage disturbances than a 'stiff' system (of low impedance and high fault level).

In the case of transmission by cable, harmonic voltages increase the dielectric stress in proportion to their crest voltages. This effect shortens the useful life of the cable. It also increases the number of faults and therefore the cost of repairs.

The effects of harmonics on Corona starting and extinction levels are a function of peak-to-peak voltage. The peak voltage depends on the phase relationship between the harmonics and the fundamental. It is thus possible for the peak voltage to be above the rating while the r.m.s. voltage is well within this limit.

The IEEE 519 standard provides typical capacity derating curves for cables feeding six-pulse convertors.

4.4.2 Transformers

The primary effect of power system harmonics on transformers is the additional heat generated by the losses caused by the harmonic content of the load current. Other problems include possible resonances between the transformer inductance and system capacitance, mechanical insulation stress (winding and lamination) due to temperature cycling and possible small core vibrations.

The presence of harmonic voltages increases the hysteresis and eddy current losses in the laminations and stresses the insulation. The increase in core losses due to harmonics depends on the effect that the harmonics have on the supply voltage and on the design of the transformer core.

The flow of harmonic currents increases the copper losses; this effect is more important in the case of converter transformers because they do not benefit from the presence of filters, which are normally connected on the a.c. system side. Apart from the extra rating required, converter transformers often develop unexpected hot spots in the tank.

Delta-connected windings can be overloaded by the circulation of triplen frequency zero-sequence currents, unless these extra currents are taken into account in the design. Under this condition a three-legged transformer design can be effectively overloaded by zero-sequence-caused harmonic fluxes. These fluxes cause additional heating in the tanks, core clamps, etc.

If the load current contains a d.c. component, the resulting saturation of the transformer magnetic circuit (described in Section 3.2) greatly increases the harmonic content of the excitation current.

Guidelines for transformer derating to take into account the harmonic content are given in the ANSI/IEEE standard C57.110 based on a derating factor [9] expressed as

$$K = \sqrt{\frac{\sum_h (I_h^2 h^2)}{\sum_h I_h^2}} \tag{4.7}$$

In terms of the above K factor, the following expression is used to determine the derated (or maximum allowed) current:

$$I_{\max} = \sqrt{\frac{1 + P_{EC.R}}{1 + KP_{EC.R}}}(I_R) \tag{4.8}$$

where I_R is the fundamental r.m.s. current under rated load conditions and $P_{EC.R}$ is the ratio of eddy-current loss to rated I^2R loss (I being the total r.m.s. current).

Examples of K-Factor Application

(i) Assume that the converter transformer used in Section 3.6.5 has a P_{EC} (eddy current loss) of 12 kW and an I^2R loss on rating of 100 kW.

In that example the calculated values of the fundamental and harmonic currents were

233.91 A(fundamental), 45.37 A(5th), 31.27 A(7th), 19.08 A(11th) and 14.03 A(13th)

Therefore $P_{EC.R} = 12/100 = 0.12$ and I_R (rated fundamental current) = 233.91 A.

From Equation (4.7)

$$K = \sqrt{\frac{\left(\frac{233.91}{233.91}\right)^2 1^2 + \left(\frac{45.84}{233.91}\right)^2 5^2 + \left(\frac{31.27}{233.91}\right)^2 7^2 + \left(\frac{19.08}{233.91}\right)^2 11^2 + \left(\frac{14.03}{233.91}\right)^2 13^2}{\left(\frac{233.91}{233.91}\right)^2 + \left(\frac{45.84}{233.91}\right)^2 + \left(\frac{31.27}{233.91}\right)^2 + \left(\frac{19.08}{233.91}\right)^2 + \left(\frac{14.03}{233.91}\right)^2}}$$

$$= 1.996$$

and from Equation (4.8) the maximum allowed current is

$$I_{\max} = \sqrt{\frac{1 + 0.12}{1 + (1.985)(0.12)}} \times (233.91) = 222 \text{ A}$$

Therefore the transformer will be overloaded if used with the nominal (i.e. 233.91 A) a.c. current rating.

Table 4.1 Harmonic currents of two heat pumps operating at nominal speed

h	Pump A		Pump B	
	Amplitude (A)	Phase (degrees)	Amplitude (A)	Phase (degrees)
1	11.87	17	14.4	2
3	7.487	0228	12.18	−10
5	3.003	−125	9.84	−17
7	1.329	−89	6.88	−24
9	0.582	67	3.99	−34
11	0.499	−55	1.63	−50
13	0.309	−227	0.39	−141
15	0.273	−68	0.88	151
17	0.154	−164	0.94	132

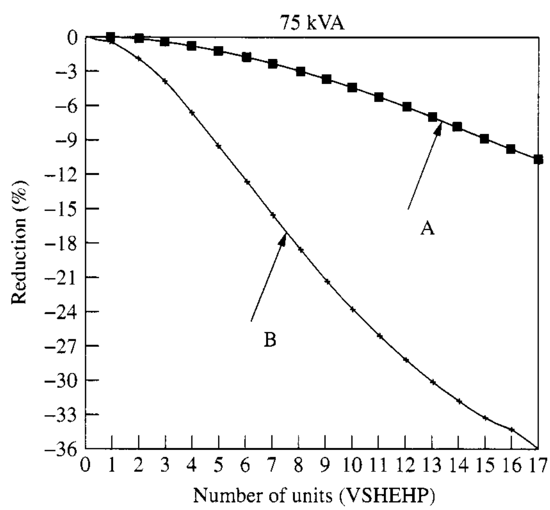

Figure 4.9 Reduction of the nominal capacity of a 75 kVA distribution transformer feeding two types of heat pump

(ii) A 75 kVA single-phase transformer feeds different proportions of heat pumps (VSHEHP) [10], involving rectification with high harmonic content (shown in Table 4.1). The derated levels, using the ANSI-C57.110 recommendation, are plotted in Figure 4.9 for different numbers of pump units.

4.4.3 Capacitor Banks

The presence of voltage distortion increases the dielectric loss in capacitors, the total loss being expressed by

$$\sum_{n=1}^{\infty} C(\tan\delta)\omega_n V_n^2 \tag{4.9}$$

where $\tan\delta = R/(1/\omega C)$ is the loss factor, $\omega_n = 2\pi f_n$ and V_n is the r.m.s. voltage of the nth harmonic.

The additional thermal stress of capacitors directly connected to the system (i.e. without series inductance) is assessed approximately with the help of a special capacitor weighted THD factor defined as

$$\mathrm{THD}_C = \frac{\sqrt{\sum_{n=1}^{N} (n \cdot V_n^2)}}{V_1} \tag{4.10}$$

Series and parallel resonances (discussed in Section 4.2) between the capacitors and the rest of the system can cause overvoltages and high currents, thus increasing dramatically the losses and overheating of capacitors, and often leading to their destruction. Therefore all possible resonances must be taken into account in the design of power factor correction capacitors and in other applications, such as those used with single-phase induction motors and in converter transients damping. These capacitors are rated according to overcurrent limiting standards, such as ANSI/IEEE 18–1980, typical values being 15% in the UK, 30% in Europe and 80% in the USA.

Moreover, the total reactive power, including fundamental and harmonics, i.e.

$$Q = \sum_{n=1}^{N} Q_n$$

should not exceed the rated reactive power (taking into account the effect of permissible overvoltage and component manufacturing tolerances).

Power factor correction capacitors are often tuned to about the third or fifth harmonic frequency by adding a small series inductance (about 9% and 4%, respectively). This makes the capacitor look inductive to frequencies above the third (or fifth) harmonic and thus avoids parallel resonances.

4.5 Power Assessment with Distorted Waveforms

4.5.1 Single-Phase System

Early attempts to include waveform distortion in the power definitions were made by Budeanu [11] and Fryze [12].

Budeanu divided the apparent power into three orthogonal components, i.e.

$$S^2 = P^2 + Q_B^2 + D^2 \tag{4.11}$$

and defined the terms reactive power

$$Q_B = \sum_{l=1}^{n} V_l I_l \sin(\varphi_l) \tag{4.12}$$

accepted by the IEC and IEEE, and complementary power (which Budeanu called fictitious)

$$P_c = \sqrt{S^2 - P^2} \tag{4.13}$$

Fryze separated the current into two orthogonal components, i_a (active) and i_b (reactive)

$$i = i_a + i_b \tag{4.14}$$

and proposed for the reactive power the following definition:

$$Q_F = VI_b = \sqrt{S^2 - P^2} \tag{4.15}$$

To add some physical meaning to the matter, Shepherd and Zakikhani [13] proposed the following decomposition for the apparent power:

$$S^2 = S_R^2 + S_X^2 + S_D^2 \tag{4.16}$$

being

$$S_R^2 = \sum_1^n V_n^2 \sum_1^n I_n^2 \cos^2(\varphi_n) \tag{4.17}$$

$$S_X^2 = \sum_1^n V_n^2 \sum_1^n I_n^2 \sin^2(\varphi_n) \tag{4.18}$$

$$S_D^2 = \sum_1^n V_n^2 \sum_1^p I_p^2 + \sum_1^m V_m^2 \left(\sum_1^n I_n^2 + \sum_1^p I_p^2 \right) \tag{4.19}$$

In these expressions, S_R is said to be active apparent power, S_X reactive apparent power and S_D distortion apparent power.

The main advantage of this decomposition is that the minimisation of S_X immediately leads to the optimisation of the power factor via the addition of a passive linear element, a property not available when applying Budeanu's definition; however, there is no justification for a power decomposition with an active component that differs from the mean value of the instantaneous power over a period (i.e. the active power).

An alternative model also involving three components was proposed by Sharon [14]

$$S^2 = P^2 + S_Q^2 + S_C^2 \tag{4.20}$$

where

$$P \sim \text{ active power}$$

$$S_Q = V\sqrt{\sum_1^n I_n^2 \sin^2(\varphi_n)} \sim \text{ a reactive power in quadrature} \tag{4.21}$$

and

$$S_C \sim \text{ a complementary reactive power}$$

Similarly to the model of Shepherd and Zakikhani, the minimisation of S_Q in this model results in maximum power factor via the connection of linear passive elements. Moreover, Sharon replaces the questionable term S_R by the more acceptable active power P.

Taking into account that, in general, the main contribution to reactive power comes from the fundamental component of the voltage, Emanuel [15] proposed the following definitions:

$$Q_1 = V_1 I_1 \sin(\varphi_1) \tag{4.22}$$

and a complementary power

$$P_C^2 = S^2 - P^2 - Q_1^2 \tag{4.23}$$

Based on Fryze's theory, Kusters and Moore [16] proposed a power definition in the time domain with the current divided into three components:

- i_p — an active component with a waveform identical to that consumed in an ideal resistance
- i_{ql}/i_{qc} — a reactive component, corresponding to either a coil or a capacitor
- i_{qlr}/i_{qcr} — a residual reactive component, the remaining current after removing the active and reactive, i.e.

$$i_{qr} = i - i_p - i_q \tag{4.24}$$

Furthermore, Kusters and Moore suggested the following decomposition for the apparent power:

$$S = P^2 + Q_l^2 + Q_{lr}^2 = P^2 + Q_c^2 + Q_{cr}^2 \tag{4.25}$$

where $P = VI_p$ is the active power, $Q_l = VI_{ql}$ the inductive reactive power, $Q_c = VI_{qc}$ the capacitive reactive power, and Q_{lr}/Q_{cl} the remaining reactive powers obtained from equation (4.24).

Also based on Fryze's definition, Emanuel [17] proposed two alternative decompositions distinguishing between the power components of the fundamental frequency (P_1, Q_1) and harmonics (P_H, Q_H), i.e.

$$S^2 = (P_1 + P_H)^2 + Q_F^2 \tag{4.26}$$

being

$$Q_F^2 = Q_B^2 + D^2 \tag{4.27}$$

and

$$S^2 = (P_1 + P_H)^2 + Q_1^2 + Q_H^2 \tag{4.28}$$

where

$$Q_H^2 = Q_F^2 - Q_1^2 \tag{4.29}$$

Further articles elaborating on the decomposition of power under non-sinusoidal conditions have been published by Czarnecki [18] and Slonin and Van Wyk [19].

In a recent contribution, Emanuel [20] makes the following statements:

- All forms of non-active powers stem from energy manifestations that have a common mark: energy oscillations between different sources, sources and loads or loads and loads. The net energy transfer linked with all the non-active powers is nil.
- Due to the unique significance of the fundamental powers, S_1, P_1 and Q_1, that comes from the fact that electric energy is a product expected to be generated and delivered and bought in the form of a 60 or 50 Hz electromagnetic field, it is useful to separate the apparent power S into fundamental S_1 and the non-fundamental S_N apparent powers:

$$S^2 = S_1^2 + S_N^2. \tag{4.30}$$

- As well as all the non-active powers, the term S_N contains also a minute amount of harmonic active power, P_H. The harmonic active power rarely exceeds $0.005P_1$. Thus, in a first approximation, an industrial nonlinear load can be evaluated from the measurements of P_1, Q_1 and S_N.
- The further subdivision of S_N into other components provides information on the required dynamic compensator or static filter capacity and level of current and voltage distortion.

Illustrative Example [21] To illustrate the practical consequences of using each of the above definitions, a comparative test study is shown with four single-phase networks with identical r.m.s. values of voltage (113.65 V) and current (16.25 A). In case A, the applied voltage is sinusoidal and the load nonlinear; case B contains voltage and current of the same, non-sinusoidal, waveform and in phase with each other; in case C the two identical waveforms are out of phase, and finally in D the voltage and current have harmonics of different orders.

The numerical information corresponding to these four cases is listed in Table 4.2 and the time-domain variation of the voltages and currents is shown in Figure 4.10.

Table 4.3 lists, for each case, the values of the powers previously defined, i.e. active P, reactive Q_B, apparent reactive S_X, reactive in quadrature Q_S, reactive of fundamental component Q_1, Fryze's reactive Q_F, distortion D, apparent S and also the power factor PF.

Table 4.2 Voltage and current phasors $V = 113.65$ V, $I = 16.25$ A

Case		$V_1\angle\alpha_1$	$V_3\angle\alpha_3$	$V_5\angle\alpha_5$	$V_7\angle\alpha_7$
A	v_A	$113.65\angle 0°$			
	i_A	$15\angle -30°$	$5.8\angle 0°$	$2\angle 0°$	$1\angle 0°$
B	v_B	$105\angle 0°$	$35\angle 0°$	$21\angle 0°$	$15\angle 0°$
	i_B	$15\angle 0°$	$5\angle 0°$	$3\angle 0°$	$(15/7)\angle 0°$
C	v_C	$105\angle 0°$	$35\angle 0°$	$21\angle 0°$	$15\angle 0°$
	i_C	$15\angle -30°$	$5\angle -90°$	$3\angle -150°$	$(15/7)\angle 150°$
D	v_D	$105\angle 0°$	$40.82\angle 180°$		$15\angle 0°$
	i_D	$15\angle -30°$		$5.44\angle -60°$	$3\angle -30°$

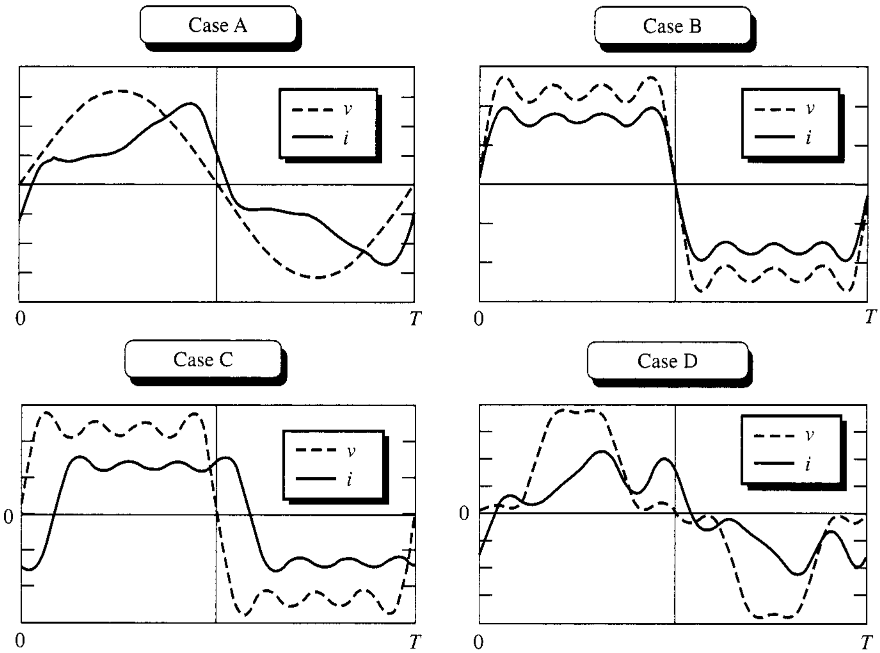

Figure 4.10 Voltage and current waveforms

Table 4.3 Alternative powers for four circuits with the same r.m.s. voltage and current but with different waveforms

Magnitude		Case A	Case B	Case C	Case D
Active	$P(W)$	1476	1845	1282	1403
Budeanu	$Q_B(Var)$	852	0	978	810
Shepherd	$S_X(VA)$	852	0	1046	811
Sharon	$Q_S(Var)$	852	0	1046	870
Emanuel	$Q_I(Var)$	852	0	788	788
Fryze	$Q_F(Var)$	1107	0	1327	1198
Budeanu	$D(VA)$	707	0	897	883
Apparent	$S(VA)/PF$	1845/0.80	1845/1	1845/0.695	1845/0.76

The rest of this section describes the results of three different comparisons made between cases A, C and D.

The first method, type 1 in Table 4.4, relates to the conventional capacitive compensation, and the table indicates the value of the capacitor that optimises the power factor $PF_{\max 1}$ and the required apparent power $S_{\min 1}$.

The second method, type 2, consists of a sinusoidal current injection of fundamental frequency that cancels the reactive current i_{r1} in the load; the corresponding apparent power and power factor are $S_{\min 2}$ and $PF_{\max 2}$, respectively.

Finally, in type 3 compensation, the current injection not only cancels the reactive component i_{r1}, but also the harmonic currents; $S_{\min 3}$ and $PF_{\max 3}$, are the apparent power and power factor, respectively.

Table 4.4 $S_{\min}$ and $PF_{\max}$ values for different compensation techniques

	Compensation	Case A	Case B	Case C
Type 1	$S_{\min 1}(VA)/C(\mu F)/PF_{\max 1}$	1636/210/0.902	1693/98/0.757	1759/81/0.798
Type 2	$S_{\min 2}(VA)/PF_{\max 2}$	1636/0.92	1636/0.785	1636/0.858
Type 3	$S_{\min 3}(VA)/PF_{\max 3}$	1476/1	1476/0.869	1476/0.950

The more relevant comments resulting from the comparison are:

(1) When the voltage is sinusoidal and the current non-sinusoidal (case A), the load is nonlinear; in this case, all the reactive powers obtained are identical, except Fryze's, which includes the distortion component. With reference to the optimisation process, $PF_{\max 1} = PF_{\max 2}$, as could be expected considering the sinusoidal nature of the voltage.

(2) If the voltage and current are in phase and of the same waveform (case B) the instantaneous power never reaches negative values; therefore $P = S$, the reactive and distortion powers are zero and the power factor unity, the load equivalent impedance being a linear resistance.

(3) When the voltage and current have the same waveform but are out of phase with each other (case C) the reactive powers calculated according to the various definitions proposed are different. In this case, the passive circuit does not distort the current waveform but alters its phase such that the power factor increases from an initial value of 0.695 to an optimal value $PF_{\max 3}$ of 0.869.

(4) Case D, due to the load nonlinearity, contains different harmonics; the third harmonic is only present in the voltage and the fifth in the current waveforms, respectively. This is the most general case; all the reactive powers calculated are different and the maximum power factor ($PF_{\max 3}$) is 0.95.

(5) The reactive powers calculated according to the various definitions are all different; the smallest is Emanuel's, since he only considers the fundamental component, and the largest Fryze's, which also includes distortion. In general,

$$Q_1 \leq Q_B \leq S_X \leq Q_S \leq Q_F \tag{4.31}$$

4.5.2 Three-Phase System

Apparent power in unbalanced three-phase systems is currently calculated using several definitions that lead to different power factor levels. Consequently the power bills will also differ, due to the reactive power tariffs and, in some countries, to the direct registration of the maximum apparent power demand.

Four different expressions have been proposed for the apparent reactive power, two of them based on Budeanu's and the other two on Fryze's definitions [22]:

(i)
$$S_v = \sqrt{\left(\left(\sum_k P_k\right)^2 + \left(\sum_k Q_{bk}\right)^2 + \left(\sum_k D_k\right)^2\right)} \tag{4.32}$$

or vector apparent power;

(ii)
$$S_a = \sum_k \sqrt{(P_k^2 + Q_{bk}^2 + D_k^2)} \tag{4.33}$$

or arithmetic apparent power.

In the above expressions P_k represents the active power, Q_{bk} Budeanu's reactive power and D_k the distortion power in phase k; these terms are generally accepted by the main international organisations, such as the IEEE and IEC.

(iii)
$$S_e = \sum_k \sqrt{(P_k^2 + Q_{fk}^2)} = \sum_k V_k I_k \tag{4.34}$$

an apparent r.m.s. power, that considers independently the power consumed in each phase.

(iv)
$$S_s = \sqrt{(P^2 + Q_f^2)} = \sqrt{\sum_k V_k^2}\sqrt{\sum_k I_k^2} \tag{4.35}$$

a system apparent power, that considers the three-phase network as a unit.

The last two expressions use the reactive powers as defined by Fryze, Q_f and Q_{fk}, which include not only the reactive but also the distortive effects.

In the vector apparent power the phase reactive powers compensate each other, but not in the other expressions; the system apparent power calculates the voltage and current of each phase individually and therefore yields the highest apparent power. In general, the following applies:

$$S_v \le S_a \le S_e \le S_s \tag{4.36}$$

The power factor of a load or system is generally accepted as a measure of the power transfer efficiency and is defined as the ratio between the electric power transformed into some other form of energy and the apparent power, i.e.

$$PF = P/S \tag{4.37}$$

Correspondingly, the relative magnitudes of the power factors calculated from the different definitions are

$$PF_v \ge PF_a \ge PF_e \ge PF_s \tag{4.38}$$

where PF_v, PF_a, PF_e and PF_s are the power factors corresponding to the vector, arithmetic, r.m.s. and system apparent powers, respectively.

For the power factor to reflect the system efficiency in three-phase networks with neutral wire, the neutral (zero sequence) currents must be included in the calculation of the equivalent current, i.e. in equation (4.35):

$$I_k^2 = I_a^2 + I_b^2 + I_c^2 + I_n^2 \tag{4.39}$$

and

$$V_k^2 = V_a^2 + V_b^2 + V_c^2 \tag{4.40}$$

where a, b, c indicate the individual phase values and n the neutral.

Illustrative Example [23,24] The simple test system shown in Figure 4.11 is used to illustrate the different performance of the proposed three-phase power definitions. Asymmetrical three-phase networks, even with sinusoidal voltage excitation and linear resistive loading, interchange reactive energy between the generator phases, despite the absence of energy storing elements. To show this effect, the circuit of Figure 4.11 is further simplified by making the line resistances equal to zero and assuming that the load is purely resistive, i.e. $Z_a = R_a$, $Z_b = R_b$, $Z_c = R_c$. Moreover, the three-phase source is assumed to be balanced and sinusoidal.

The following different operating conditions are compared in Table 4.5.

Case i	$R_a = 0\ \Omega$	$R_b = 5\ \Omega$	$R_c = 25\ \Omega$	without neutral wire
Case ii	$R_a = 5\ \Omega$	$R_b = 5\ \Omega$	$R_c = 25\ \Omega$	without neutral wire
Case iii	$R_a = 5\ \Omega$	$R_b = 5\ \Omega$	$R_c = 25\ \Omega$	with neutral wire
Case iv	$R_a = 5\ \Omega$	$R_b = 5\ \Omega$	$R_c = 5\ \Omega$	without neutral wire

As shown in Table 4.5, in case i, phases a and c generate reactive power while phase b absorbs reactive power, even though the overall reactive power requirement is zero. The effect is attenuated when a load is connected to phase a (case ii), but

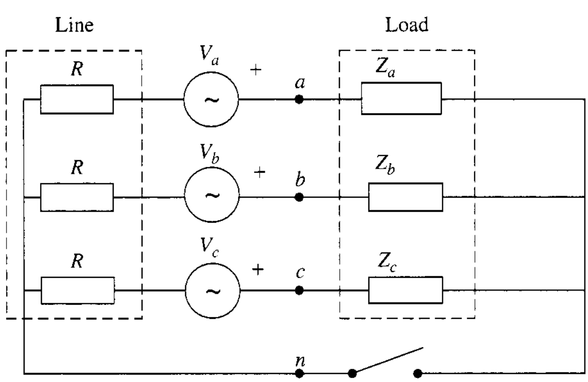

Figure 4.11 Three-phase test system

Table 4.5 Apparent powers and power factors for test cases i to iv

Case	SV_a	SV_b	SV_c	SV	$S_a = S_e$	S_s	PF_v	$PF_a = PF$	PF_s
i	32.4 − j12.46	27 + j15.58	5.4 − j3.11	64.8	72.12	81.52	1	0.898	0.794
ii	14.7 − j5.7	14.7 + j5.7	4.9	34.36	36.47	39.58	1	0.942	0.868
iii	18	18	3.6	39.6	39.6	44.53	1	1	0.889
iv	18	18	18	54	54	54	1	1	1

*All the powers are expressed in kVA.

there is still reactive power in two phases. When the circuit topology is changed by connecting a neutral wire (case iii) the generating source stops generating reactive power; the reactive power generation is also absent when phase c is made equal to phases a and b (case iv), i.e. when the network is perfectly balanced, and this is the case of maximum efficiency.

Table 4.5 shows for each case the apparent powers resulting from the different definitions as well as the corresponding power factors. The table also shows the complex apparent powers generated by the sources V_a, V_b, V_c. It can be seen that the most pessimistic power factor is PF_s, which only gives the value of one when the resistive load is perfectly balanced; on the other hand PF_v remains unity in all cases since the total load demand for reactive power, being resistive, is zero.

Next, a set of test cases v to x involve the circuit of Figure 4.11 with a perfectly balanced source feeding either linear or nonlinear loads.

The calculations performed in each case involve the four power factors defined above, the resistive losses in the line, and the line loss ratios of each of the cases to that of case v (the balanced case), which is used as a reference; the latter will show that only one of the definitions coincides with that ratio.

As the powers consumed by the load are different in all cases, for the purpose of comparison, the power of case v is also used as a reference, i.e. 54 kW in the test system.

Case v: V_{ns} (non-sinusoidal voltages), balanced load, 3 or 4 wires

$$
\begin{aligned}
R_a &= R_b = R_c = 5\ \Omega,\ R = 0.2\ \Omega/\text{phase} \\
v_a &= 298.51\sqrt{2}\sin(\omega t) + 29.85\sqrt{2}\sin(5\ \omega t);\quad V_a = 300\ \text{V} \\
i_a &= 59.70\sqrt{2}\sin(\omega t) + 5.97\sqrt{2}\sin(5\ \omega t);\quad I_a = 60\ \text{A} \\
P &= 3 \times 298.51 \times 59.70 + 3 \times 29.85 \times 5.97 = 54\,000\ \text{W}
\end{aligned}
$$

The power loss in the line is $P_v = 3 \times (60)^2 \times 0.2 = 2160$ W. It should be noted that this value is the same as case iv (the balanced sinusoidal circuit) for the same line resistance.

Case vi(a): V_{ns}, unbalanced load, 4 wires

$$
\begin{aligned}
R_a &= R_b = 5\ \Omega,\ R_c = 25\ \Omega \\
v_a &= 298.51\sqrt{2}\sin(\omega t) + 29.85\sqrt{2}\sin(5\ \omega t) \\
i_a &= 59.70\sqrt{2}\sin(\omega t) + 5.97\sqrt{2}\sin(5\ \omega t);\quad I_a = 60\ \text{A} \\
i_b &= 59.70\sqrt{2}\sin(\omega t - 120) + 5.97\sqrt{2}\sin(5\ \omega t + 120);\quad I_b = 60\ \text{A} \\
i_c &= 11.94\sqrt{2}\sin(\omega t + 120) + 1.19\sqrt{2}\sin(5\ \omega t - 120);\quad I_c = 12\ \text{A}
\end{aligned}
$$

The topology of this case does not permit distorted or reactive power. To calculate the line loss in relation to the power base, the calculated currents are multiplied by factor K, which is the ratio of the power in case v to that of vi(a); therefore

$$K = 54\,000/39\,600 = 1.36$$

$$P_{vi} = [(60 \times 1.3636)^2 + (60 \times 1.3636)^2 + (12 \times 1.3636)^2] \times 0.2 = 2731.23\ \text{W}$$

The ratio $P_v/P_{vi} = 0.79$ is the square of PF_s. It is observed that although the consumed power has no reactive or distorted component, due to load unbalance, line losses are larger and the power factors PF_v, PF_a and PF_e do not reflect that loss of network efficiency; on the other hand PF_s represents faithfully this increment. It should be noted that the value P_{iii} (of case iii) would be identical to P_{vi} if the line resistance (i.e. 0.2 Ω/phase) had been represented.

Case vi(b): V_{ns}, unbalanced load, 3 wires

$$R_a = R_b = 5\ \Omega,\ R_c = 25\ \Omega,\ R = 0.2\ \Omega/\text{phase}$$
$$v_a = 298.51\sqrt{2}\sin(\omega t) + 29.85\sqrt{2}\sin(5\ \omega t)$$
$$i_a = 52.34\sqrt{2}\sin(\omega t + 21.05) + 5.234\sqrt{2}\sin(5\ \omega t - 21.05);\quad I_a = 52.60\ \text{A}$$
$$i_b = 52.34\sqrt{2}\sin(\omega t - 141.05) + 5.234\sqrt{2}\sin(5\ \omega t + 141.05);\quad I_b = 52.60\ \text{A}$$
$$i_c = 16.34\sqrt{2}\sin(\omega t + 120) + 1.628\sqrt{2}\sin(5\ \omega t - 120);\quad I_c = 16.36\ \text{A}$$

The different behaviour in this case with respect to reactive and distorted power is due to the new topology, which excludes the neutral wire.

The line losses, normalised to 54 kW, are $P_{vi}(b) = 2865.3$ and the ratio $P_v/P_{vi} =$ 0.75 is again the square of PF_s.

Case vii: V_s (sinusoidal), nonlinear load, I_{ns} (non sinusoidal) and balanced

$$R = 0.2\ \Omega/\text{phase}$$
$$v_a = 300\sqrt{2}\sin(\omega t).$$
$$i_a = 60\sqrt{2}\sin(\omega t) + 6\sqrt{2}\sin(5\ \omega t);\quad I_a = I_b = I_c = 60.3\ \text{A}$$

In this case the load nonlinearity produces a harmonic component not present in the voltage source. The power consumed is 54 kW and the values of the various power factors are identical and close to unity, in spite of the fifth harmonic current, because the fundamental component of the current is balanced and its power factor is unity.

Case viii: V_s, nonlinear load, I_{ns} unbalanced

$$R = 0.2\ \Omega/\text{phase}$$
$$v_a = 300\sqrt{2}\sin(\omega t).$$

This case has the same currents as in vi(b), yielding lower power factors, due to the fact that the load active power decreases in the absence of fifth harmonic voltage.

Case ix: V_{ns}, nonlinear load, I_{ns} balanced

$$R = 0.2\ \Omega/\text{phase}$$
$$v_a = 298.51\sqrt{2}\sin(\omega t) + 29.85\sqrt{2}\sin(5\ \omega t)$$
$$i_a = 59.70\sqrt{2}\sin(\omega t) + 5.97\sqrt{2}\sin(5\ \omega t) + 10\sqrt{2}\sin(7\ \omega t);\quad I_a = 60.83\ \text{A}$$

In this case the currents are as in v, but the load nonlinearity injects seventh harmonic, which results in a reduction of power factors with respect to the base case.

Case x: V_{ns}, nonlinear load, I_{ns} unbalanced

$$R = 0.2\ \Omega/\text{phase}$$
$$v_a = 298.51\sqrt{2}\sin(\omega t) + 29.85\sqrt{2}\sin(5\ \omega t)$$
$$i_a = 59.70\sqrt{2}\sin(\omega t);\quad I_a = 59.70\ \text{A}$$
$$i_b = 59.70\sqrt{2}\sin(\omega t - 120) + 5.97\sqrt{2}\sin(5\ \omega t + 120);\quad I_b = 60\ \text{A}$$
$$i_c = 59.70\sqrt{2}\sin(\omega t + 120) + 11.94\sqrt{2}\sin(5\ \omega t - 120);\quad I_b = 60.88\ \text{A}$$

Table 4.6 Circuit characteristics for test cases v to x

Case	Volt components	Impedance	Neutral	Current components
v	ωi, $\omega 5$	$R_a = R_b = R_c = 5\ \Omega$	No	$\omega 1$, $\omega 5$, balanced
vi(a)	ωi, $\omega 5$	$R_a = R_b = 5$; $R_c = 25\ \Omega$	Yes	$\omega 1$, $\omega 5$, unbalanced
vi(b)	$\omega 1$, $\omega 5$	$R_a = R_b = 5$; $R_c = 25\ \Omega$	No	$\omega 1$, $\omega 5$, unbalanced
vii	$\omega 1$	Nonlinear	Yes	$\omega 1$, $\omega 5$, balanced
viii	$\omega 1$	Nonlinear	Yes	$\omega 1$, $\omega 5$, unbalanced
ix	$\omega 1$, $\omega 5$	Nonlinear	Yes	$\omega 1$, $\omega 5$, balanced
x	$\omega 1$, $\omega 5$	Nonlinear	Yes	$\omega 1$, balanced; $\omega 5$, unbalanced

Table 4.7 Apparent powers and power factors of test cases v to x

Case	S_v	$S_a = S_e$	S_s	PF_v	$PF_a = PF_e$	PF_s	P_j	P_v/P_j	PF_s^2
v	54 000	54 000	54 000	1	1	1	2160	1	1
vi(a)	39 600	39 600	44 529.5	1	1	0.889	2731.2	0.790	0.790
vi(b)	34 436.8	36 470.1	39 578.3	0.997	0.942	0.868	2865.3	0.753	0.753
vii	54 269.3	54 269.3	54 269.3	0.995	0.995	0.995	2181.6	0.990	0.990
viii	34 385.1	36 470.1	39 578.3	0.994	0.937	0.863	2893.9	0.746	0.746
ix	54 744.8	54 744.8	54 744.8	0.986	0.986	0.986	2220	0.972	0.972
x	54 117.5	54 176.0	54 177.9	0.997	0.996	0.996	2174.2	0.993	0.993

The fundamental component is as in case v; the nonlinear load contains unbalanced fifth harmonic, chosen to ensure that the consumed active power is the same as that of the base circuit; this case also shows a decrease of the power factors with respect to the base case.

The main characteristics of the seven cases considered in this section are shown in Table 4.6. Table 4.7 illustrates, for each case, the magnitudes of the apparent powers; P_j is the power loss in the line, normalised to the base power of case v, P_v/P_j is the ratio of the line power loss for the balanced line and that corresponding to each case. Finally, the table lists the magnitude of PF_s (the square of the system power factor), which coincides always with P_v/P_j.

4.5.3 Power Factor Under Harmonic Distortion [25]

In general, the instantaneous values of the voltage and current components can be expressed as

$$v = \sum_{1}^{n} \sqrt{2} V_n \sin(n\omega t + \alpha_n) + \sum^{m} \sqrt{2} V_m \sin(m\omega t + \alpha_m) \tag{4.41}$$

$$i = \sum_{1}^{n} \sqrt{2} I_n \sin(n\omega t + \alpha_n + \phi_n) + \sum^{p} \sqrt{2} I_p \sin(p\omega t + \alpha_p) \tag{4.42}$$

and the power factor is given by

$$\text{p.f.} = \frac{1/T\int_0^T vi\,\mathrm{d}t}{V_{\text{rms}}I_{\text{rms}}} = \frac{\sum_1^n V_n I_n \cos(\phi_n)}{\left\{\left(\sum_1^n V_n^2 + \sum^m V_m^2\right)\left(\sum_1^n I_n^2 + \sum^p I_p^2\right)\right\}^{1/2}} \tag{4.43}$$

This factor represents a figure of merit of the character of the power consumption. A low value indicates poor utilisation of the source-power capacity needed by the load.

If the voltage waveform is sinusoidal, equation (4.43) reduces to

$$\text{p.f.} = \frac{V_1 I_1 \cos(\phi_1)}{V_1 I_{\text{rms}}} = \frac{I_1}{I_{\text{rms}}}\cdot\cos(\phi_1) = \mu\cos(\phi_1) \tag{4.44}$$

where $\cos(\phi_1)$ is the displacement power factor (DPF) between the fundamental components of voltage and current, and μ is a current distortion factor. It is worth mentioning that only DPF information is available from conventional instrumentation.

Thus unity power factor can only be achieved when $\mu = 1$ since $\cos(\phi_1)$ in equation (4.44) can not be greater than one.

Power factor compensation is not straightforward with distorted waveforms. As lossless devices are normally used for the compensation, the minimisation of the apparent power should lead directly to the optimum power factor. If, for example, a capacitance C is added in parallel to the load characterised by equations (4.41) and (4.42), the general expression for the apparent power in terms of C is

$$S = \left(\sum_1^n V_n^2 + \sum^m V_m^2\right)^{1/2} \cdot \left\{\sum_1^n (I_n^2 + V_n^2 n^2\omega^2 C^2 + 2V_n I_n n\omega C\sin(\phi_n)) + \sum^m V_m^2 m^2\omega^2 C^2 + \sum^p I_p^2\right\}^{1/2} \tag{4.45}$$

The differentiation of this equation with respect to C and its subsequent equating to zero leads to an optimum value of the linear capacitance, i.e.

$$C_{\text{opt}} = -\frac{1/\omega\sum_1^n V_n n I_n \sin(\phi_n)}{\sum_1^n V_n^2 n^2 + \sum^m V_m^2 m^2} \tag{4.46}$$

The object of capacitor compensation is to improve the displacement factor if the voltage is sinusoidal. Improvement in the values of distortion factor are achieved by filters, higher pulse numbers, or current waveform modification. These techniques are discussed in Chapter 6.

4.5.4 Effect of Harmonics on Measuring Instruments

Measuring instruments initially calibrated on purely sinusoidal alternating current and subsequently used on a distorted electricity supply can be prone to error. The magnitude and direction of the harmonic power flow are important for revenue considerations as the sign of the meter error is decided by the direction of flow.

Studies have shown that errors due to harmonic content vary greatly as to the type of meter, and that both positive and negative metering errors are possible.

The classic energy measuring instrument is the Ferraris motor type kilowatt-hour meter. Its inherent design is electromagnetic, producing driving and braking fluxes which impinge on its rotor, developing a torque. Secondary flux-producing elements are provided for compensation purposes to improve the instrument accuracy and to compensate for friction in the register. These flux-producing elements, providing primary and secondary torques, are essentially nonlinear in regard to amplitude and frequency. The nonlinear elements include the voltage and current elements and overload magnetic shunts, and the frequency-sensitive elements include the disc, the quadrature and anti-friction loops.

The response of this meter to frequencies outside the design parameter is inefficient, and large inaccuracies result. An expression for total power as seen by a meter is

$$\underset{(P_\mathrm{T})}{\text{Total power}} = \underset{(P_\mathrm{dc})}{V_\mathrm{dc} I_\mathrm{dc}} + \underset{(P_\mathrm{F})}{V_F I_F \cos\phi_F} + \underset{(P_\mathrm{H})}{V_H I_H \cos\phi_H} \tag{4.47}$$

The meter will not measure P_dc but will be sensitive to its presence; it will measure P_F accurately and P_H inaccurately, the error being determined by the frequency. The total harmonic power P_H is obtained by adding all components derived from the products of voltages and currents of the same frequencies, both above and below the fundamental frequency.

Any d.c. power supplied to or generated by the customer will cause an error proportional to the power ratio $P_\mathrm{dc}/P_\mathrm{T}$, with the error sign related to the direction of power flow. Similarly, any deficiency in measuring harmonic power P_H will cause an error represented by $\pm KP_\mathrm{H}/P_\mathrm{T}$, where the factor K is dependent on the frequency-response characteristics of the meter, and the error sign again will be related to power flow direction.

D.c. power and harmonic voltages or currents alone should not produce torques, but will degrade the capability of a meter to measure fundamental frequency power. Direct currents distort the working fluxes and alter the incremental permeability of the magnetic elements. Fluxes produced by harmonic currents combine with spurious fluxes of the same frequency that may be present due to the imperfection of the meter element and produce secondary torques.

The kilowatt-hour meter, based on the Ferraris (eddy current) motor principle, has been found generally to read high to the extent of up to several percentage points with a consumer generating harmonics through thyristor-controlled variable speed equipment (particularly if even harmonics and d.c. are involved) and notably if there is also a low power factor.

Converter loads using the 'burst firing' principle can cause kilowatt-hour meters to read high by several percentage points (cases in excess of 6% have been quoted), largely attributable to the lack of current damping during the no-load interval.

It appears that consumers that generate harmonics are automatically penalised by a higher apparent electricity consumption, which may well offset the supply authority's additional losses [26]. It is therefore in the consumer's own interest to reduce harmonic generation to the greatest possible extent.

There is no evidence that the reading of kVA-demand meters is affected by network harmonics. However, kW-demand meters operating on the time interval Ferraris motor principle will read possibly a few percentage points high, as shown earlier for energy meters.

Harmonics present a problem to measurement of VAR values, since this is a quantity defined with respect to sinusoidal waveforms.

The present trend is to use solid state instruments which can measure true power irrespective of the waveforms. Modern r.m.s. responding voltmeters and ammeters are relatively immune to the influences of waveform distortion. In such meters, the input voltage or current is processed using an electronic multiplier type, such as variable transconductance, log/antilog, time division, and thermal and digital sampling. All these can be configured to respond to the r.m.s. value independently of the harmonic amplitude or phase, as long as the harmonics are within the operating bandwidth of the instrument and the crest factor (the ratio of peak to r.m.s.) of the waveform is not excessively large. However, absolute average and peak responding meters which are calibrated in r.m.s. are not suitable in the presence of harmonic distortion.

A seldom mentioned effect, but nevertheless an important one, is that measurement and calibration laboratories, working to small tolerances of accuracy, may lose confidence in their results.

4.6 Harmonic Interference with Ripple Control Systems

Ripple signals are often used for the remote control of street lighting circuits and for load reduction (such as domestic hot water heaters) during peak times of the day.

Electricity suppliers have in the past experienced some practical difficulties with their ripple control equipment as a result of harmonic interference.

Since ripple relays are essentially voltage-operated (high-impedance) devices, harmonic interference can cause signal blocking or relay maloperation if present in sufficient amplitude. The exact amplitude at which the voltage harmonic will affect the relay is a function of the relay detection circuit (sensitivity and selectivity) and the proximity of the ripple injection frequency of the interfering harmonic.

Signal blocking occurs when sufficient interfering voltage renders the relay unable to detect the presence of the signal. Capacitors can produce the same effect due to their capability to absorb the ripple signal. Relay maloperation occurs when the presence of the harmonic voltage (usually in the absence of the signal) causes the relay to change state. The latter problem has effectively been solved by the use of suitably encoded switching signals in present generations of ripple relays.

Earlier ripple relays were electromechanical devices employing mechanical filter assemblies. Although their response was slow, they did achieve very good selectivity. However, these relays commonly suffered from maloperation because they had inadequate signal encoding to cater for any harmonic interference which got pass the filters.

However, modern ripple relays are basically electronic equivalents of their electromechanical forerunners. They usually employ piezoelectric or active filter circuits and a high degree of signal encoding to minimise maloperation. They have filter response curves defined by international standards [27] and they require the use of precision electronic components (high stability and reliability) to ensure a satisfactory performance over the life of the relay. Even better immunity can be achieved with the use of digital filtering techniques, which are ideally suited for the detection of signals of a specified frequency in the presence of harmonic voltages.

4.7 Harmonic Interference with Power System Protection

Harmonics can distort or degrade the operating characteristics of protective relays depending on the design features and principles of operation. Digital relays and algorithms that rely on sample data or zero crossings are particularly prone to error when harmonic distortion is present.

In most cases, the changes in operating characteristics are small and do not present a problem. Early tests [28] indicate that for most types of relays operation is not affected significantly for harmonic voltage levels of less than 20%. Most studies carried out so far conclude that it is difficult to predict relay performance without testing; the studies published have evaluated electromechanical and electronic relays but there is no information on digital relays [29]. However with the increased content of large power conversion equipment, this can be a potential problem.

Current harmonic distortion can also affect the interruption capability of circuit breakers and fuses. Possible reasons are higher di/dt at zero crossings, the current sensing ability of thermal magnetic breakers and a reduction in the trip point due to extra heating of the solenoid. The fuses, being thermally activated, are inherently r.m.s. overcurrent devices; the fuse ribbons are also susceptible to the extra skin effect of the harmonic frequencies.

4.7.1 Harmonic Problems During Fault Conditions

Protective functions are usually developed in terms of fundamental voltages and/or currents, and any harmonics present in the fault waveforms are either filtered out or ignored altogether. The latter is particularly the case for electromagnetic relay applications, such as overcurrent protection. Electromechanical relays have significant inertia associated with them such that often they are inherently less sensitive to higher harmonics.

More important is the effect of harmonic frequencies on impedance measurement. Distance relay settings are based on fundamental impedances of transmission lines, and the presence of harmonic current (particularly third harmonic) in a fault situation could cause considerable measurement errors relative to the fundamental-based settings.

High harmonic content is common where fault current flows through high resistivity ground (i.e. the ground impedance is dominant) so the possibility of maloperation is great unless only the fundamental waveforms are captured.

In solid fault situations, the fundamental components of current and voltage are much more dominant (notwithstanding the d.c. asymmetry associated with fault waveforms).

However, because of current transformer saturation, secondary induced distortion of current waveforms, particularly with large d.c. offsets in the primary waveforms, will occur. The presence of secondary harmonics in such instances can be a real problem, i.e. whenever current transformer saturation occurs it is very difficult to recover the fundamental current waveform.

Whenever high secondary e.m.f. exists during steady-state conditions, the nonlinear current transformer exciting impedance only causes odd harmonic distortion. During saturation under transient conditions, however, any harmonics can be produced, with dominance of second and third harmonic components [30].

Fortunately, these are often design problems. Correct choice of equipment in relation to the system requirements can eliminate many of the difficulties associated with the current and voltage transformers.

Filtering of the current and voltage waveforms, particularly in digital protection systems, is of special importance to distance protection schemes. Although the implementation is not always simple, the recovery of fundamental frequency data has been greatly improved by the use of digital techniques [31].

4.7.2 Harmonic Problems Outside Fault Conditions

The effective insensitivity of protective apparatus to normal system load conditions implies that, generally, the harmonic content of the waveforms is not a problem outside fault conditions.

The most notable exemption is probably the problem encountered in energisation of power transformers. In practice, constructive use of the high harmonic content of magnetising inrush currents prevents (most of the time!) tripping of the high voltage circuit breaker by the transformer protection due to the excessively high peaks experienced during energisation.

The actual peak magnitude of the inrush current depends on the air-core inductance of the transformer and the winding resistance plus the point on the voltage wave at which switching occurs [32]. Residual flux in the core prior to switching also increases the problem or alleviates it slightly, depending on the polarity of flux with regard to the initial instantaneous voltage.

Since the secondary current is zero during energisation, the heavy inrush current would inevitably cause the differential protection to operate unless it is rendered inoperative.

The simple approach is to use a time-delay differential scheme, but this could result in serious damage to the transformer should a fault be present at energisation.

In practice, information on the uncharacteristic second harmonic component present during inrush is used to restrain the protection, but protection is still active should an internal fault develop during energisation.

4.8 Effect of Harmonics on Consumer Equipment

This is a broad subject discussed in many journal articles; a selected bibliography on the topic can be found in a paper by the IEEE Task Force on the Effects of Harmonics on Equipment [33]. A concise summary of the main effects is made below:

(1) *Television receivers* Harmonics which affect the peak voltage can cause changes in TV picture size and brightness. Inter-harmonics cause amplitude modulation of the fundamental frequency; even a 0.5% inter-harmonic level can produce periodic enlargement and reduction of the image of the cathode ray tube.

(2) *Fluorescent and mercury arc lighting* These appliances sometimes have capacitors which, with the inductance of the ballast and circuit, produce a resonant frequency. If this corresponds to a generated harmonic, excessive heating and failure may result. However, the resonant frequency of most lamps is in the range 75–80 Hz and should not interact with the power supply. Audible noise is another possible effect of harmonic voltage distortion.

(3) *Computers* [34] There are designer-imposed limits as to acceptable harmonic distortion in computer and data processing system supply circuits. Harmonic rate (geometric) measured in vacuum must be less than −3% (Honeywell, DEC) or 5% (IBM). CDC specifies that the ratio of peak to effective value of the supply voltage must equal 1.41 ± 0.1.

(4) *Power electronic equipment* Notches in the voltage waveform resulting from current commutations may affect the synchronisation of other converter equipment or any other apparatus controlled by voltage zeros. Harmonics could theoretically affect thyristor-controlled variable-speed drives of the same consumer in several ways: (i) voltage notching (causing brief voltage dips in the supply) can cause maloperation via a thyristor through misfiring; (ii) harmonic voltages can cause the firing of the gating circuits at other than the required instant; (iii) resonance effects between different equipment can result in over-voltages and hunting.

The problems described above could also be experienced by other consumers if connected to the same (415 V or 11 kV) busbar. Consumers without problems with the simultaneous operation of their own thyristor-controlled equipment are unlikely to interfere with other consumers. Consumers on different busbars could interfere with each other but 'electrical' remoteness (separation by impedances in the form of lines and transformers) will tend to reduce the problem.

4.9 Interference with Communications

Noise on communication circuits degrades the transmission quality and can interfere with signalling. At low levels noise causes annoyance and at high levels loss of information, which in extreme cases can render a communication circuit unusable.

The continuously changing power transmission environment demands regular reconsideration of the interference problem when telephone lines are placed in the vicinity of the power system.

The signal to noise ratio commonly used in communication circuits as a measure of the transmission quality must be used with caution when considering power system interference because of the different power levels of the signals involved, i.e. in megawatts (power circuit) and milliwatts (communication circuit). Thus even a small unbalanced audio-frequency component within the power network may easily produce unacceptable noise level when coupled into a metallic communication circuit.

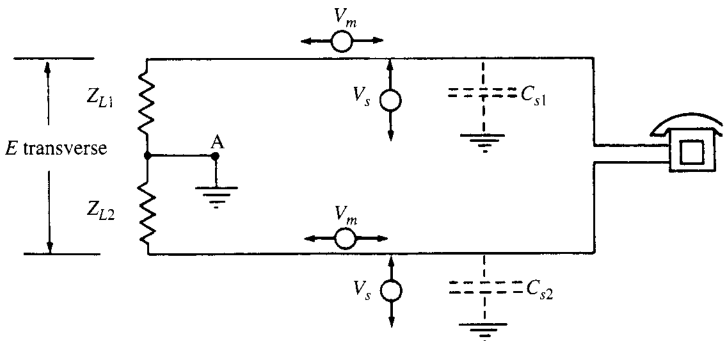

Figure 4.12 Simple model of a telephone circuit

Moreover, the purpose of the power system is to transmit energy at high efficiency but with relatively low waveform purity; on the other hand, in a communication circuit the waveform must not be significantly distorted, as the intelligence being conveyed may be destroyed, while the power efficiency is of secondary importance.

This section describes the factors influencing communication interference and the means by which noise may be reduced to within acceptable levels

4.9.1 Simple Model of a Telephone Circuit

A physical telephone circuit consists of a twisted pair of wires with associated terminal equipment, and a simplified model of such a circuit is illustrated in Figure 4.12.

For safety and practical reasons telephone circuits are referenced to earth and thus the equivalent circuit of the telephone system of Figure 4.12 includes the terminal impedances Z_{L1} and Z_{L2} to earth. An electromagnetic induced voltage is modelled as voltage source V_m, and an electrostatic induced voltage as V_s. The terminal impedances, Z_{L1} and Z_{L2}, are generally of high value and the telephone line self-impedance, being much smaller, may be neglected.

In the absence of an earth conductor the earth return circuit is completed by the stray capacitances C_{S1} and C_{S2}.

4.9.2 Factors Influencing Interference

Three factors combine to produce a noise problem on a communication line:

(1) *Power system influence* This depends on the source of audio-frequency components within the power system and the relative magnitude of unbalanced harmonic currents and voltages present in the power circuit in the vicinity of the communication circuit.

(2) *Coupling to communication circuits* This factor involves the coupling of interfering currents and voltages into a communication system.

(3) *Effect on communication circuits (susceptiveness)* The effect of the noise interference on a communication circuit is dependent on the characteristics of the circuit and associated apparatus.

All three factors must be present for the problem to develop. Complete elimination of the interference problem is usually impractical, and the degree of the problem will be a function primarily of the basic factors that have the highest influence.

Some harmonic currents in the power line are confined to the phase conductors, thus flowing in one phase and returning by the other two (the 'balanced' circuit); these are the positive- and negative-sequence harmonic currents. Other harmonic components flow in phase with each other in the phase conductors and return through the neutral or earth; these are the zero-sequence or residual currents.

4.9.3 Coupling to Communication Circuits

Coupling refers to the mutual impedance existing between the power and telephone lines. It depends on the separation between them, the length of exposure, earth resistivity and frequency.

Noise voltages may be impressed on telephone circuits in several ways, i.e. by loop induction, by longitudinal electromagnetic induction, by longitudinal electrostatic induction and by conduction.

Loop Induction Loop induction occurs when a voltage is induced directly into the metallic loop formed by the two wires of a telephone circuit. This type of induction manifests itself directly as a transverse voltage across the terminations of the telephone circuit. It is cancelled out by regular transpositions of aerial wires or by the use of twisted pairs of cables. As these are standard practices in communication circuits, loop induction is not generally a problem.

In the case of crossings, telephone lines often come so near the power line that the loop effect may be important [35].

Longitudinal Electromagnetic Induction Longitudinal electromagnetic induction occurs when an e.m.f. is induced along the conductors of a telephone circuit. The residual current in a power line sets up a magnetic field, which causes flux lines to intersect with any neighbouring telephone line and induces an e.m.f. longitudinally on it. This type of coupling, illustrated in Figure 4.13, constitutes the most common form of noise induction into communication lines.

For the close spacing of joint overhead transmission power and telephone lines, the magnetic coupling from balanced currents is important. For roadside spacings or greater, the earth return or residual current is the predominant source of interference.

The residual current I_R in Figure 4.13 returns via earth to V_p, hence the current loop so formed has a large cross-sectional area for overhead transmission lines. Likewise, aerial open wire telephone circuits may have large cross-sectional areas.

This leads to a longitudinal electromagnetic induction on telephone circuits given by

$$V_m = MI_R \tag{4.48}$$

where M is the mutual impedance between the power and telephone systems.

The most widely accepted model for determining the mutual impedance between power and telephone systems was developed by Carson [36].

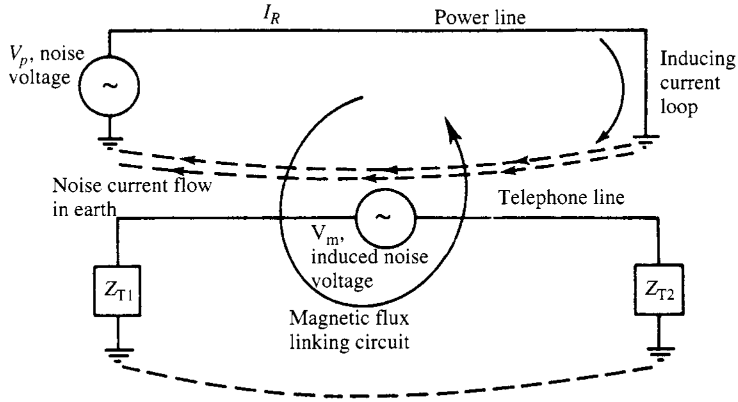

Figure 4.13 Electromagnetic induction

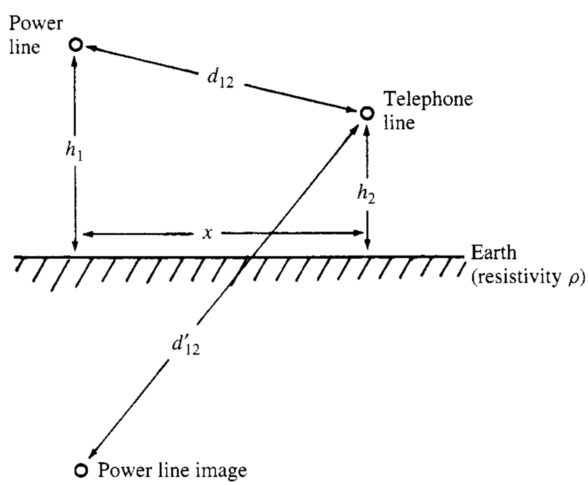

Figure 4.14 Power and telephone line configuration

In general, electromagnetic propagation at power frequencies is not very sensitive to earth structure and resistivity and so Carson's model (which assumes a uniform flat earth) is valid for most cases.

Carson's equation for the mutual impedance of a power line at height h_1 and a telephone line at height h_2, as in Figure 4.14, is given by

$$M = \frac{j\omega\mu_0}{2\pi}\left[\ln\frac{d'_{12}}{d_{12}} - 2j\int_0^{\infty}\left[\sqrt{(u^2+j)} - u\right]e^{-u\alpha(h_1+h_2)}\cos(u\alpha x)\,\mathrm{d}u\right] \tag{4.49}$$

where M is the mutual impedance per unit length, x is the horizontal separation of power and telephone lines, h_1 is the height of the power line above ground (which is negative if below ground), h_2 is the height of the telephone line above ground (which is again negative if below ground), $d_{12} = \sqrt{[(h_1 - h_2)^2 + x^2]}$ is the radial distance between lines, $d'_{12} = \sqrt{[(h_1 + h_2)^2 + x^2]}$ is the radial distance between one line and the underground image of the other, $\omega = 2\pi f$ is the angular frequency of the inducing current, μ_0 is the rationalised permeability of free space, $\alpha = \sqrt{2}/\delta$, $\delta = \sqrt{(2\rho/\mu_0\omega)}$, is the skin depth of uniform earth with resistivity ρ, and ρ is the resistivity of the earth in ohm-metres.

The first term of Carson's equation is the mutual impedance between two conductors as if they were above a perfectly conducting earth, while the second term 'corrects' for finite earth resistivity. Unfortunately, the second term usually dominates for typical parameters.

Carson's solution relies on measurements of the earth resistivity, which can be difficult to obtain, and simplified forms of Carson's equation are often used to obtain good approximations for the mutual impedance.

The factors influencing the mutual impedance between power and telecommunication lines can be summarised as follows: it increases with increasing values of the residual current loop area; it increases for increasing common distance run; it increases with frequency; it increases with earth resistivity; it decreases for increasing separation between circuits.

Longitudinal Electrostatic Induction Longitudinal electrostatic induction occurs when an e.m.f. is induced between the conductors and earth.

The simplest way of visualising electrostatic induction is by considering the capacitances in an exposure between a single power wire and a single telephone wire, as illustrated in Figure 4.15. The voltage of the power wire to ground (residual voltage), V_r, divides over the capacitance between the power and telephone wire, C_{PT} and the telephone wire and ground, C_{TC} in the ratio of their impedances, i.e.

$$V_s = \frac{Z_{\mathrm{T}}}{1/(j\omega C_{\mathrm{PT}}) + Z_{\mathrm{T}}} \cdot V_r \tag{4.50}$$

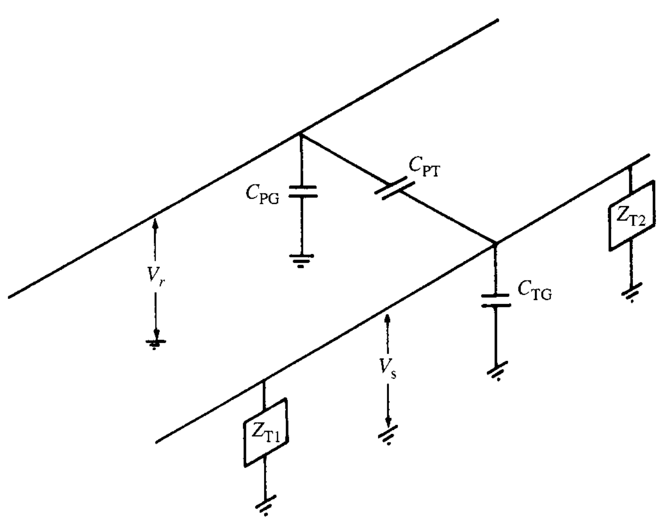

Figure 4.15 Electrostatic induction

where

$$Z_{\mathrm{T}} = \frac{1}{j\omega C_{\mathrm{TG}} + (1/Z_{\mathrm{T1}}) + (1/Z_{\mathrm{T2}})}$$

Because of the loading effect of Z_{T1} and Z_{T2} and the relative separations, V_s is very small as compared with V_m and can easily be neutralised by cable screening.

Electrostatic induction is serious only when the residual voltage, V_r is large (e.g. single-wire power lines) or when C_{PT} is large (for example, in joint construction and crossings, where the two lines are very close together). Generally the telephone line is terminated in impedances which are small compared to the capacitive impedances and thus reduce the induced voltage, V_s. This form of induction is often a problem on long telephone lines in the neighbourhood of very high voltage transmission lines.

For instance, for the case of roadside separation between the power and communication lines, with an earth resistivity of 100 metre-ohms, the 60 Hz mutual impedance is about one half ohm per mile. The longitudinal electric field in a telephone circuit on the same right of way as the power line is about 10 V per mile. Therefore five miles of exposure would induce 50 V on the telephone circuit.

Conductive Coupling There is always some residual current flowing in the neutral of the power system due to out-of-balance components. With a multiple earthed neutral (MEN) system, some of this residual current will return to the transformer by the neutral wire, and some via earth. The earth currents will cause a local rise of earth potential at the earth electrode.

If one end of a telephone line is earth referenced in the area of influence of this earth potential rise, then a longitudinal voltage may be impressed on the line.

In the circuit of Figure 4.12, if a local earth potential rise occurred at A this would cause unequal currents to flow through Z_{L1} and Z_{L2}, giving rise to a transverse noise voltage in the telephone circuit.

This source of interference is an increasing problem due to the following factors:

(1) MEN earth systems are carrying higher levels of noisy current.

(2) The earth resistances of telephone exchange earth systems are increasing as a result of the use of less lead armoured cable; it is very costly to achieve low earth resistance.

(3) Despite (2), the telephone exchange earth is often a relatively low impedance earth return circuit for a MEN earth system feeding the exchange, and hence a considerable noise voltage can be impressed onto the earth system from a MEN system.

4.9.4 Effect on Communication Circuits (Susceptiveness)

Telephone Circuit Susceptiveness A voice band telephone channel is normally designed to pass frequencies between 300 and 3000 Hz. Although harmonics in this range of frequencies are very small compared with the fundamental, they still have an effect on the telephone reception.

The effect that a 'noisy' power line will have on a communication line may be ascertained by considering the susceptiveness of the circuit to the effects of inductive interference. Three characteristics are of importance in this respect: (i) the relative interfering effects of different frequencies; (ii) the balance of the communication circuit; and (iii) shielding effects of metallic cable sheaths and other buried metallic plant.

Harmonic Weights Standard weighting curves are used to take into account the response of the telephone equipment and the sensitivity of the human ear to the harmonic frequencies. Two weighting factors are in common use:

(1) the psophometric weighting by the CCITT [37], extensively used in Europe;
(2) the C-message weighting by Bell Telephone Systems (BTS) and Edison Electric Institute (EEI) [38], used in the USA and Canada.

Figure 4.16 shows that the difference between these two weighting curves (when normalised) is very slight and that the human ear in combination with a telephone set has a sensitivity to audio-frequencies that peaks at about 1 kHz.

However, harmonics between 1000 and 3000 Hz are hardly diminished in the weighting curves and yet modern electronic telephone sets appear to be more sensitive to these higher frequencies; so perhaps the weightings should be revised accordingly.

Using these weighting factors, the total weighted transverse (or metallic) noise is obtained from the expression:

$$V_m = \sqrt{\sum_{n=1}^{N} (V_{cn} K_n B_n C_n)^2} \tag{4.51}$$

where V_m is the metallic mode weighted voltage, V_{cn} is the longitudinal induced voltage, N is the maximum order of harmonic to be considered, C_n is the psophometric or the C-message weighting factor of harmonic n, K_n is the telephone circuit shielding factor at harmonic n and B_n is the telephone circuit balance at harmonic n. The

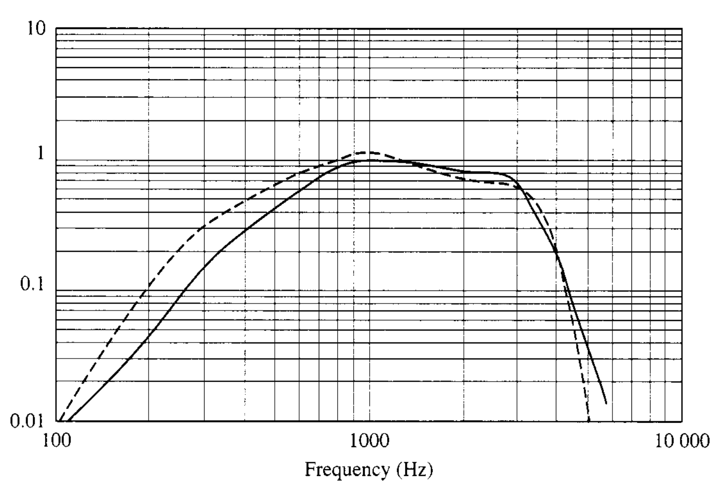

Figure 4.16 C-message (———) and psophometric weighting (– --- –) factors

last two coefficients require detailed information on the telephone systems cables and design as well as knowledge of the local earth resistivities.

Psophometric Weighting

Telephone Form Factor In the psophometric system the level of interference is described in terms of a telephone form factor (TFF), which is a dimensionless value that ignores the geometrical configuration of the coupling and is expressed as

$$\text{TFF} = \sqrt{\sum_{n=1}^{N}\left(\frac{U_n}{U}.F_n\right)^2} \tag{4.52}$$

where U_n is the component at harmonic n of the disturbing voltage, N is the maximum harmonic order to be considered,

$$U = \sqrt{\sum_{n=1}^{N} U_n^2}$$

is the line to neutral total r.m.s. voltage, $F_n = p_n\, nf_0/800$, p_n is the psophometric weighting factor and f_0 is the fundamental frequency (50 Hz). The required limit of TFF is typically 1%.

The CCITT directives recommend that the total psophometric weighted noise on a telephone circuit has an e.m.f. (i.e. open circuit voltage) of less than 1 mV. When measuring the noise voltage the telephone circuit is terminated with its characteristic impedance (which is a resistance of the order of 600 Ω) and the noise voltage is measured across such resistance. Therefore, the psophometrically weighted noise across the terminating resistor must be less than 0.5 mV.

As a general rule, if the TFF is greater than 0.5 mV it is likely to cause interference to telephone services. It must be stressed that the TFF is only a guideline measurement; it is not satisfactory as the sole measure of interference to a communication line as it takes no account of coupling and exposure factors.

Equivalent Disturbing Current The CCITT also defines an equivalent disturbing current (I_P), expressed as

$$I_P = (1/p_{800})\sqrt{\sum_f (h_f p_f I_f)^2} \tag{4.53}$$

where I_f is the component of frequency f of the current causing the disturbance, p_f is the psophometric weighting factor at frequency f, and h_f is a factor which is a function of frequency and takes into account the type of coupling between the lines concerned (by convention $h_{800} = 1$).

C-Message Weighting

Telephone Influence Factor The C-message weighting system uses the telephone influence factor (TIF) instead of the TFF of the psophometric system. Again, TIF is a dimensionless value used to describe the interference of a power transmission line on a telephone line, and is expressed as

$$\mathrm{TIF} = \frac{\sqrt{\sum_{n=1}^{N} (U_n W_n)^2}}{U} \tag{4.54}$$

where U_n is the single frequency r.m.s. voltage at harmonic n, N is the maximum harmonic order to be considered, U is the total line to neutral voltage (r.m.s.), C_n is the C-message weighting factor, f_0 is the fundamental frequency (60 Hz), and W_n is the single frequency TIF weighting at harmonic n (the relationship between the C-message and TIF weighting factors is $W_n = C_n\ 5n\ f_0$ and the corresponding coefficients for the first 25 harmonics are shown in Table 4.8).

The TIF weights account for the fact that mutual coupling between circuits increases linearly with frequency, while the C-message weights do not take this into consideration. Because of this coupling relationship the TIF weights peak at about 2.6 kHz (as compared with the 1 kHz peak of the C-message weighting curve). The TIF index

Table 4.8 C-message and TIF weighting coefficients

Frequency (Hz)	Harmonic	C-message weight, C	TIF W
60 (50)	1	0.0017	0.5 (0.71)
120 (100)	2	0.0167	10.0 (8.91)
180 (150)	3	0.0333	30.0 (35.5)
240 (200)	4	0.0875	105 (89.1)
300 (250)	5	0.1500	225 (178)
360 (300)	6	0.222	400 (295)
420 (350)	7	0.310	650 (376)
480 (400)	8	0.396	950 (484)
540 (450)	9	0.489	1320 (582)
600 (500)	10	0.597	1790 (661)
660 (550)	11	0.685	2260 (733)
720 (600)	12	0.767	2760 (794)
780 (650)	13	0.862	3360 (851)
840 (700)	14	0.912	3830 (902)
900 (750)	15	0.967	4350 (955)
960 (800)	16	0.977	4690 (1000)
1020 (850)	17	1.000	5100 (1035)
1080 (900)	18	1.000	5400 (1072)
1140 (950)	19	0.988	5630 (1109)
1200 (1000)	20	0.977	5860 (1122)
1260 (1050)	21	0.960	6050 (1109)
1320 (1100)	22	0.944	6230 (1072)
1380 (1150)	23	0.923	6370 (1035)
1440 (1200)	24	0.924	6650 (1000)
1500 (1250)	25	0.891	6680 (977)

Note: Number in brackets refer to the CCITT values and 50 Hz fundamental models the effectiveness of induction between adjacent circuits, and is thus particularly useful to assess the interference of power distribution circuits on analogue-type telephone systems (many recent telephone circuits, however, are of digital-type design).

Instead of voltage, the TIF is more usefully expressed in terms of line current because the electromagnetic induction relates to line current amplitude. Also, the line currents are best represented by their sequence of rotation, i.e. positive, negative and zero sequences, respectively. The relationship between the phase and sequence currents is:

$$\begin{bmatrix} I^+ \\ I^- \\ I^0 \end{bmatrix} = 1/\sqrt{3} \begin{bmatrix} 1 & a & a^2 \\ 1 & a^2 & a \\ 1 & 1 & 1 \end{bmatrix} \begin{bmatrix} I_R \\ I_Y \\ I_B \end{bmatrix} \tag{4.55}$$

where $a = 1\angle 120°$.

As the power circuit is three-phase and the audio circuit single phase, the latter can not distinguish between positive- and negative-sequence signals, but the effect of zero sequence is very different.

Telephone circuits are more affected by zero-sequence harmonics because these are in phase in the three phases and add arithmetically. Generally, standards are less tolerant of zero-sequence harmonics to take this fact into account. When only the zero-sequence signals are included in equation (4.54), the term *residual* is used in the TIF (which in the case of a balanced three-phase system include only triplen frequency components). The term balanced is used when the signals included in equation (4.54) are only of positive and negative sequence. The latter contribution to the induced noise is important in the immediate proximity of the transmission line, while the residual signals are the dominant ones at greater distances from the line. When the balanced signals are expected to contribute significantly to the induced noise, they must be included in the calculation of the TIF, i.e.

$$\mathrm{TIF} = \sqrt{\mathrm{TIF}_r^2 + \mathrm{TIF}_b^2} \tag{4.56}$$

where the suffixes r and b indicate residual and balanced, respectively.

Equivalent Disturbing Current

$$I_{\mathrm{eq}} = \sqrt{\sum_{n=1}^{N} (H_n C_n I_n)^2} \tag{4.57}$$

where I_n is the effective disturbing current at harmonic n (generally corresponding to residual mode currents), N is the maximum harmonic order to be considered, C_n is the C-message weighting factor, and H_n is the weighting factor normalised to reference frequency (1000 Hz) that accounts for the frequency dependence of mutual coupling, shielding and communication circuit balance at harmonic n.

Again, when the balanced mode harmonic currents are taken into account the effective disturbing current is then specified as

$$I_n = \sqrt{(I_{rn})^2 + (K_b I_{bn})^2} \tag{4.58}$$

where I_{rn} is the total residual mode current at harmonic n, I_{bn} is the balanced mode current at harmonic n, and K_b is the ratio of balanced mode coupling to the residual mode coupling at reference frequency.

IT and kVT Products The THD, TFF and TIF indices do not provide information about the amplitude of the voltage (or current) to which they relate. The IT and VT (or kVT) products incorporate that information. In these products the voltages or currents of the power transmission line are represented by a single voltage or current obtained by weighting each harmonic voltage or current with the corresponding factor of the system (BTS-EEI or CCITT) used.

The VT product incorporates the line-to-line voltage amplitude, i.e.

$$\mathrm{VT} = \sqrt{\sum_{n=1}^{N} (W_n V_n)^2} \tag{4.59}$$

where V_n is the single frequency r.m.s. line-to-line voltage at harmonic n, N is the maximum harmonic order to be considered and $W_n = C_n\ 5\ nf_0$ is the single frequency TIF weighting at harmonic n.

The IT product is derived similarly using currents instead of voltages. However, there is some confusion about whether to use the current in one phase or some kind of sum of the three phase currents; if the latter is used it would have to be a phasor sum, which is usually close to zero. Our interpretation is that when the analogue telephone circuit is far from the three-phase power line, the value to be used is the phasor combination of the three line currents. On the other hand, if the telephone circuit is in the vicinity of the power line then the calculation based on one phase current or a non-phasor combination of the three phases will be a better indicator of telephone interference. In [39] the line current is multiplied by a factor of $\sqrt{3}$ to account for the three-phase arrangement. Ideally, when assessing the interfering ability of a power line the effect of the balanced sequence components needs to be represented, including the geometry of the line, the separation of the telephone cable and the soil conditions; these effects are significant in typical road width type separations.

Taking into account the above ambiguities and the fact that analogue communication circuits are gradually being replaced by digital ones, the TIF and IT indices are not used extensively by telecommunication companies, as they do not reflect the vulnerabilities of services operating over telephone cables. The weightings currently used are based on the response of the human ear in conjunction with the telephone receiver. However, modern telecommunication devices tend to use all of the available spectrum that the cable can propagate. Even devices such as modems, which are restricted to the traditional telephony band from 300 Hz to 3400 Hz, utilise the high-frequency part of this band (1000 to 3400 Hz) much more extensively than the human voice does, and are consequently much more sensitive to interference at these frequencies. Thus it is possible to have acceptable TIF or psophometric noise on a telephone cable but still have severe service degradation.

More modern telecommunication services such as ADSL utilise the spectrum from about 25 kHz up to 1 MHz and beyond. As they are not using the low frequencies they are proving to be tolerant of typical power line noise, but they are likely to be sensitive to electrical fast transients.

Illustrative Example A 4.16 kV 60 Hz distribution system bus is selected to supply a three-phase 4.5 MVA purely resistive load. The corresponding fundamental frequency current in the line is

$$I_1 = (P_1/3)/(V_1/\sqrt{3}) = (4500/3)/(4.16/\sqrt{3}) = 624.5376 \text{ A}$$

and the load resistance is

$$R = (V_1/\sqrt{3})/I_1 = (4160/\sqrt{3})/(624.537) = 3.8457 \ \Omega$$

Two different voltage waveforms, both with the same THD, are used to illustrate the relative effect of different frequencies on the IT and TIF indices. The voltage waveforms include either

(1) 75 V of zero sequence third harmonic and 177 V of negative sequence fifth, or
(2) 177 V of zero sequence third harmonic and 75 V of negative sequence fifth.

In both cases the total harmonic distortion (THD) is the same, i.e.

$$\text{THD} = 100\left(\sqrt{V_3^2 + V_5^2}/V_1\right) = 100\left(\sqrt{75^2 + 177^2}/4160\right) = 4.62\%$$

which is below the 5% limit recommended by IEEE standard 519 for the 4.16 kV distribution system.

The ANSI 368 standard indicates that telephone interference from a 4.16 kV distribution system is unlikely to occur when the IT index is below 10 000. Let us check the two cases against this limit using the TIF weightings (taken from Table 4.8) for the fundamental (0.5), the third (30) and fifth (225) harmonic frequencies:

Case (i):

$$I_3 = \frac{(V_3/\sqrt{3})}{R} = \frac{(75/\sqrt{3})}{3.8457} = 11.26 \text{ A}$$

$$I_5 = \frac{(V_5/\sqrt{3})}{R} = \frac{(177/\sqrt{3})}{3.8457} = 26.57 \text{ A}$$

Therefore the following IT indices result (using the questionable $\sqrt{3}$ factor as suggested in [39]):

$$\text{IT}_1 = (624.537)(0.5)\sqrt{3} = 540.865$$

$$\text{IT}_3 = (11.26)(30)\sqrt{3} = 585.071$$

$$\text{IT}_5 = (26.57)(225)\sqrt{3} = 10\,355.752$$

and the total IT including the balanced and residual components becomes:

$$\text{IT (total)} = \sqrt{(540.865)^2 + (585.087)^2 + (10\,354.63)^2} = 10\,386.358$$

which is above the ANSI standard limit.

The TIF index for this case is

$$\mathrm{TIF} = \frac{\sqrt{(0.5x4160)^2 + (225x75)^2 + (30x177)^2}}{\sqrt{(4160)^2 + (75)^2 + (177)^2}} = 9.59$$

Case (ii):

$$\mathrm{I}_3 = \frac{(177/\sqrt{3})}{3.8457} = 26.57\ \mathrm{A}$$

$$\mathrm{I}_5 = \frac{(75/\sqrt{3})}{3.8457} = 11.26\ \mathrm{A}$$

and the corresponding ITs

$$\mathrm{IT}_3 = (26.57)(30)\sqrt{3} = 1380.77$$

$$\mathrm{IT}_5 = (11.26)(225)\sqrt{3} = 4388.103$$

$$\mathrm{IT(total)} = \sqrt{(540.865)^2 + (1380.62)^2 + (4388.15)^2} = 4631.83$$

which is well below the 10 000 limit.

The total TIF index for this case is

$$\mathrm{TIF} = \frac{\sqrt{(0.5x4160)^2 + (30x177)^2 + (225x75)^2}}{\sqrt{(4160)^2 + (177)^2 + (75)^2}} = 4.28$$

Typical requirements of TIF are between 15 and 50.

4.9.5 Telephone Circuit Balance to Earth

If the telephone line or terminal equivalent is not perfectly balanced with respect to earth, longitudinally induced voltage on that line can be transformed into transversed voltage, and it is as a transversed voltage across the ear piece that we can hear noise on the telephone.

Consider the simple telephone circuit of Figure 4.12. If the impedances to earth of the two wires that form a telephone circuit are different, and assuming that the two wires are subjected to the same longitudinal induction, different currents will flow in each wire. Because the self-impedances of the wires will be very similar, the different current flows will give rise to a transverse voltage between the pair.

Factors which may affect the telephone circuit balance to earth are (i) any leakage paths to earth, e.g. across-the-pole insulators or through-cable insulation; and (ii) any unbalanced terminal equipment, either subscriber's equipment (e.g. extension bells) or exchange equipment. Some of these factors may be corrected simply and quickly to reduce an induced noise problem, while others rely on the proper design and manufacture of the equipment or plant.

Balance is simply defined as

$$20 \log_{10} \left(\frac{\text{Longitudinal voltage}}{\text{Transverse voltage}} \right)$$

and for the majority of relay sets it is of the order of 45–50 dB.

The most critical component balance-wise is generally any terminal relay set in a communication connection. As there are large numbers of such relay sets, it is not practical to set balance objectives at a very high level as this would significantly increase the cost of the communication network.

4.9.6 Shielding

A metallic or conducting earthed screen such as a cable sheath which encloses the telephone circuit over the length of an exposure is totally effective in eliminating electrostatic induction. Buried cables with a metallic screen or sheath are also immune to electrostatic induction due to the conducting effect of the earth.

Metallic screens or sheaths are only partly effective in reducing the effects of electromagnetic induction.

The mechanism of electromagnetic shielding is as follows. The power line current causes longitudinal voltages to be induced in the wires of the cable and also in the shield. The resulting current flow in the shield is in the opposite direction to the inducing current in the power line. This induced current in the shield generates in turn a voltage in the wires of the cable opposing the voltage induced by the power line current, thus tending to neutralise the latter.

If the shield current is large and/or well coupled to the inner conductors, reasonable shielding factors can be obtained.

The shielding factor of a cable is the fraction by which the shield reduces the voltage induced into the core; the shielding factor (K) of an installed shielded cable system is given by

$$K = \frac{\text{d.c. shield resistance} + \text{earth resistance}}{\text{a.c. shield resistance} + \text{earth resistance}} \tag{4.60}$$

This simple formula shows that the shielding factor may be reduced by decreasing the resistance and/or increasing the inductance of the sheath and/or decreasing the earth resistance.

A good shielding factor for a cable can be achieved if the following set of conditions hold:

(1) At points where the shield is earthed, the earth resistance must be low (typically 1–2 Ω).
(2) There is a low resistance shield, i.e. plenty of metal in the cable shield to keep the resistance low.
(3) Steel tapes are usually required.

All these factors mean that shielded cables used for noise shielding are relatively expensive; typically they can be expected to cost some 30–50% more than standard unshielded types.

4.9.7 Mitigation Techniques

The steps to be taken when a noise problem is known to exist are:

(1) Check the transverse noise voltage in the telephone circuit. If the e.m.f. is less than 1 mV no further steps need be taken; if it is greater than 1 mV, then proceed as below.
(2) Determine whether the mode of induction is of electrostatic or electromagnetic type.
(3) Test telephone line and termination balance to earth.
(4) Where considered useful, derive by test the telephone form factor in various parts of the power system to try to isolate the source of noise.

With the information gained from these tests, the following mitigation methods are considered:

(1) A reduction of the influence of the power system, which can be achieved by (i) physical relocation of either system (usually an expensive exercise); (ii) replacing the copper wire with light conducting fibres (fibre optics); or (iii) reduction of the harmonic content in the power system (the appropriate techniques are discussed in Chapter 6).
(2) Reducing coupling is not usually a practical proposition, except in cases of earth potential rise, where a noisy incoming multiple earthed neutral can be excluded if required.
(3) Reducing the susceptiveness of the communication circuits can be achieved by the use of noise chokes, noise-neutralising transformers, shielded cable and derived circuits.

Noise Chokes Reducing factors upwards of 25 dB are achievable in certain circumstances. Generally, noise chokes are only useful for improving the balance of substandard terminal relay sets. They work by increasing the a.c. longitudinal line impedance, thereby reducing the noise current and with it the transverse noise voltage level.

Noise-Neutralising Transformers These work by inducing an equal but opposite (in-phase) noise voltage in affected cable pairs, thus reducing the induced longitudinal noise. Reduction factors of 15–20 dB can be achieved.

Shielded Cable Reduction factors upward of 60 dB can easily be achieved, but it is costly. Generally the use of shielded cable is only applicable for new work.

Derived Circuits By providing the circuits via pulse code modulation (PCM) or frequency division multiplexing (FDM), the system can be made relatively immune to noise induction. In each case the degree of improvement depends on the circumstances.

It must be emphasised that the actual reduction factors depend on the particular circumstances and the factors mentioned above are not achievable in all cases.

4.10 Audible Noise from Electric Motors

The major causes of sound from electric motors are

- torque pulsations in induction and permanent magnet machines;
- torque and normal force pulsations in the doubly salient structure of the switched reluctance machines.

Motors powered by pulse width modulation exhibit their predominant sound levels at the modulating frequency. The level of sound is not a function of the load but is inversely proportional to the motor speed.

The highest levels are produced by switched reluctance motors and are related to the load torque.

4.11 Discussion

In the absence of an extremely intelligent harmonic traffic controller, the presence of waveform distortion is normally detected by its effects on power system components or on personnel or plant outside the power system. Some effects, such as telephone interference, are immediately obvious to the senses and can thus be mitigated at an early stage of the problem without excessive disturbance. Other effects, such as a resonant condition, can occur at unexpected locations lacking monitoring facilities and will often cause expensive damage to plant components such as capacitor banks and inadequate filter equipment. The financial consequences of harmonic overloading and equipment failure, not discussed in this chapter, should be an essential part in the design of modern power systems.

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