NEAR EAST UNIVERSITY

Faculty of Engineering

Department of Electrical and Electronic Engineering

EFFECTS OF HARMONICS ON LARGE INSTALLATION

Graduation Project
EE- 400

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CONTENTS

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ABSTRACT

Providing consumer (public) with a high quality, cheap and continuous power supply are the objectives electrical energy producing and distributing firms.High quality means achieving constant voltage with sinusoidal waveform through a constant network frequency. However in such energy applications some difficulties may be encountered. Distortion of the sinusoidal waveform may take place as a resulst different elements (devices) connected to the energy distribution system and difficiencies encountered by those elements (like generation harmonic).

In general, distortion of perfect sinusoidal voltage refers to he occurrence components called harmonics. The main factors causing these components called harmonics. The main factors causing these components are the nonlinear characteristic y magnetic and electrical circuits.

Adversely factors affecting the power system as a result of harmonics must be prevented and eliminated when problems occur. One of the most imported methods to eliminate such problem is to manufacture devices don't produces harmonics another method however is to use harmonic filters.

In this study, Ercan Airport power distribution system is considered as a case study. The data collected by the harmonic analyser prepared using analyser were prepared using excel file and graphs were plotted

INTRODUCTION

Basically harmonics are the results of either non linear or non sinusoidal sources or both. The occurrence of harmonic current or voltage in power system refers to the distortion in sinusoidal signal is referred to as non sinusoidal signal. Using Fourier Analyses this signal are expressed interms of fundamental frequency and other frequency components. This analysis allow to expressed the non sinusoidal signal in mathematical expression as the sum of sinusoidal signals with different frequencies hands, providing a simple way to analysis the harmonics. Harmonics, cause technical and economics problem. Additional lose additional voltage drops, resonance, power factor variation,etc.

CHAPTER ONE

1.ANALYSIS OF NON-SINUSOIDAL WAVEFORMS

1.1. Waveforms

Up to the present, we have been consider sinusoidal alternating waveforms as Show in figure 1.1.1 and 1.1.2 respectively.

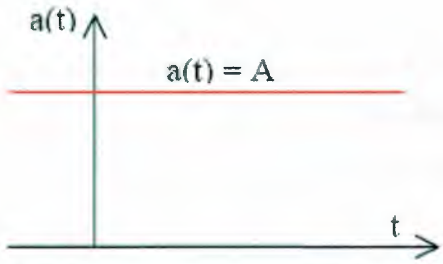


Figure 1.1.1 – Direct Waveform

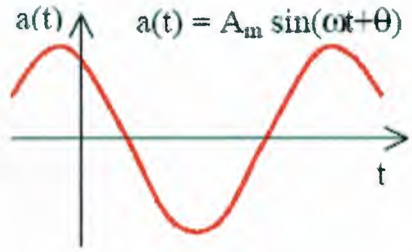


Figure 1.1.2 – Sinusoidal Waveform

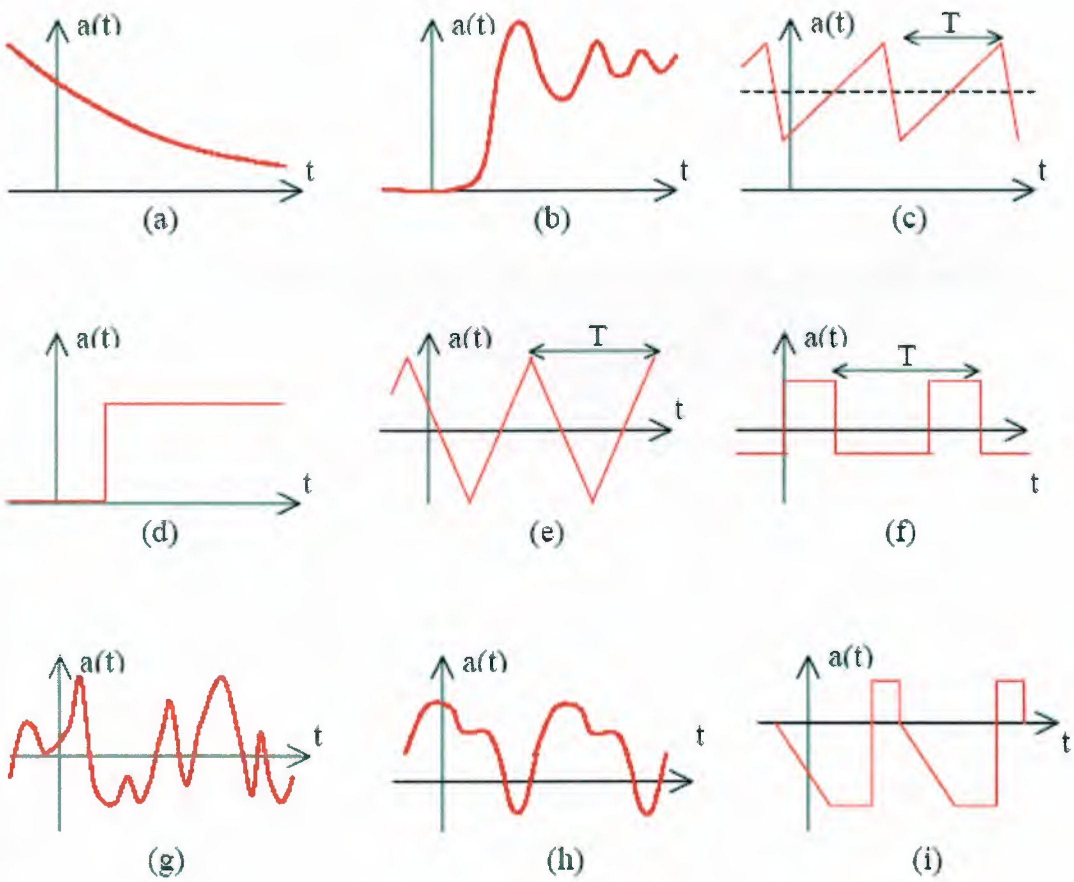


Figure 1.1.3 Waveforms

It can be seen that the waveforms of Figure 1.1.3 (a), (b), (c) and (d) are uni-directional, although not purely direct. Waveforms of Figure 1.1.3 (e) and (f) are repetitive waveforms with zero mean value, while figure1.1.3 (c), (h) and (i) are repetitive waveforms with finite mean values. Figure 1.1.3 (g) is alternating but non-repetitive and mean value is also non-zero.Thus we see that there are basically two groups of waveforms, those that are repetitive and those which are non-repetitive. These will be analysed separately in the coming sections. In a repetitive waveform, only one period "T" needs to be defined and can be broken up to a fundamental component (corresponding to the period T) and its harmonics. A uni-directional term (direct component) may also be present. This series of terms is known as Fourier Series named after the French mathematician who first presented the series in 1822.

1.2. Fourier Series

The Fourier series states that any practical periodic function (period T or frequency $\omega_0 = 2\pi/T$) can be represented as an infinite sum of sinusoidal waveforms (or sinusoids) that have frequencies which are an integral multiple of ω_o.

$$f(t) = F_o + F_1 \cos(\omega_0 t + \theta_1) + F_2 \cos(2\omega_0 t + \theta_2) + F_3 \cos(3\omega_0 t + \theta_3) + F_4 \cos(4\omega_0 t + \theta_4) + F_5 \cos(5\omega_0 t + \theta_5) + \dots\dots\dots$$

Usually the series is expressed as a direct term ($A_o/2$) and a series of cosine terms and sine terms.

$$f(t) = A_o/2 + A_1 \cos \omega_0 t + A_2 \cos 2\omega_0 t + A_3 \cos 3\omega_0 t + A_4 \cos 4\omega_0 t + \dots + B_1 \sin \omega_0 t + B_2 \sin 2\omega_0 t + B_3 \sin 3\omega_0 t + B_4 \sin 4\omega_0 t + \dots..$$

$$f(t) = \frac{A_o}{2} + \sum_{n=1}^{\infty} (A_n \cos n\omega_o t + B_n \sin n\omega_o t)$$

This, along with the Superposition theorem, allows us to find the behaviour of circuits to arbitrary periodic inputs. Before going on to the analysis of the Fourier series, let us consider some of the general properties of waveforms which will come in useful in the analysis.

1.3. Symmetry in Waveforms

Many periodic waveforms exhibit symmetry. The following three types of symmetry help to reduce tedious calculations in the analysis.

(1) Even symmetry

(2) Odd symmetry

(3) Half-wave symmetry

1.3.1. Even Symmetry

A function f(t) exhibits even symmetry, when the region before the y-axis is the mirror image of the region after the y-axis.

i.e. $f(t) = f(-t)$

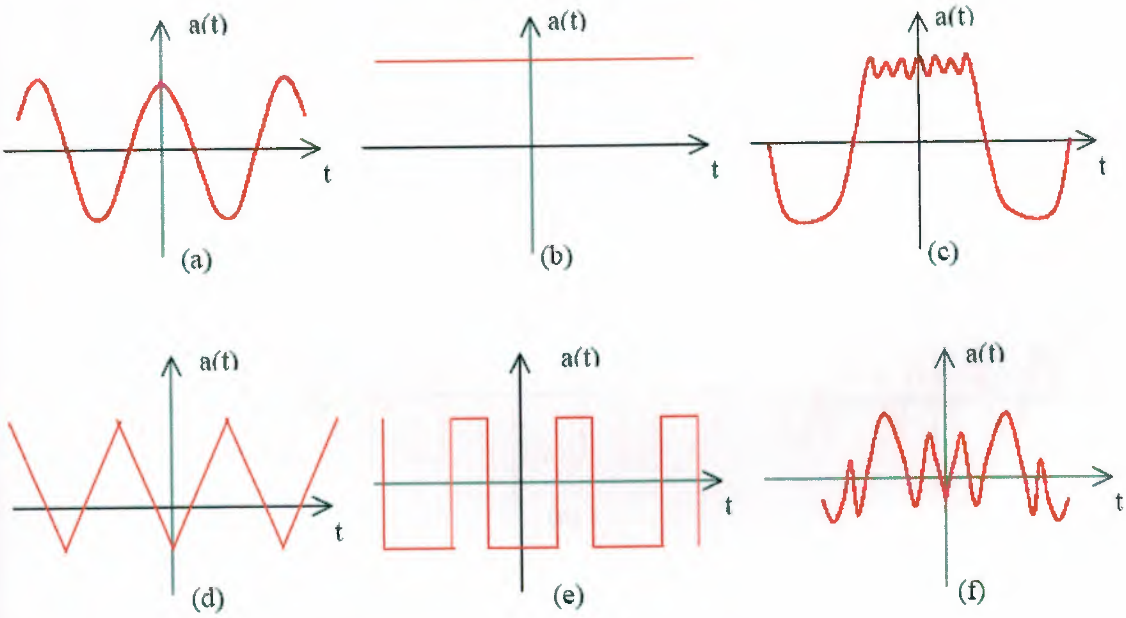


Figure 1.3.1 Waveforms with Even symmetry

The two simplest forms of the Even Function or waveform with even symmetry are the cosine waveform and the direct waveform as shown in figure 1.3.1 (a) and (b).It can also be seen from the waveforms seen in the figure 1.3.1 that even symmetry can exist in both periodic and non-periodic waveforms, and that both direct terms as well as varying terms can exist in such waveforms.It is also evident, that if the waveform is defined for only t ≥ 0, the remaining part of the waveform is automatically known by symmetry.

1.3.2. Odd Symmetry

A function f(t) exhibits even symmetry, when the region before the y-axis is the negative of the mirror image of the region after the y-axis.

$$\text{i.e. } f(t) = (\ -)\ f(-t)$$

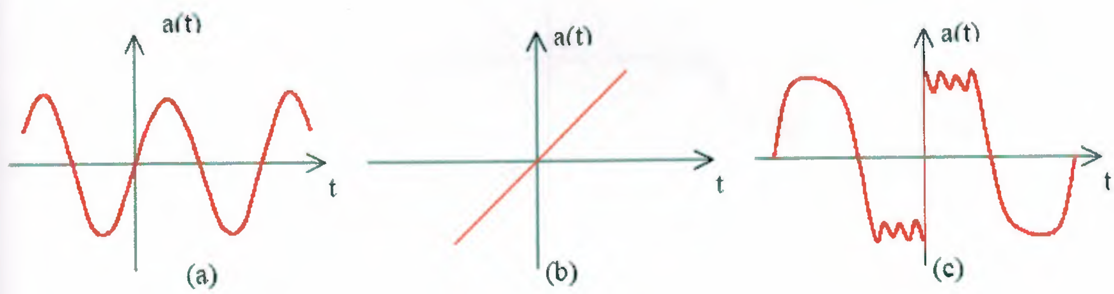


Figure 1.3.2

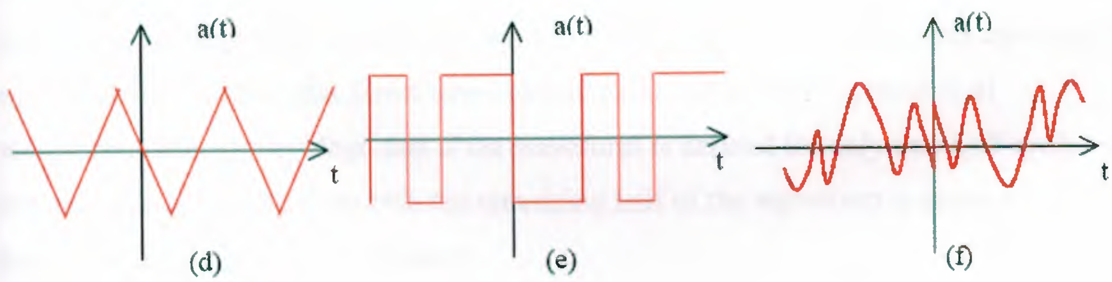


Figure 1.3.3 Waveforms with Odd symmetry

The two simplest forms of the Odd Function or waveform with odd symmetry are the sine waveform and the ramp waveform as shown in figure 1.3.3 (a) and (b). It can also be seen from the waveforms seen in the figure 1.3.3 that odd symmetry can exist in both periodic and non-periodic waveforms,and that only varying terms can exist in such waveforms. Note that direct terms cannot exist in odd waveforms.It is also evident, that if the waveform is defined for only t ≥ 0, the remaining part of the waveform is automatically known by the properties of symmetry.

1.3.3. Half-wave Symmetry

A function f(t) exhibits half-wave symmetry, when one half of the waveform is exactly equal to the negative of the previous or the next half of the waveform.

$$\text{i.e. } f(t) = (-)f(t-\tfrac{T}{2}) = (-)f(t+\tfrac{T}{2})$$

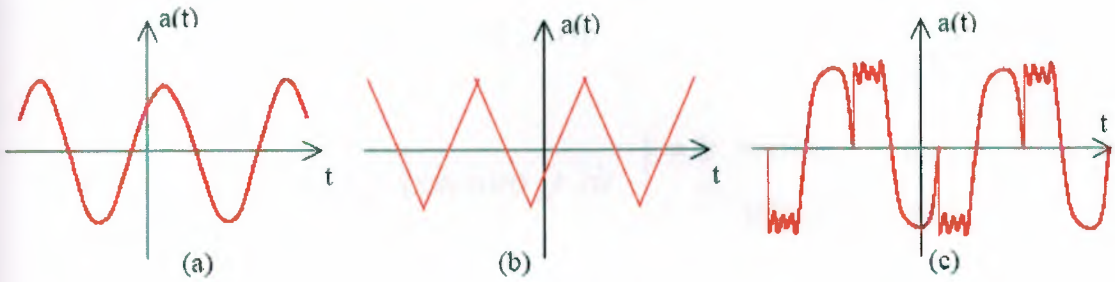


Figure 1.3.4 Waveforms with Half-wave symmetry

The simplest form of Half-wave Symmetry is the sinusoidal waveform as shown in figure 1.3.4(a). It can also be seen from the waveforms in the figure 1.3.4 that half-wave symmetry can only exist in periodic waveforms, and that only varying terms can exist in such waveforms. Note that direct terms cannot exist in half-wave symmetrical waveforms. It is also evident, that if the waveform is defined for only **one half cycle**, not necessarily starting from t=0, the remaining half of the waveform is automatically known by the properties of symmetry.

1.4. Some useful Trigonometric Properties

The sinusoidal waveform being symmetrical does not have a mean value, and thus when integrated over a complete cycle or integral number of cycles will have zero value. From this the following properties follow. [Note: $\omega_o T = 2\pi$)]

$$\int_{t_o}^{t_o+T} \sin \omega_o t.dt = 0 \qquad \int_{t_o}^{t_o+T} \cos \omega_o t.dt = 0$$

$$\int_{t_o}^{t_o+T} \sin n\omega_o t.dt = 0 \qquad \int_{t_o}^{t_o+T} \cos n\omega_o t.dt = 0$$

$$\int_{t_o}^{t_o+T} \sin n\omega_o t.\cos m\omega_o t.dt = 0 \quad \text{for all values of } \mathbf{m} \text{ and } \mathbf{n}$$

$$\int_{t_o}^{t_o+T} \sin n\omega_o t.\sin m\omega_o t.dt \begin{cases} = 0 & \text{when } n \neq m \\ = \frac{T}{2} & \text{when } n = m \end{cases}$$

$$\int_{t_o}^{t_o+T} \cos n\omega_o t.\cos m\omega_o t.dt \begin{cases} = 0 & \text{when } n \neq m \\ = \frac{T}{2} & \text{when } n = m \end{cases}$$

Evaluation of Coefficients A_n and B_n

$$f(t) = \frac{A_o}{2} + \sum_{n=1}^{\infty} (A_n \cos n\omega_o t + B_n \sin n\omega_o t)$$

You will notice that the first term of the Fourier Series is written as $A_o/2$ rather than A_o. This is because it can be shown that A_o can also be evaluated using the same general expression as for A_n with n=0. It is also worth noting that $A_o/2$ also corresponds to the direct component of the waveform and may be obtained directly as the mean value of the waveform.

Let us now consider the general method of evaluation of coefficients.

Consider the integration of both sides of the Fourier series as follows.

$$\int_{t_0}^{t_o+T} f(t)\cdot dt = \int_{t_0}^{t_0+T} \frac{A_o}{2}\cdot dt + \int_{t_0}^{t_0+T} \sum_{n=1}^{\infty} (A_n \cos n\omega_o t + B_n \sin n\omega_o t)\cdot dt$$

using the properties of trigonometric functions derived earlier, it is evident that only the first term on the right hand side of the equation can give a non zero integral.

i.e. $$\int_{t_0}^{t_0+T} f(t)\cdot dt = \int_{t_0}^{t_0+T} \frac{A_o}{2}\cdot dt + 0 = \frac{A_o}{2}\cdot T$$

$$\therefore A_o = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot dt$$

or from mean value we have which gives the same result.

$$\frac{A_o}{2} = \frac{1}{T}\int_{t_o}^{t_o+T} f(t)\cdot dt$$

Consider the integration of both sides of the Fourier series, after multiplying each term by cos $n\omega_o t$ as follows.

$$\int_{t_0}^{t_o+T} f(t)\cdot \cos n\omega_o t\cdot dt = \int_{t_0}^{t_o+T} \frac{A_o}{2}\cdot \cos n\omega_o t\cdot dt + \int_{t_0}^{t_o+T} \sum_{n=1}^{\infty}(A_n \cos n\omega_o t + B_n \sin n\omega_o t)\cdot \cos n\omega_o t\cdot dt$$

using the properties of trigonometric functions derived earlier, it is evident that only cos $n\omega_o t$ term on the right hand side of the equation can give a non zero integral.

$$\therefore A_n = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot \cos n\omega_o t\cdot dt$$

Similarly integration of both sides of the Fourier series, after multiplying by $\sin n\omega_o t$ gives

$$\therefore B_n = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot \sin n\omega_o t\cdot dt$$

1.5. Analysis of Symmetrical Waveforms

1.5.1. Even Symmetry

When even symmetry is present, the waveform from 0 to T/2 also corresponds to the mirror image of the waveform from –T/2 to 0.Therefore it is useful to select to = –T/2 and integrate from t = –T/2.

f(t) = f(-t)

$$A_n = \frac{2}{T}\int_0^T f(t)\cdot \cos n\omega_o t \cdot dt$$

$$A_n = \frac{2}{T}\int_{-T/2}^{0} f(t)\cdot \cos n\omega_o t \cdot dt + \frac{2}{T}\int_0^{T/2} f(t)\cdot \cos n\omega_o t \cdot dt$$

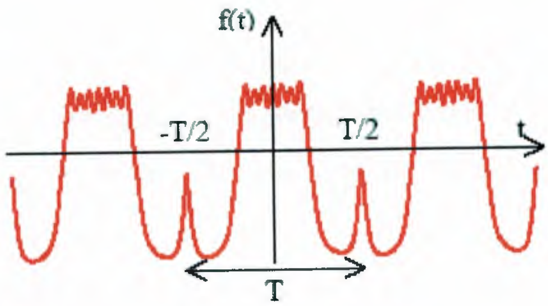


Figure 1.5.1 Analysis of even waveform

If in the first part of the expression, if the variable 't' is replaced by the variable '-t' the equation may be re-written as

$$A_n = \frac{2}{T}\int_{T/2}^{0} f(-t)\cdot \cos(-n\omega_o t)\cdot(-dt) + \frac{2}{T}\int_0^{T/2} f(t)\cdot \cos n\omega_o t \cdot dt$$

Since the function is even, f(-t) = f(t), and cos(-nωot) = cos(nωot).

Thus the equation may be simplified to

$$A_n = (-)\frac{2}{T}\int_{T/2}^{0} f(t)\cdot \cos(n\omega_o t)\cdot dt + \frac{2}{T}\int_0^{T/2} f(t)\cdot \cos n\omega_o t \cdot dt$$

The negative sign in front of the first integral can be replaced by interchanging the upper and lower limits of the integral. In this case it is seen that the first integral term and the second integral term are identical. Thus

$$A_n = \frac{2\times 2}{T}\int_0^{T/2} f(t)\cdot \cos n\omega_o t\cdot dt$$

Thus in the case of even symmetry, the value of An can be calculated as twice the integral over half the cycle from zero.A similar analysis can be done to calculate Bn. In this case we would have

$$B_n = \frac{2}{T}\int_{T/2}^{0} f(-t)\cdot \sin(-n\omega_o t)\cdot(-dt) + \frac{2}{T}\int_0^{T/2} f(t)\cdot \sin n\omega_o t\cdot dt$$

Since the function is even, $f(-t) = f(t)$, and $\sin(-n\omega_o t) = -\sin(n\omega_o t)$.

In this case the two terms are equal in magnitude but have opposite signs so that they cancel out.Therefore $B_n = 0$ for all values of **n** when the waveform has **even symmetry**.

Thus an even waveform will have only cosine terms and a direct term.

$$f(t) = \frac{A_o}{2} + \sum_{n=1}^{\infty} A_n \cos n\omega_o t$$

$$A_n = \frac{4}{T}\int_0^{T/2} f(t)\cdot \cos n\omega_o t\cdot dt \quad - \frac{2}{T}\int_0^{T/2} f(t)\cdot \cos n\omega_o t\cdot dt$$

1.5.2 Odd Symmetry

When odd symmetry is present, the waveform from 0 to T/2 also corresponds to the negated mirror image of the waveform from –T/2 to 0.

Therefore as for even symmetry $t_o = -T/2$ is selected and integrated from $t = -T/2$.

$f(t) = -f(-t)$

$$A_n = \frac{2}{T}\int_0^T f(t)\cdot\cos n\omega_o t\cdot dt$$

$$A_n = \frac{2}{T}\int_{-T/2}^{0} f(t)\cdot\cos n\omega_o t\cdot dt + \frac{2}{T}\int_0^{T/2} f(t)\cdot\cos n\omega_o t\cdot dt$$

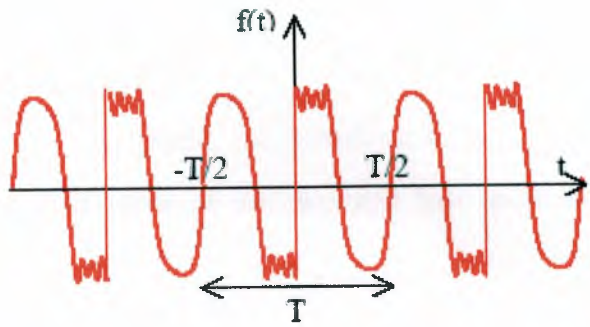


Figure 1.5.2 – Analysis of odd waveform

In the first part of the expression, if the variable 't' is replaced by the variable '-t' it can be easily seen that this part of the expression is exactly equal to the negative of the second part.

$$\therefore A_n = (-)\frac{2}{T}\int_0^{T/2} f(t)\cdot\cos n\omega_o t\cdot dt + \frac{2}{T}\int_0^{T/2} f(t)\cdot\cos n\omega_o t\cdot dt = 0$$

for all **n** for odd waveform In a similar way, for Bn, the two terms can be seen to exactly add up.

Thus

$$B_n = \frac{2\times 2}{T}\int_0^{T/2} f(t)\cdot\sin n\omega_o t\cdot dt$$

Thus in the case of odd symmetry, the value of B_n can be calculated as twice the integral over half the cycle from zero.

Thus an **odd waveform** will have only **sine** terms and no direct term.

$$f(t)=\sum_{n=1}^{\infty}B_n \sin n\omega_o t$$

$$B_n=\frac{4}{T}\int_0^{T/2} f(t)\cdot \sin n\omega_o t\cdot dt$$

1.5.3. Half-wave Symmetry

When half-wave symmetry is present, the waveform from $(t_0+T/2)$ to (t_0+T) also corresponds to the negated value of the previous half cycle waveform from **to** to $(t_o+T/2)$.

$$A_n=\frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot \cos n\omega_o t\cdot dt$$

$$A_n=\frac{2}{T}\int_{t_o}^{t_o+T/2} f(t)\cdot \cos n\omega_o t\cdot dt+\frac{2}{T}\int_{t_o+T/2}^{t_o+T} f(t)\cdot \cos n\omega_o t\cdot dt$$

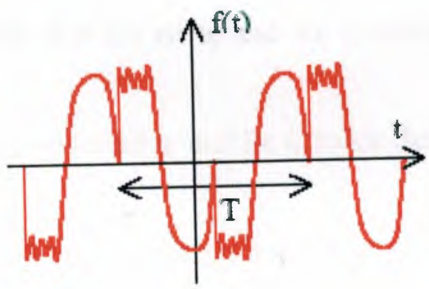


Figure 1.5.3– Analysis of waveform with half wave symmetry

In the second part of the expression, the variable 't' is replaced by the variable 't-T/2'.

$$A_n=\frac{2}{T}\int_{t_o}^{t_o+T/2} f(t)\cdot \cos n\omega_o t\cdot dt+\frac{2}{T}\int_{t_o}^{t_o+T/2} f(t-T/2)\cdot \cos n\omega_o (t-T/2)\cdot d(t-T/2)$$

f(t-T/2) = - f(t) for half-wave symmetry, and since $\omega_o T = 2\pi$, cos $n\omega_o(t - T/2)$ = cos $(n\omega_o t - n\pi)$ which has a value of (-)cos $n\omega_o t$ when n is odd and has a value of cos $n\omega_o t$ when n is even.From the above it follows that the second term is equal to the first term when n is odd and the negative of the first term when n is even.Thus when n is odd and

$A_n = 0$ when n is even

$$A_n = \frac{2\times 2}{T}\int_{t_o}^{t_o+T/2} f(t)\cdot\cos n\omega_o t\cdot dt$$

Similarly it can be shown that

$$B_n = \frac{2\times 2}{T}\int_{t_o}^{t_o+T/2} f(t)\cdot\sin n\omega_o t\cdot dt$$

when n is odd and $B_n = 0$ when n is even Thus it is seen that in the case of **half-wave symmetry**, even harmonics do not exist and that for the odd harmonics the coefficients An and Bn can be obtained by taking double the integral over any half cycle.It is to be noted that many practical waveforms have half-wave symmetry due to naturalcauses.

Summary of Analysis of waveforms with symmetrical properties

1. *With* ***even symmetry****, Bn is 0 for all n, and An is twice the integral over half the* cycle from zero time.

2. With **odd symmetry**, An is 0 for all n, and Bn is twice the integral over half the cycle *from zero time.*

3. With **half-wave symmetry**, An and Bn are 0 for even n, and twice the integral over any half cycle for odd n.

4. If **half-wave symmetry** and either **even symmetry** or **odd symmetry** are present,then An and Bn are 0 for even n, and four times the integral over the quarter cycle for odd n for **An** or **Bn** respectively and zero for the remaining coefficient.

5. It is also to be noted that in any waveform, **Ao/2** corresponds to the mean value of the waveform and that sometimes a symmetrical property may be obtained by subtracting this value from the waveform.

1.6. Piecewise Continuous waveforms

Most waveforms occurring in practice are continuous and single valued (i.e. having a single value at any particular instant). However when sudden changes occur (such as in switching operations) or in square waveforms, theoretically vertical lines could occur in the waveform giving multi-values at these instants. As long as these multi-values occur over finite bounds,the waveform is single-valued and continuous in pieces, or said to be Piecewise continuous.

Figure 1.6.1 – Piecewise continuous waveform

Figure 9 shows such a waveform. Analysis can be carried out using the Fourier Series for both continuous or piecewise continuous waveforms. However in the case of piecewise continuous waveforms, the value calculated from the Fourier Series for the waveform at the discontinuities would correspond to the mean value of the vertical region. However this is not a practical problem as practical waveforms will not have exactly vertical changes but those occurring over very small intervals of time.

1.7.Frequency Spectrum

The frequency spectrum is the plot showing each of the harmonic amplitudes against frequency. In the case of periodic waveforms, these occur at distinct points corresponding to practical waveforms, the higher harmonics have significantly lower amplitudes compared to the lower harmonics. For smooth waveforms, the higher harmonics will be negligible, but for waveforms with finite discontinuities (such as square waveform) the harmonics do not decrease very rapidly. The harmonic magnitudes are taken as 2 2n n B A + for the nth harmonic and has thus a positive value. Each component also has a phase angle which can be determined.

1.8. Effective Value of a Periodic Waveform

The effective value of a periodic waveform is also defined in terms of power dissipation and is hence the same as the r.m.s. value of the waveform.

$$A_{effective} = \sqrt{\frac{1}{T}\int_{to}^{to+T} a^2(t)\cdot dt}$$

Since the periodic waveform may be defined as

$$a(t) = \frac{A_o}{2} + \sum_{n=1}^{\infty} A_n \cos n\omega_o t + \sum_{n=1}^{\infty} B_n \sin n\omega_o t$$

$$A_{eff} = \sqrt{\frac{1}{T}\int_{to}^{to+T}\left[\frac{A_o}{2} + \sum_{n=1}^{\infty} A_n \cos n\omega_o t + \sum_{n=1}^{\infty} B_n \sin n\omega_o t\right]^2 dt}$$

$$A_{eff} = \sqrt{\frac{1}{T}\int_{to}^{to+T}\left[\left(\frac{A_o}{2}\right)^2 + \sum_{n=1}^{\infty}(A_n \cos n\omega_o t)^2 + \sum_{n=1}^{\infty}(B_n \sin n\omega_o t)^2 + \sum product\ terms\right]dt}$$

Using the trigonometric properties derived earlier, only the square terms will give non-zero integrals. The Product terms will all give zero integrals.

$$A_{eff} = \sqrt{\frac{1}{T}\left[\left(\frac{A_o}{2}\right)^2\cdot T + \sum_{n=1}^{\infty} A_n^{\ 2}\cdot\frac{T}{2} + \sum_{n=1}^{\infty} B_n^{\ 2}\cdot\frac{T}{2}\right]} = \sqrt{\left(\frac{A_o}{2}\right)^2 + \sum_{n=1}^{\infty}\frac{A_n^{\ 2}}{2} + \sum_{n=1}^{\infty}\frac{B_n^{\ 2}}{2}}$$

$\frac{A_o}{2}$ is the d.c. term, and $\frac{\sqrt{(A_n^2 + B_n^2)}}{\sqrt{2}}$ is the r.m.s. value of the nth harmonic.

Thus the effective value or r.m.s. value of a periodic waveform is the square root of the sum of the squares of the r.m.s. components.

1.9. Calculation of Power and Power Factor associated with Periodic Waveforms

Consider the voltage waveform and the current waveform to be available as Fourier Series of the same fundamental frequency ω_o as follows.

$$v(t) = V_{dc} + \sum_{n=1}^{\infty} V_n \sin(n\omega t + \alpha_n) \quad \text{and} \quad i(t) = V_{dc} + \sum_{n=1}^{\infty} I_n \sin(n\omega t + \beta_n)$$

$$p(t) = v(t).i(t) \qquad \text{and, average power } P \text{ is given by}$$

$$P = \frac{1}{T}\int_{to}^{to+T} v(t).i(t).dt = \frac{1}{T}\int_{to}^{to+T} [V_{dc} + \sum V_n \sin(n\omega\, t + \alpha_n)] \cdot [I_{dc} + I_n \sin(n\omega\, t + \beta_n)] dt$$

Using the trigonometric properties, it can be easily seen that only similar terms from v and Ican give rise to non-zero integrals.

$$P = V_{dc}.I_{dc} + \Sigma\, (1/2) V_n I_n \cos(\alpha_n - \beta_n) = V_{dc}.I_{dc} + \Sigma\, V_{rms,n} I_{rms,n} \cos(\alpha_n - \beta_n)$$

Thus the total power is given as the sum of the powers of the individual harmonics including the fundamental and the direct term.

The **overall power factor** of a periodic waveform is defined as the ratio of the active power to the apparent power. Thus

Overall power factor = $P/(V_{rms} I_{rms})$

In the case of non-sinusoidal waveforms, the power factor is not associated with lead or lag as these no longer have any meaning.

The **fundamental displacement factor** corresponds to power factor of the fundamental. It tells us by how much the fundamental component of current is displaced from the fundamental component of voltage, and hence is also associated with the terms lead and lag.

1.10. Analysis of Circuits in the presence of Harmonics in the Source

Due to the presence of non-linear devices in the system, voltages and currents get distorted from the sinusoidal. Thus it becomes necessary to analyse circuits in the presence of distortion in the source. This can be done by using the Fourier Series of the supply voltage and the principle of superposition.

For each frequency component, the circuit is analysed as for pure sinusoidal quantities using normal complex number analysis, and the results are summed up to give the resultant waveform.

1.11. Complex form of the Fourier Series

It would have been noted that the only frequency terms that were considered were positive frequency terms going up to infinity but that time was not limited to positive values.Mathematically speaking, frequency can have negative values, but as will be obvious,negative frequency terms would have a positive frequency term giving the same Fourier component. In the complex form, negative frequency terms are also defined.Using the trigonometric expressions

$e^{j\theta} = \cos\theta + j\sin\theta$ and $e^{j\theta} = \cos\theta + j\sin\theta$

we may rewrite the Fourier series in the following manner.

This can be re-written in the following form

$$f(t) = \frac{A_o}{2} + \sum_{n=1}^{\infty} (A_n \cos n\omega_o t + B_n \sin n\omega_o t)$$

$$f(t) = \frac{A_o}{2} + \sum_{n=1}^{\infty} A_n \cdot \left(\frac{e^{jn\omega_o t} + e^{-jn\omega_o t}}{2} \right) + B_n \cdot \left(\frac{e^{jn\omega_o t} - e^{-jn\omega_o t}}{2j} \right)$$

It is to be noted that B_o is always 0, so that the **j0** with A_o may be written as j B_o. Also

$e^{j0} = 1$

Thus defining;

$$C_n = \frac{A_n - jB_n}{2}, \text{ we have } C_0 = \frac{A_0 - j0}{2} \text{ and } C_{-n} = \frac{A_{-n} - jB_{-n}}{2}$$

the term on the right hand side outside the summation can be written as $C_0\, e^{j0}$ and the first term inside the summation becomes $C_n\, e^{-jn\omega_0 t}$

Since $A_n = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot \cos n\omega_o t \cdot dt$, $A_{-n} = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot \cos(-n\omega_o t)\cdot dt = A_n$

and $B_n = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot \sin n\omega_o t\cdot dt$, $B_{-n} = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot \sin(-n\omega_o t)\cdot dt = -B_n$

$$\therefore C_{-n} = \frac{A_{-n} - jB_{-n}}{2} = \frac{A_n + jB_n}{2}$$

That is,the second term inside the summation becomes $C_{-n}\, e^{-jn\omega_0 t}$.Thus the three sets of terms in the equation correspond to the zero term, the positive terms and the negative terms of frequency. Therefore the Fourier Series may be written in complex form as

$$f(t) = \sum_{n=-\infty}^{\infty} C_n \cdot e^{jn\omega_o t}$$

and the Fourier coefficient C_n can be calculated as follows.

$$C_n = \frac{A_n - jB_n}{2} = \frac{1}{2}\cdot\frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cdot[\cos n\omega_o t - j\sin n\omega_o t]\cdot dt = \frac{1}{T}\int_{t_o}^{t_o+T} f(t)\cdot e^{-jn\omega_o t}\cdot dt$$

CHAPTER TWO

HARMONICS

2.1 Definition of Harmonics

Harmonic is defined as a sinusoidal component of a periodic wave having a frequency that is an integral multiple of the fundamental frequency. For example, a component, the frequency of which is five times the fundamental frequency, is called a 5^{th} harmonic. The theoretical maximum amplitude of each harmonic current produced by a converter is equal to that of the fundamental component divided by harmonic order. For example, the 5^{th} harmonic is equal to 20 percent of the load current; and the 7^{th} harmonic is equal to 14.3 percent; and so on. These values are for an idealized square wave and, in practice, will be less because of system impedance. The harmonic components are assumed to be in phase with the fundamental. The resulting wave shape will depend on the magnitude and the phase relation of each of the harmonic components.Fig 2.1.1

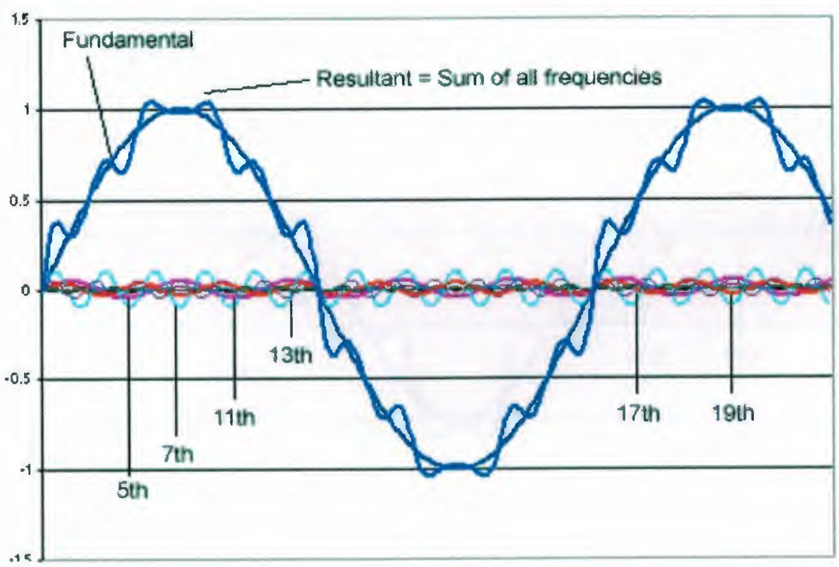


Fig 2.1.1 Harmonics component

Harmonic distortion factor (HDF) standards are needed to ensure that users are provided with a suitable voltage supply wave form; limit distortions to levels that system components can tolerate; and prevent the power system from interfering with the operation of other systems. In order to compare levels of harmonic distortion in a power system, the HDF is used, and is defined in the Institute of Electrical and Electronic Engineers (IEEE) Standard 519-1992 as:

$$HDF = \frac{(SumOfSquaresOfAllHarmonics)^{1/2}}{(SquareOfAmplitudeOfFundamental)} x100\%$$

The amount of voltage distortion that can be tolerated on a power system is dependent upon the equipment connected to it and this equipment's susceptibility to nonsinusoidal wave shapes. Power utility companies may be more stringent or relaxed in their specifications for the HDF, and may use different formulas than those given in IEEE Standards. In Canada, for example, the requirements for HDF vary from utility to utility, but in general they range from 1 to 5 percent depending on the system voltage level. The higher the voltage level, the more stringent the harmonic limitations requirements. It is, therefore, necessary to check with the power company as to their requirements in limiting harmonic voltages and currents as this may have substantial impact on the drive and filter design. IEEE Standard 519-1992 specifies guidelines with regard to limiting the harmonic voltage and current distortion factor.

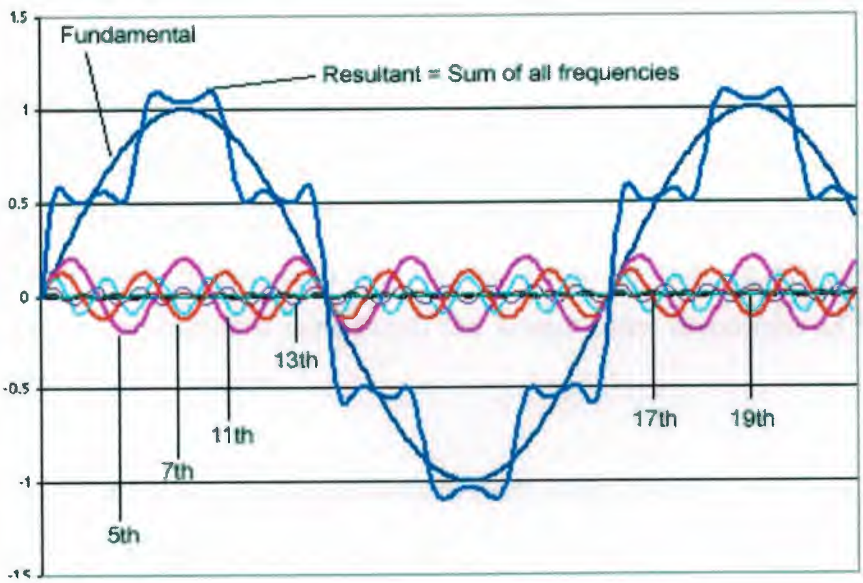


Fig 2.1.2 Highly distorted waveform

2.2 Harmonic sources

The common sources of harmonics in utility or industrial electrical systems are: rectifiers, dc motor drives, adjustable frequency ac drives, uninterruptible power supplies (UPS), arc furnaces, static volt amperes reactive (VAR) generators, cyclo converters, and static motor starters.

A static power converter generates harmonic currents the order of which is given by:

n = kp ±1

where: n = order of the harmonic

k = an integer 1, 2, 3 . . .

p = number of pulses of the converter system

Table 2.1.1 Voltage distortion limits for medium and high voltage power systems

Power System Voltage Level	Dedicated System Converter*	General Power System
Medium Voltage 2.4 – 69kv	8%	5%
High Voltage 115kv and Above	1.5%	1.5%

*A dedicated system is one servicing only converters or loads not affected by voltage distortion.

A 6-pulse converter would generate harmonic currents of the order 5th, 7th, 11th, 13th, 17th, 19th, 23rd, 25th, etc. For a 12-pulse converter configuration, the harmonics generated are 11th, 13th, 23rd, 25th, etc. Therefore, a 12-pulse converter system provides a significant reduction in the voltage distortion and, equally important, it eliminates (assuming balanced conditions) the lowest order harmonics of 5th and 7^{th} that are typical of most concern.

In addition there are other non-linear devices such as arc discharge devices used in arc furnaces and fluorescent lamps. Power supplies for electronic equipment such as UPS, numerical controlled machine and computers, and any load that requires other than a sinusoidal current will be a source of harmonic currents.

2.3 Harmonic technical history

From approximately 1910 to the 1960s, the main non-linear loads came from those few larger users in the electro-chemical and electro-metallurgical industries. They developed means of limiting the harmonic currents that their processes developed and thus minimized the effect on power systems and other users.

Small and medium sized adjustable speed drives used motor-generator (M-G) sets to feed dc motors and a few adjustable speed ac drives used wound rotor motors. Still other variable speed drives were steam driven. For the M-G sets, the mechanical linkage between the two systems transmitted power between them and at the same time electrically isolated each system from the other. However, these M-G sets were bulky and tended to be high maintenance pieces of equipment.

The first attempt at electrical rectification was accomplished through mechanical means. A motor driven cam physically opened and closed switches at precisely the right instant on the voltage wave form to supply dc voltage and current to load. At best, this approach was cumbersome since timing the switches and keeping them timed was extremely difficult. In addition, contact arcing plus mechanical wear also made this equipment a high maintenance item. Mechanical rectifiers were soon replaced by static equipment including mercury, selenium, and silicon diodes, thyristors and finally insulated gate bipolar transistors (IGBTs). With the invention and development of the thyristor, cost effective equipment became available to allow standard squirrel cage induction motors to drive pumps, fans, and machines with the ability to control the speed of these drives. The technology grew rapidly and the applications of these drives covered all aspects of process drives in all industries. These non-linear type loads increased dramatically in just the decade of the 1970s. This growth has continued and will continue.

Although solid-state rectification appeared to be the panacea to the problems of the older methods, other system problems soon became noticeable , especially as the total converter load became a substantial section of the total system power requirements.

The most noticeable initial problem was the inherent poor power factor associated with static power converters. Economics (utility demand billing) as well as system voltage regulation requirements made it desirable to improve the overall system power factor which normally was accomplished using shunt power factor correction capacitors. However, when these capacitor banks were applied, other problems involving harmonic voltages and currents affecting these capacitors and other related equipment became prevalent.

Another initial problem was the excessive amount of interference induced into telephone circuits due to mutual coupling between the electrical system and the communication system at these harmonic frequencies.

More recent problems involve the performance of computers, numerical controlled machines, and other sophisticated electronic equipment which are very sensitive to power line pollution. These devices can respond incorrectly to normal inputs, give false signals, or possibly not respond at all. More recently, neutrals of four-wire systems (480/277V; 120/208V) have been the latest power system element being affected by harmonics.

2.4 Resonance

The application of capacitors in a harmonic environment necessitates the consideration of the potential problem of an excited harmonic resonance condition. Inductive reactance increases directly with frequency and capacitive reactance decreases directly with frequency. At the resonant frequency of any inductive-capacitive (L-C) circuit, the inductive reactance equals the capacitive reactance.

Two forms of resonance which must be considered: series resonance and parallel resonance.

For the series circuit , the total impedance at the resonant frequency reduces to the resistance component only. For the case where this component is small, high current magnitudes at the exciting frequency will flow. From a practical viewpoint, series resonance conditions are source limited. As long as the harmonic source is not excessively large compared to the power factor capacitor bank, the harmonic current is typically within limits.

Parallel resonance is similar to series resonance where the capacitance reactance equals the inductive reactance. However, the parallel impedance is significantly different. At the resonant frequency the impedance is very high and when excited from a source at this frequency, a high circulating current will flow in the capacitance-inductance loop – although the source current is small in comparison. The most common situation, which leads to parallel resonance in industry, is where the source inductance resonates with the power factor capacitor bank at a frequency excited by the facilities' harmonic sources. In this situation, the harmonics are amplifie d by the resonant condition and are not source limited.

In actual electrical systems utilizing power factor correction capacitors, either type of resonance or a combination of both may occur if the resonant point happens to be close to one of the frequencies generated by harmonic sources in the system. The result may be the flow of excessive amounts of harmonic current and/or the appearance of excessive harmonic overvoltages. Possible consequences of such an occurrence are excessive capacitor fuse operation, capacitor failures, unexplained protective relay tripping, telephone interference, or overheating of other electrical equipment.

2.5 Electrical loads

Until recently, almost all loads were linear, and those that were not were a small portion of the total as to have little effect on system design and operation. Then came the electronic revolution and electronic loads, such as computers, UPS equipment, and variable speed motor drives, have proliferated. These electronic loads are mostly non-linear, and have become a large enough factor to have serious consequences in distribution systems. Overheated neutral conductors, failed transformers, malfunctioning generators, and motor burnouts have been experienced, even though loads were apparently well within equipment ratings. Motor, incandescent lighting, and heating loads are linear in nature. That is, the load impedance is essentially constant regardless of the applied voltage. For ac, the current increases proportionately as the voltage increases and decreases proportionately as the voltage decreases. This current is in phase with the voltage for a resistive circuit with a power factor of unity. It lags the voltage by some phase angle for the more typical partially inductive circuit with a power factor commonly between 0.80 and 0.95, and leads the voltage by some phase angle for the occasional capacitative circuit, but is always proportional to the voltage. For a sinusoidal voltage, the current is also sinusoidal and at the same frequency. Fig 2.5.1 and Fig 2.5.2

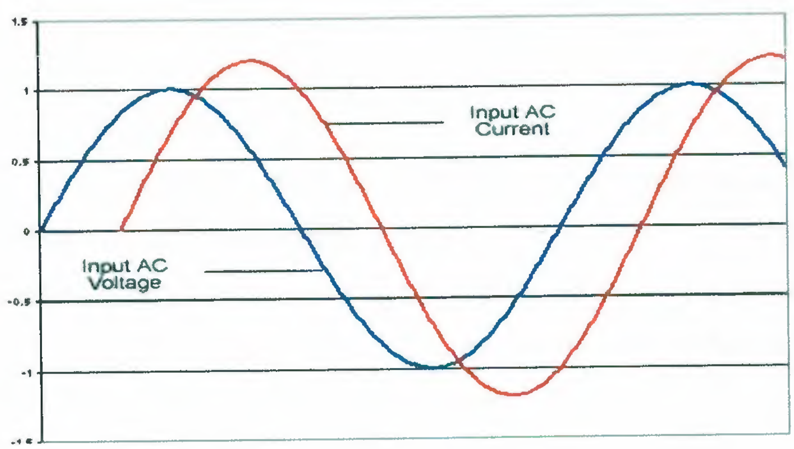


Fig 2.5.1 Linear Load Waveforms

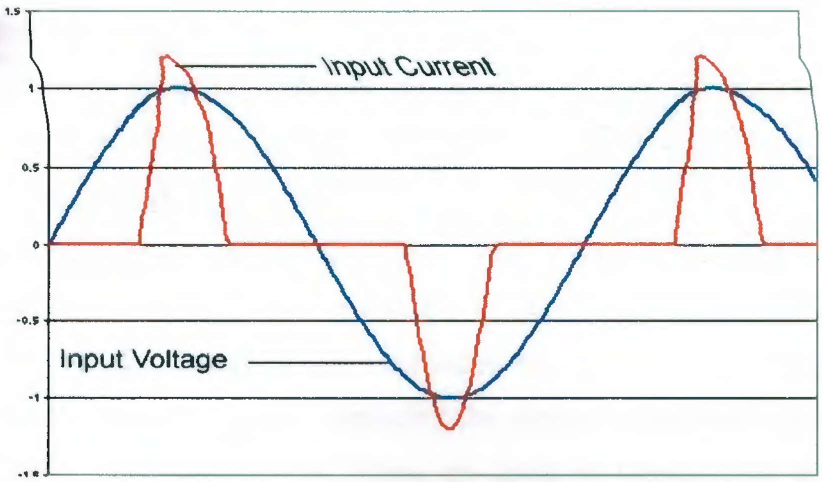


Fig 2.5.2 Non Linear Waveforms

A non-linear load is one in which the load current is not proportional to the voltage. Often, the load current is not continuous. It can be switched on for only part of the cycle, as in a thyristor-controlled circuit; or pulsed, as in a controlled rectifier circuit, a computer, or power to a UPS.

Non-linear loads can often create considerable harmonic distortion on the system. Harmonic currents cause excessive heating in magnetic steel cores of transformers and motors. Odd-order harmonics are additive in the neutral conductors of the system, and some of the pulsed currents do not cancel out in the neutral, even when the three phases of the system are carefully balanced. The result is overloaded neutral conductors. Also, many of these non-linear loads have a low power factor, increasing the cost of utility power where power factor penalty clauses apply.

Non-linear load currents are nonsinusoidal and even when the source voltage is a clean sinewave, the non-linear loads can distort that voltage wave, making it nonsinusoidal. It is essential that special characteristics of non-linear loads are understood so that failures on critical systems are avoided.

In rectifiers for dc loads and dc motor speed controls, the incoming ac is rectified and in many cases filtered to remove the ripple voltage.

In ac speed controls, the incoming ac is rectified to dc, which is then inverted by pulsing circuits back to adjustable -frequency ac. The same steps are used in UPS systems to obtain constant frequency 60 or 415 hertz ac power.

In power supplies for computers, office machines, programmable controllers, and similar electronic equipment, the ac is converted to low-voltage dc, with high-speed switching circuits for controlling the voltage. The dc is used directly by the microprocessors or central processing unit (CPU).

In the past, most motor-driven computer peripherals, such as tape drives and cooling fans, had ac motors. However, in the latest equipment these peripherals use dc motors, increasing the dc load of the computer system.

Conventional rectifier-type power supplies consist of a transformer to raise or lower the voltage, a rectifier, and filtering to remove the voltage variations or ripple from the dc output. Where voltage change is not necessary, the transformer may be eliminated. Voltage control can be obtained by replacing the diodes in the rectifier with thyristors [silicon-controlled rectifiers (SCRs)]. These are gated on (conducting) at any point in the cycle, turn off automatically as the current passes through zero, and are gated on again at the same point in each subsequent half-cycle.

Rectifiers use three-phase power for a 6-pulse circuit , or use transformers to increase the number of phases to create a 12- or 24- pulse circuit. The greater the number of pulses, the less filtering is needed to provide a smooth, ripple -free dc output. There are many variations of these basic rectifier circuits.

A characteristic of all rectifier circuits is that they are non-linear and draw currents of high harmonic content from the source. Diode full-wave rectifiers are least non-linear, conducting as soon as the forward voltage overcomes the small (about 0.7 V) forward bias required. Phase-controlled rectifiers using thyristors do not begin to conduct until gated on and are, therefore, more non-linear.

The standard power supply, with a transformer and an iron-core choke in the filter, is large, heavy, inefficient, and costly. Manufacturers of computers and other microprocessor-based electronic equipment have almost completely changed over to the switching-mode type of power supply, which eliminates the heavy iron-core input transformer and filter choke.

The switcher controls the voltage, switching at a frequency of from 20 to 100 kilohertz (kHz). A newer switcher operates in the megahertz (MHz) range. A transformer on the switcher output provides some voltage control and isolation of the load from the source. The high switching frequency means that the transformer can be small and light. It requires only a ferrite core instead of a steel core. Voltage sensors and control circuits vary the switcher duty cycle (on time) to produce the required output voltage under varying load conditions.

Switching mode power supplies (SMPS) are highly non-linear and a major source of harmonic distortion and noise. The high-frequency harmonics extend into the radio-frequency (RF) range, requiring most manufacturers to include filters in the incoming line to meet FCC requirements on limiting conducted and radiated interference. Modern *computers, from the individual PC to the largest mainframe, and most other* microprocessor-based electronic equipment use SMPS and are a major source of non-linear load problems.

2.6 Neutral currents

Multiples of the 3rd harmonic current are additive in the common neutral of a three-phase system, but the mechanism that causes this is little understood.

In a three-phase, four-wire system, single -phase line-to-neutral load currents flow in each phase conductor and return in the common neutral conductor. The three 60 hertz phase currents are separated by 120°; and for balanced three-phase loads, they are equal. When they return in the neutral, they cancel each other out, adding up to zero at all points. Therefore, for balanced three-phase, 60 hertz loads, neutral current is zero.

For 2nd harmonic currents separated by 120°, cancellation in the neutral would also be complete with zero neutral current. This is true in the same way for all even harmonics.

For 3rd harmonic currents, the return currents from each of the three phases are in phase in the neutral and so the total 3rd harmonic neutral current is the arithmetic sum of the three individual 3rd harmonic phase currents. This is also true for odd multiples of the 3rd harmonic (9th, 15th, 21st, etc.).

The theoretical neutral current with harmonics is at least 1.73 and perhaps as much as 3.0 times the phase current. For pulsed loads, the pulses can occur in each phase at a different time. They will return in the common neutral, but they will be separated by time; therefore, there will be no cancellation. If none of the pulses overlap, the neutral current can be three times the phase current.

The effects of additive harmonics in the neutral were first recognized in the National Electrical Code (NEC) many years ago, when Section 220-22 prohibited reduced neutral conductor size for that portion of the load consisting of discharge lighting. The effects of electronic equipment were recognized in the 1987 NEC when the prohibition in Section 220-22 against reducing the neutral was expanded to include non-linear loads.

There are several solutions to minimize this problem. Not only must neutral conductor sizes not be reduced for these loads, but they must often be increased. Many engineers are designing with neutral conductors sized for at least 150 percent of the true root mean square (RMS) phase current, including the harmonic content.

The harmonic content of the loads may be reduced by means of line filters. Since the manufacturers of the electronic equipment seldom install line filters beyond the minimum necessary to meet FCC requirements, the power-line filters must usually be separate units installed between the source and the loads.

For large computers, UPS, or other non-linear loads, the final isolation transformer should be located as close as possible to the load. The neutral conductors from the wye secondary must be oversized as noted, but the conductor lengths would be relatively short. Nothing upstream from the transformer will be affected. However, the transformer may have to be derated.

2.7 Derating power equipment

The ratings of transformers and generators are based on the heating created by load currents of an undistorted 60 hertz sinewave. When the load currents are non-linear and have a substantial harmonic content, they cause considerably more heating than the same number of amperes of pure sinewave. There are two major reasons for this.

When steel is magnetized, the minute particles known as magnetic domains reverse direction as the current alternates, and the magnetic polarity also reverses. The magnetizing of the steel is not 100 percent efficient, since energy is required to overcome the friction of the magnetic domains. This createshysteresis losses which are greater for a given RMS current at the higher-frequency harmonics, where themagnetic reversals are more rapid than at the fundamental 60 hertz.

Also, alternating magnetic fields induce currents into the steel laminations when the changing magnetic flux cuts through a conductor. These "eddy currents" flow through the resistance of the steel, generating eddy-current heating losses. Because of the higher frequencies, eddy-current losses are considerably greater for harmonic currents than they are for the same RMS value of 60 hertz current. A lesser, but still considerable, heating effect at higher frequencies is caused by the "skin effect" in the conductors. Currents at higher frequencies are not distributed evenly through the cross-section of the conductor. The magnetic fields tend to force the current flow toward the outside or skin of the conductor. This effect increases as the frequency increases, and also as the magnitude of current increases. At higher frequencies, the center of the conductor carries little or no current. Therefore, the effective cross-section of the conductor is decreased, and its resistance is increased. It behaves as a smaller conductor of lower capacity. As a result, a given current at harmonic frequencies causes more conductor heating than the same current at 60 hertz.

The result of hysteresis, eddy current, and skin effect is that the transformer or generator carrying no more than its full-rated RMS current, but supplying non-linear loads with a high harmonic content, will overheat, sometimes to the point of failure. Transformers and generators loaded to less than 70 percent of their rating have been shut down because of over-temperature. Rectifier transformers specifically designed for non-linear industrial rectifier loads have been manufactured to reduce these effects. At this time, transformers specially designed for other electronic loads are not available, and standard transformers must be derated.

As the harmonic currents are drawn by the loads, they act on the impedance of the source, causing harmonic distortion of the source voltage. Motors are normally linear loads, but when the supply voltage has harmonic distortion, the motors draw harmonic currents. These harmonic currents cause excessive motor heating from higher hysteresis and eddy-current losses in the motor laminations and skin effect in the windings. Thus, motors supplied from sources with voltage distorted by other non-linear loads will also overheat unless they are derated.

The solutions to overheating of transformers, generators, and motors as a result of non-linear loads are the same as those for neutral overheating. The equipment must be derated or the harmonic content must be reduced by line filters, or both. There are no standards for the required derating, although considerable research is being done to determine these requirements. Derating can be done by observation, based on the temperature rises of the affected equipment. In initial design, equipment must be oversized by an amount determined by judgement and experience to permit the necessary derating.

2.8 Generator control problems

Harmonic currents can cause serious problems for generator installations in addition to excessive heating. Modern generators use electronic means to regulate the output voltage of the generator, to control the speed of the engine or prime mover (thus the output frequency of the generator), to parallel generators, and to share the load proportionately among the paralleled units.

Many of these control devices use circuits that measure the zero crossing point of the voltage or current wave. At 60 hertz this is acceptable; but with a high harmonic content, there may be many more zero crossings than the normal ones for 60 hertz. This can cause hunting and instability in speed and frequency control, and can make the *paralleling of generators difficult or impossible.*

Load sharing depends on measurement of the load on each unit. The RMS value of the current is simple to determine for a pure 60 hertz sinewave, but using controls based on 60 hertz RMS where harmonics are present will give false readings, sometimes too high and at other times too low. Only more complex true RMS measurements will provide proper operation.

Therefore, it is urgent that the generator and control manufacturers are informed of the load characteristics if a generator is to be used alone or in parallel with non-linear loads. If this is not done, the installation may not perform properly and it may be costly to obtain correct operation.

2.9 UPS output harmonic distortion

UPSs are used to supply clean power to computers under all conditions, including total utility power failure. They range from large systems of thousands of kVA for major computer installations to small units of a few hundred VA for PCs. If the loads distort the power supplied by the UPS, then the power fed to the loads will not be truly "clean."

When harmonic currents are drawn by the load, they cause voltage distortion of the source. Since the voltage drop across the source for a given current is proportional to the impedance of the source (E = I x Z), the distortion caused by a given harmonic current is lower for a low-impedance source and higher for a high-impedance source.

The total harmonic distortion (THD) of the UPS for a given load depends on the UPS design and output impedance. This is true whether the UPS is the static type or the rotary M-G type. Most UPS manufacturers specify the output distortion of their equipment; <5 percent total harmonic distortion (as a percentage of the fundamental) is typical. However, many manufacturers add a disclaimer, such as "based on linear loads" or "for reactive and inductive loads." Such a disclaimer means that the THD figure only applies under linear load conditions. Before purchasing any UPS, make certain that it is capable of supplying the actual types of non-linear loads to be connected to it. Discuss the prospective loads with the manufacturer because correcting problems may be costly. UPS equipment utilizing IGBT rectifier technology, working at a 6 kHz frequency avoids harmonic contamination of the source. Input current wave forms have been measured at less than 3 percent THD at 100 percent load and a maximum of 5 percent at 50 percent load. These low levels of harmonics are achieved without the use of any passive harmonic resonant filtering. The input power factor of IGBT rectifiers is 0.98 lagging, so there is no need for power factor correcting capacitors.

CHAPTER THREE

SOLUTION OF HARMONICS PROBLEMS

3.1 AC system response to harmonics

An ac system's response to harmonic currents can be very complex. If only one steady-state sinusoidal voltage or current were applied, ohm's law (E = IZ) could be applied. The results of Fourier analysis of complex current wave forms indicate that such a simple relationship is not enough to describe the networks involved. Any number of inputs can be imposed upon a network so a more complete analytical tool is needed. Such a tool is that of complex functions.

Using the complex function H(s), with s as the complex frequency, its value (real or complex numbers) can indicate constants, linear changes, exponentials, sinusoids or any frequency, or any conceivable input in any combination. If a network with only one pair of input terminals is present, the response is H(s) = E(s)/I(s) = Z(s), the input impedance. Likewise, H(s) = I(s)/E(s) = Y(s), which is the input admittance.

In both cases, the network function H(s) relates the voltage and current (complex or real quantities) at the same pair of terminals. It is also called a driving-point or input function. The units of the driving-point function may be either ohms or mhos.

The forced response and the form of the free response or transient behavior can be found from the appropriate driving-point function. For a current input, the nature of the free response, which is the only response when the input is opened, is determined by the poles of Z(s). For a voltage source, the free response, which is the only response when the source is shorted out depends upon the poles of Y(s).

Poles are not the only points of importance. Another phenomenon called a zero is also of interest. To understand exactly what the poles and zeros mean, we must first realize that all circuits and networks contain capacitors (C), inductors (L), and resistors (R); either placed there intentionally or simply there as a consequence of distributed parameters. The R, L, and C components combine in various ways to form factors of H(s). Suppose that H(s) = Z(s) for a particular R-L-C network. If Z(s) has poles at particular points of s, it means that the impedance of the network is infinite or at least extremely high for most practical circuits. This means if any current at all can be forced into the network, its voltage response will tend toward infinity. On the other hand, if H(s) has zeros at some points of s, this means that the impedance of the network has

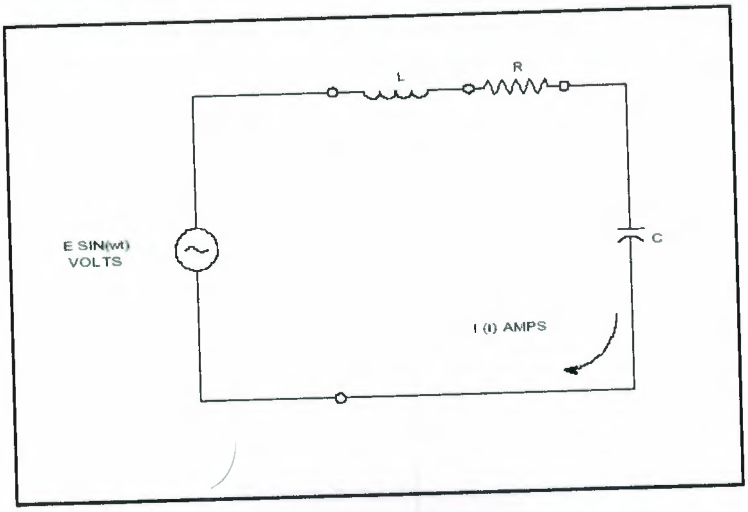


Fig. 3.1.1 Series RLC circuit

The location of the poles and zeros is plotted on the complex frequency or s plane as shown in figure 3.1.1

The extreme resolving power of H(s) can be visualized in figure 3.1.3 H(s) describes a complex surface consisting of peaks, which are the poles and depressions, which are the zeros. The value of –X, which is represented on the –b axis, describes how the circuit responds to transient inputs. If the capacitor and the inductor remain constant and the resistor value varies, several things can happen. If R is infinite, then –X is infinite and all responses of the circuit are damped to zero. If R is very large compared to L and C, the circuit will be overdamped or heavily damped. When X is small, meaning R is small, the circuit becomes underdamped.

Finally, if all resistance could be removed from the circuit (R = 0-ohms), the poles would move to the jω-axis. In this theoretical case, the circuit would become an oscillator whose frequency would be ωo. The oscillation would start as soon as any energy was introduced. In practice there is always some resistance in the circuit so the circuit response would eventually die out.

The damping factor describes how the energy in the circuit is redistributed or dissipated when the input is changed or removed. This distribution and/or dissipation of energy has to occur because the inductors and capacitors act as energy storage devices. Consequently, the network's response cannot stop when the input is removed if energy has been stored in the network.

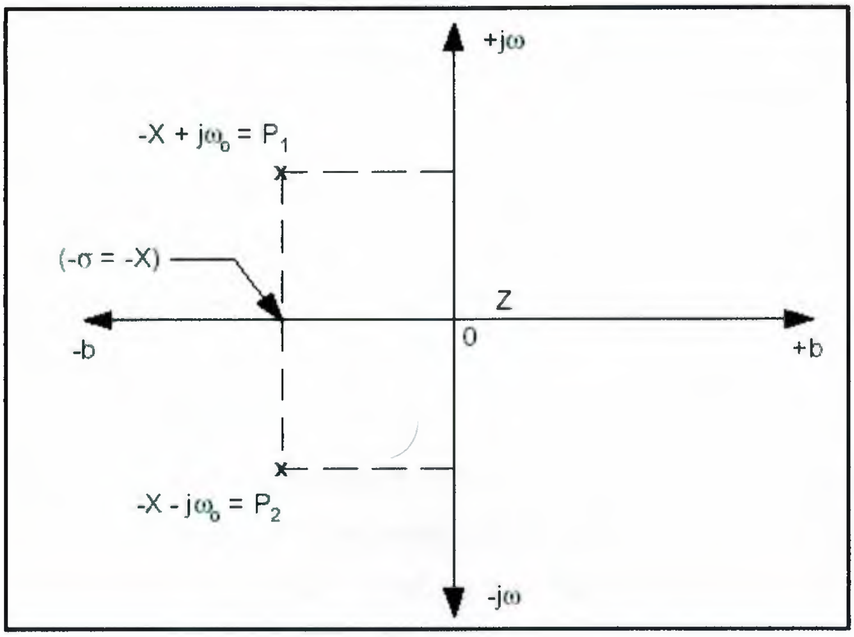


Fig. 3.1.2 Poles and zeros on the s-plane

From Fig 3.1.3, the cutaway view made along the +jw axis shows the network response to the steady-state source. That is b = 0, so s = ±jω only. Figure 3.2.1 shows the magnitude of H(jω) for any positive frequency, ω. This is the same curve traced out by the right-hand edge of the complex surface in figure 3.1.3 Finally, figure 3.1.5 shows the impedance Z(jω) which is the inverse of figure 3.2.1.

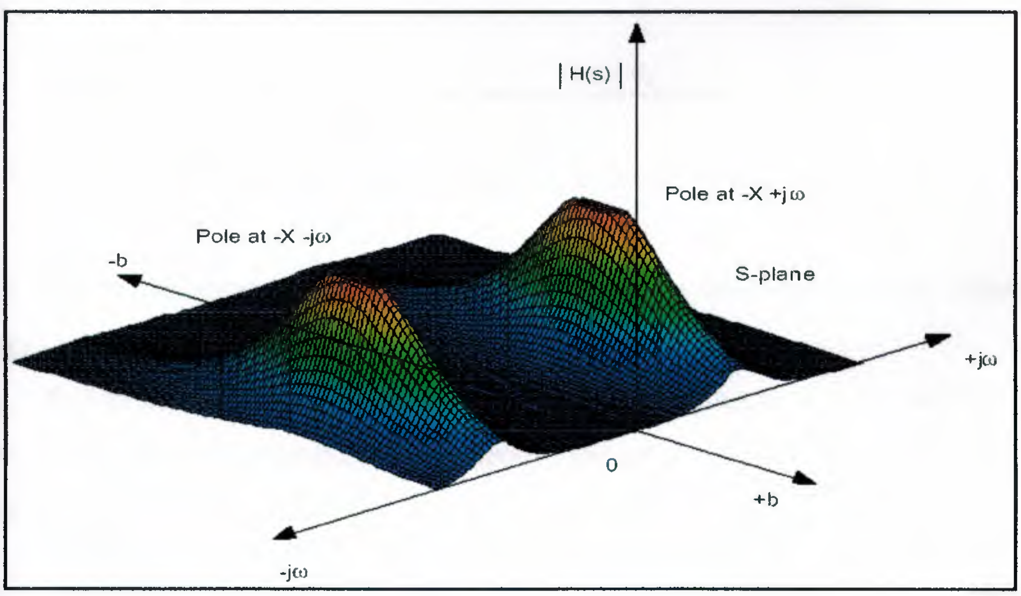


Fig. 3.1.3 Surface view of H(s) for all complex frequency in s-plane

The most important point for either Y(jω) or Z (jω) occurs at the resonant-frequency, $\omega_{\circ}$. Here, the effects of the inductor and capacitor exactly cancel each other; leaving only the conductance 1/R or the resistance R, respectively.

Only at this point is the power factor unity, figure 3.1.5. For frequencies lower than $\omega_{\circ}$, the network looks capacitive (generates VARS); and for frequencies higher than $\omega_{\circ}$, the network looks inductive (absorbs VARS).

Summarizing transient behavior:

1) The impedance of the network is not constant with respect to frequency.

2) The impedance Z(s) or admittance Y(s) functions develop drastic changes at various critical, complex frequencies, which are the roots of the poles and zeros.

3) The values of R, L, and C will change the network function if any one value is changed.

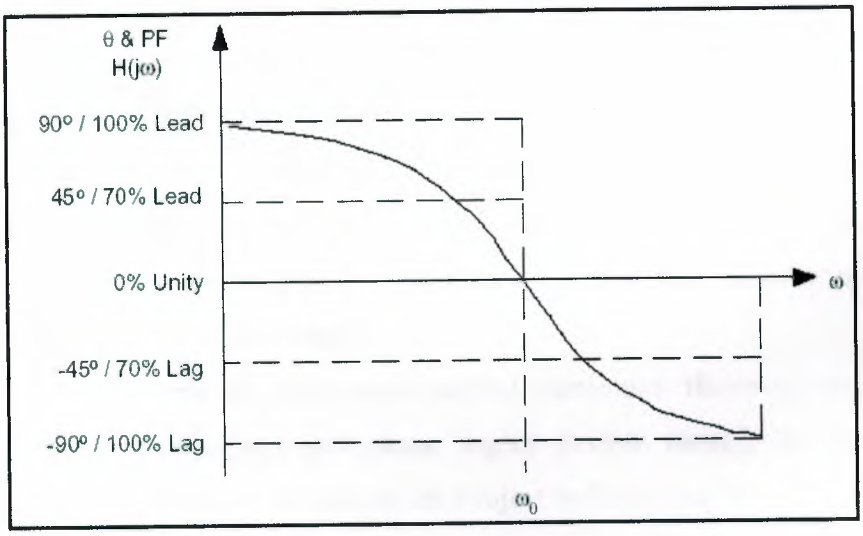


Fig. 3.1.4 Phase angle and power factor vs. frequency

4) The damping, or energy storage manifesting itself as transient behavior , changes with R/L or RC depending on the network.

5) The network function H(s) becomes more complex with each addition of an inductor , resistor, or capacitor; consequently, the source and/or load impedance will affect the network as well.

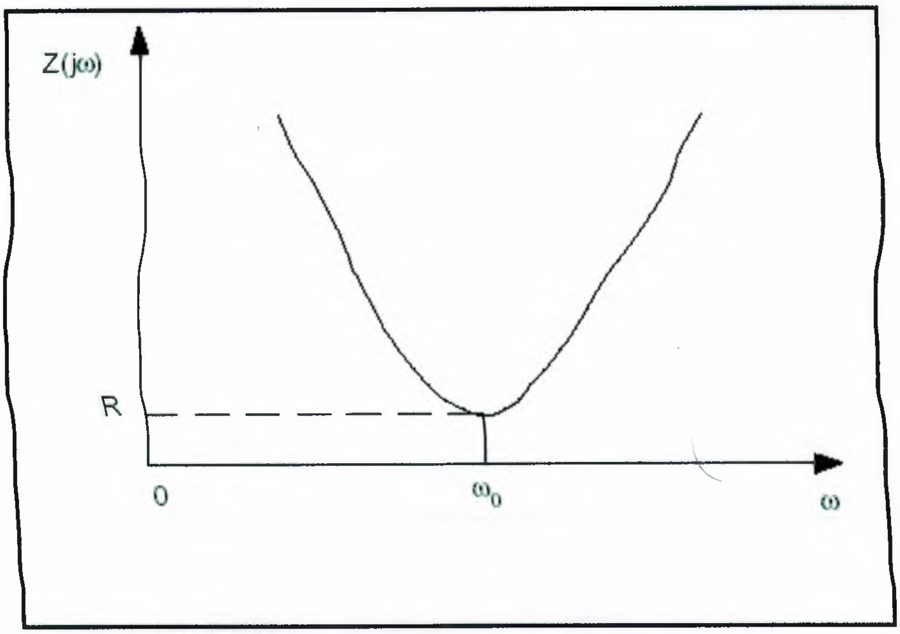


Fig. 3.1.5 Impedance vs. Frequency

3.2 *Solution of harmonic problems*

The use of phase multiplication should be considered in the design stages of a particular power supply. The phase multiplication technique requires special transformer configurations. Retrofits to existing power supplies, i.e., user loads would entail a cost far outweighing the benefits of the scheme.

Phase multiplication theoretically will cancel normal harmonics. However, in practice, both current sharing distributions and phase angles deviate enough to allow only incomplete cancellation. Most references on the subject indicate that 10 to 25 percent of the maximum harmonic magnitudes will remain.

Once systems without higher order phases are in place, there is little the power system engineer can do but resort to external filtering or real-time cancellation methods.

Tuned passive filters contain several branches of series R-L-C elements. These are connected in parallel and externally to the non-linear load with the objective of filtering the major harmonics generated by the load.

The harmonic currents do not "magically" disappear; rather they flow through the branches whose impedances have been set to result in a minimum voltage distortion response.

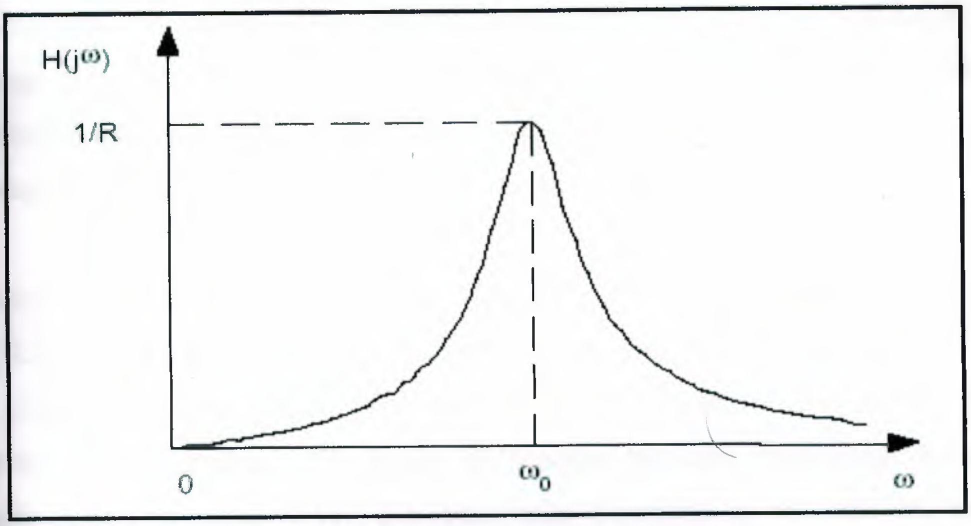


Fig. 3.2.1 Response vs. frequency for steady state input

An example of a tuned passive filter containing elements for the 5th and 7th harmonics and a high pass filter for all the higher harmonics is shown in figure 3.2.2 The filters represent a constant net capacitive VAR generation at the fundamental frequency. The inductor Lp is added to re-establish an overall unity power factor.

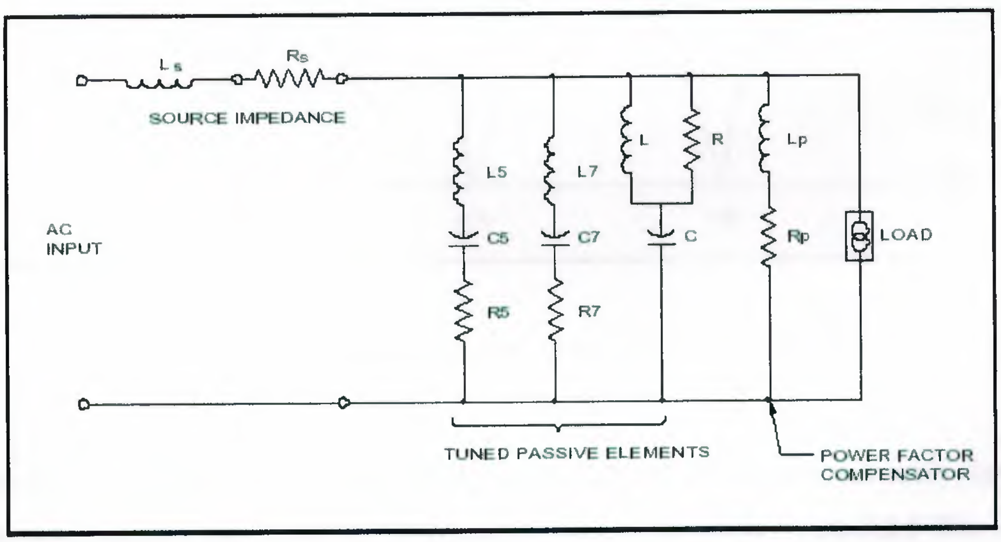


Fig. 3.2.2 Tuned passive filter

An examination of figures 3.2.2 and 3.2.3 together will show the complexity of this arrangement. Experience shows there will be eight poles and at least four zeros within the range of harmonic frequencies normally encountered. The filter is located, selected, and tuned with both the load characteristics and source impedance (Ls and Rs) in mind. Changing the load, the source, or the filter's location can result in degraded or unpredictable performance. The units can be costly to design and set, bulky, frequency and component variation sensitive, and intolerant of major system design changes.

A passive filter configured for 3rd, 5th, 7th, 9th, and higher harmonics is shown in figure 3.2.4 The phase control of the current through the inductor Lp can continuously adjust *filter/load power factor. With careful design, placement, and tuning, it will* effectively handle higher kVA loads of 6-pulse rectifiers exhibiting incomplete harmonic cancellation. However, the drawbacks discussed in the previous sections should not be underestimated.

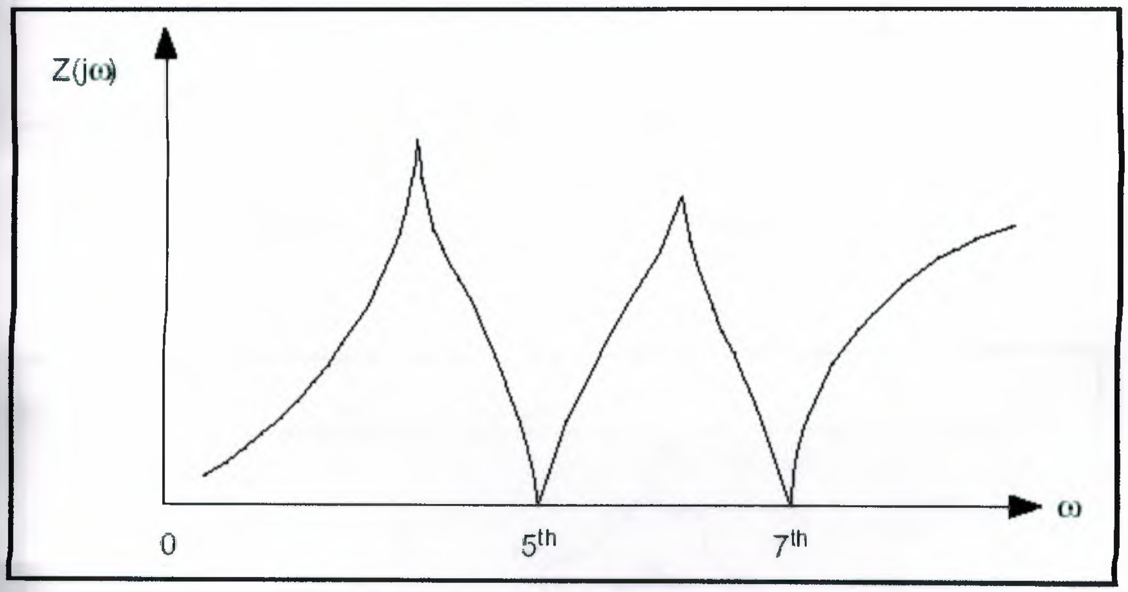


Fig. 3.2.4 Impedance poles and zeros of tuned passive filter

The active load current compensator is shown in figure 3.2.6 It performs every function of the tuned passive devices with VAR compensation shown in figure 3.2.5 The active technique requires an efficient, high speed, switching amplifier. This technique is superior to passive filtering in the following areas.

(1) It will power-up/power-down without transients.

(2) Will go into a self-protection mode should a load short-circuit (crowbar) appear.

(3) Can tolerate some variation in frequency without compromising performance.

(4) Will respond to voltage step variation or load current step variation without unwanted transients.

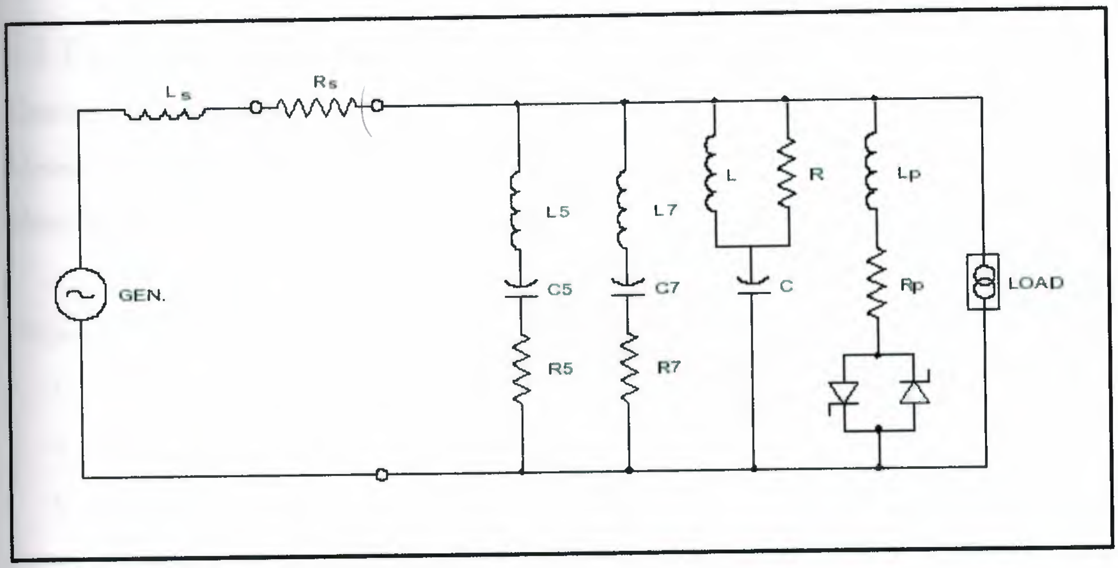


Fig. 3.2.5 Tuned passive filter with static VAR control

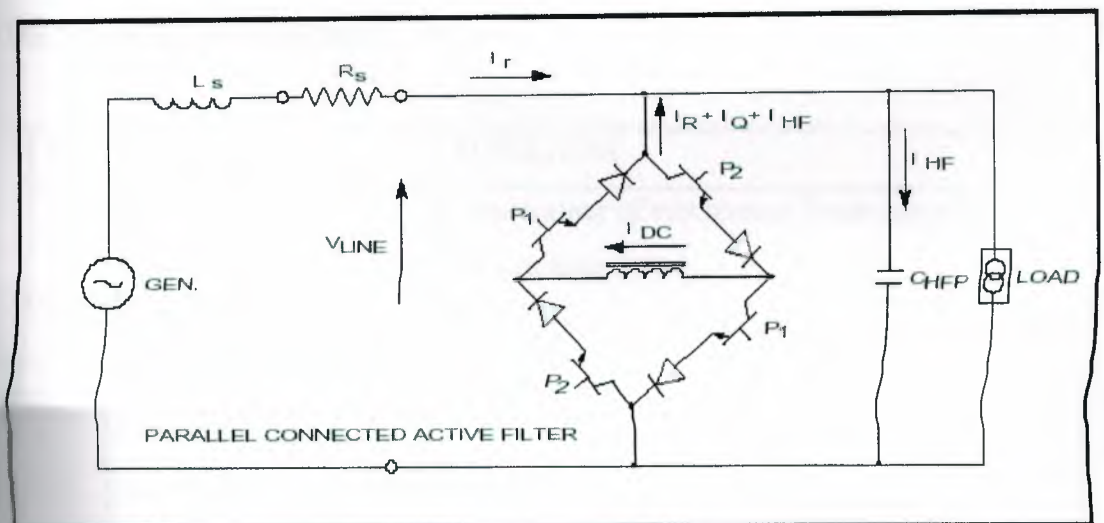


Fig. 3.2.6 Active load current compensation

(5) Effectively cancels the harmonic currents at its output node by "real-time" addition and subtraction. In effect, it "linearizes" the load's non-linear current.

CHAPTER FOUR

MEASUREMENT AND ANALYSIS OF HARMONICS IN A LOW VOLTAGE ELECTRICAL ENERGY COMPOUND

4.1 Example Application

General Information

Measurement Date : 30/12/2004 – 01/01/2005

Measurement Device: Energy Test 2020E

Objective of Measurement;

- Determining power factor
- Investigate the harmonic structure.
- Analysis of network during alternating loads.
- Determining the possibilities of enlarging the network.
- The control the voltage waves.

Measured Loads:

The following were measured .

NO	NAME	AÇIKLAMA
1.	Transormer-1 Mechanical	Transformer of mechanical loads composition not available
2.	Transformer-Illumination	Transformer of illumination loads composition not available
3.	UPS entry	
4.	UPS exit	

Measurement Condition:

Measurement data were recorded as an excel file.

Measurement time :

<u>Transformer-2 Main Entry</u>	30.12.2004	14:41 - 14:56
	Recording interval	5 second
	Recording number	184
<u>UPS Entry</u>	30.12.2004	15:48 - 15:58
	Recording interval	5 second
	Recording number	128
<u>UPS Exit</u>	30.12.2004	16:18 – 16:27
	Recording interval	5 second
	Recording number	112
<u>Transformer -1 Mechanical</u>	30.12.2004	6:56 – 17:02
	Recording interval	5 second
	Recording number	75
<u>Transformer -1 Mechanical</u>	30.12.2004 – 31.12.2004	17:28 – 21:26
	Recording interval	60 second
	Recording number	1680
<u>Transformer -2 Illumination</u>	31.12.2004 – 01.01.2005	21:53 – 11:43
	Recording interval	30 second
	Recording number	1662

Transformer 1; 800 kVA power,supplies two parallel connected machine departments.

Machine deparments load approximately 1166kW.

Transformer 2; 800kVA power,supplies mainly illumination loads.

4.2 Transformer 1- Mechanical

Power Factor :

Average power factor varies between 0.57-0.64. Power factor should be greater then 0.949. Ideally 1.00.

Voltage :

Neutral phase average voltage measured at ~ 250V.

A 10 V variation of voltage in neutral – phase was measured .

Variation in voltage is tolerable

Network Load :

Load in transformer were %40 during measurements.

Reactive Energy :

Average of total reactive power = 170kVA.

<u>During Measurements ;</u>

Apperent power of loads on transformer = **210-245kVA.**

Active power of loads on transformer = **120-150kW**

Reactive power of loads on transformer = **160- 180kVAr**

The ratio of active power to apparent power becomes lower than expected value because loads on transformer obtain power from network.

Obviously load has a power factor value as low as **0.57-0.64** (Fig 4.2.3). Harmonic voltage distortion could be neglected in 300A current applied transformer that has very low load value as obtained above.

Harmonic current distortion is between **%5 - %9** (Fig 4.2.1 –Fig 4.2.2)

The value of 5^{th} harmonic current is above the limitation of IEEE 519-1992 Standard.

Reactive power demand, taking into consideration the existing harmonic in the network; Altough there is no reactive penalty problem (according to the regulation of this country). Power demand is using compensation system of suitable power and type.

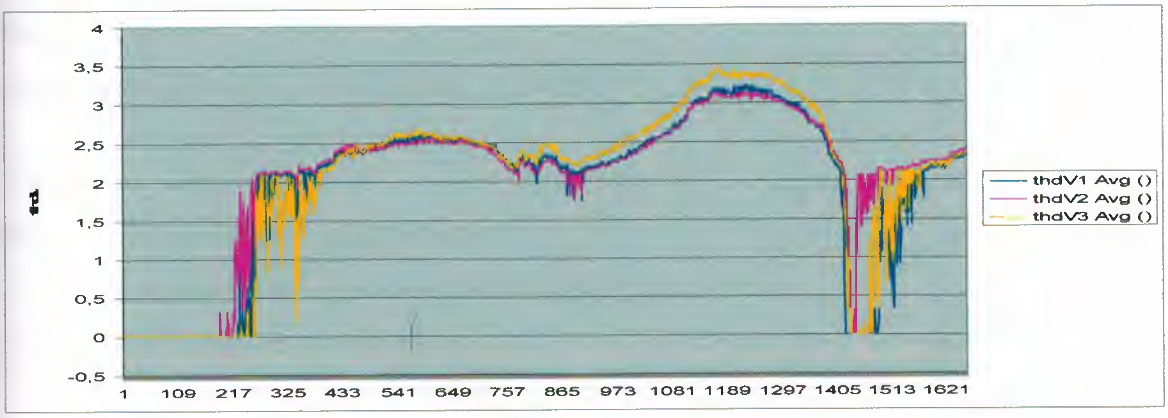


Fig 4.2.1 Transformer -1 Total harmonic distortion of voltage

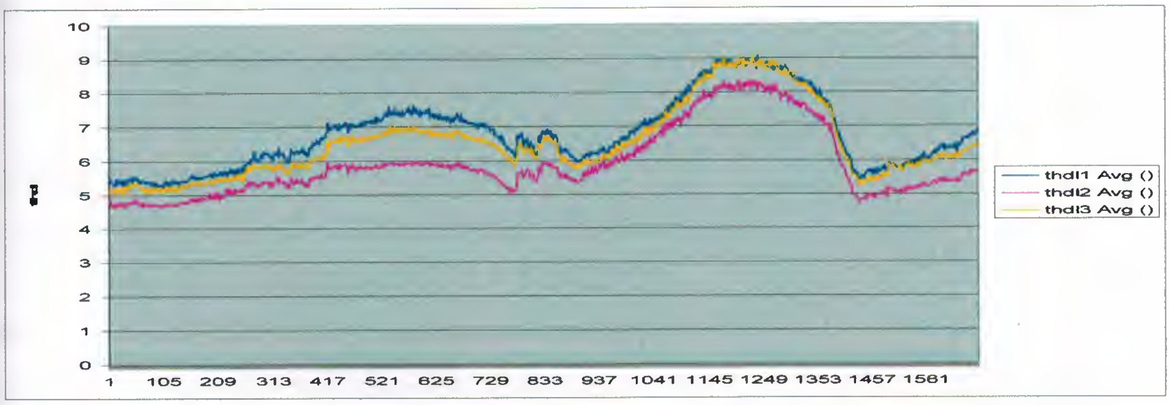


Fig 4.2.2 Total harmonic distortion of current

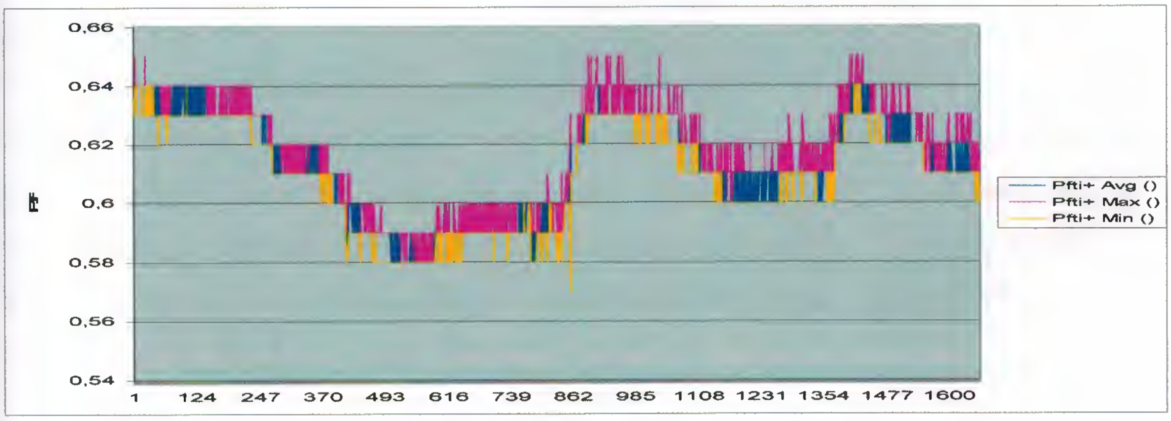


Fig 4.2.3 Power factor

4.3 Transformer 2- Illumination

Power Factor :

Average power factor varies between 0.93-0.97.

Voltage :

Neutral phase average voltage measured at ~ 245V.

A 10 V variation of voltage in neutral – phase was measured.

Variation in voltage is tolerable

Network Load :

Load in transformer were %50 during measurements.

Reactive Energy :

Average of total reactive power = 120kVA.

<u>During Measurements ;</u>

Apperent power of loads on transformer = **240-350kVA.**

Active power of loads on transformer = **235-325kW**

Reactive power of loads on transformer = **60- 120kVAr**

Value of power factor was observed between **0,93 – 0,97** .

In this load value ,20 % harmonic currents distortion was obtained when 500 A current applied to the transformer (harmonics 3, 5, and 7) (Fig 4.3.4-Fig 4.3.5). The major components are 3^{rd} and 5^{th} harmonics.These harmonics cause 8% distortion on phase-neutral voltage.(Fig 4.3.7-Fig 4.3.8)

This values are indication of an expected characteristic of loads that are mainly illumination loads.3^{rd} harmonics is obtained in three different phase that are produced in the some phazer according to the characteristic of loads. Neutraly ,these harmonics in three phase will combine and flow forward to the neutral base.(Fig 4.3.1—Fig 4.3.2)

Therefore ,185 A current is measured on neutral base.This value created by 3^{rd} harmonic (150 Hz) not by 50 Hz.(Fig 4.3.3 -4.3.6)

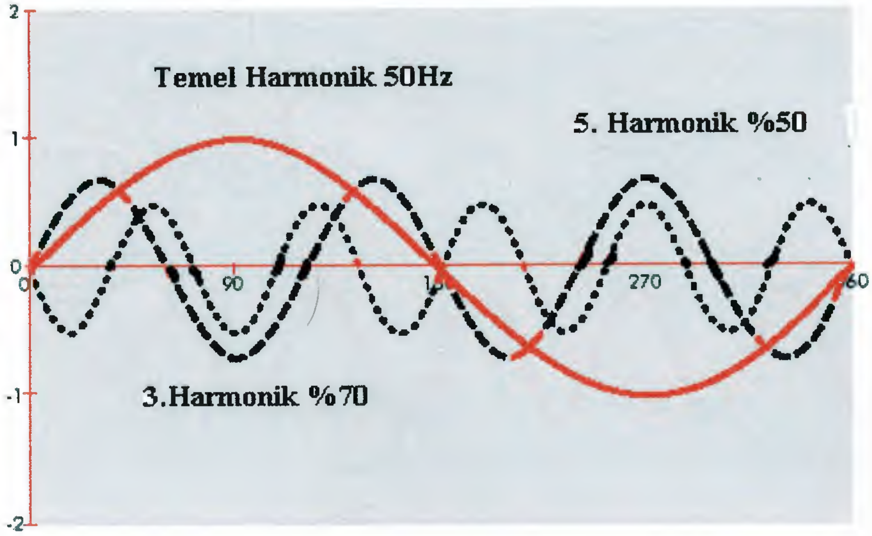


Fig 4.3.1 Fundamental harmonic ,3rd & 5th harmonics

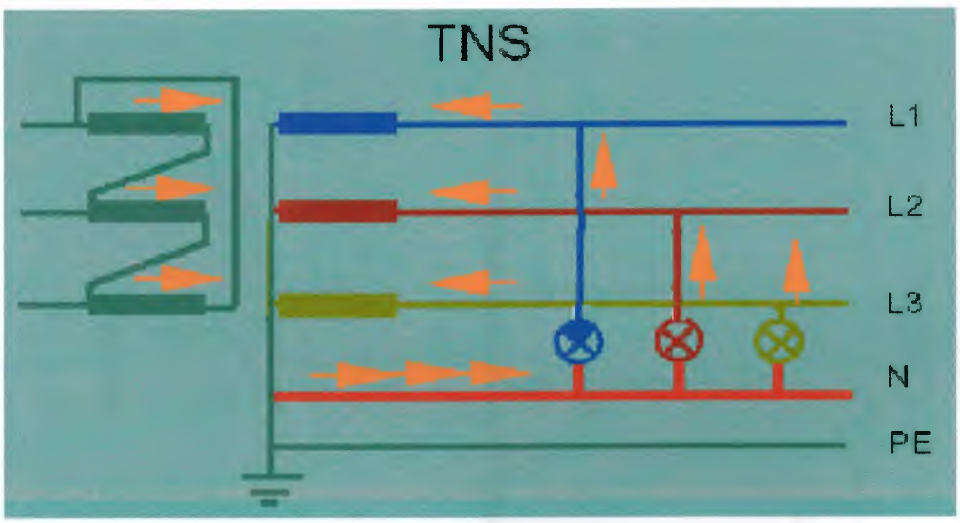


Fig 4.3.2 Harmonic currents

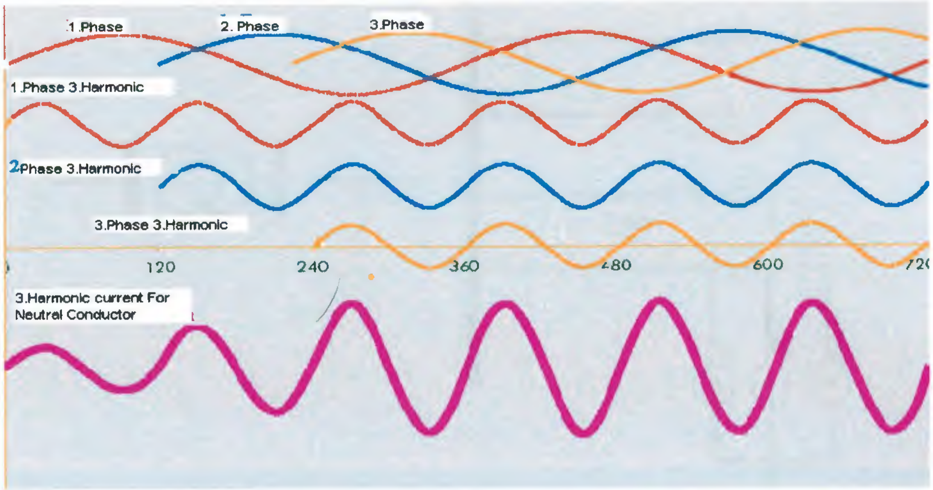


Fig 4.3.3 Neutral current

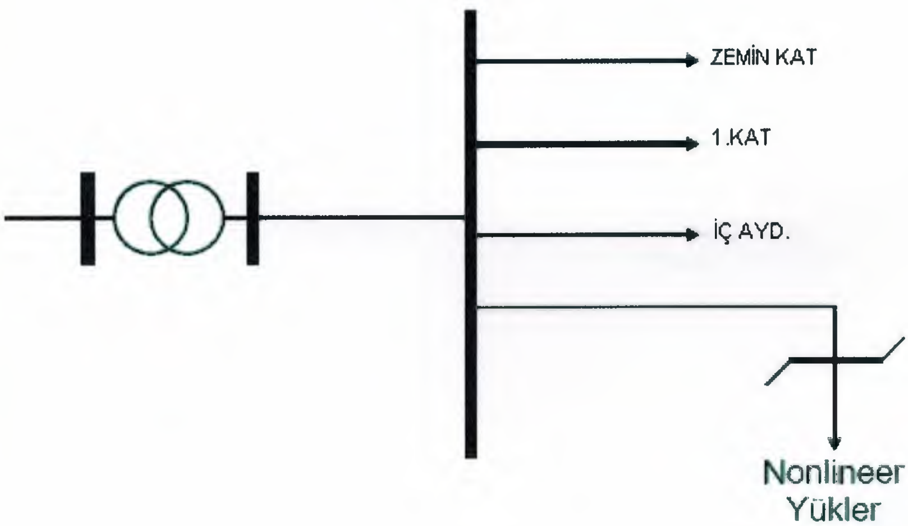


Fig 4.3.4 Transformer 2

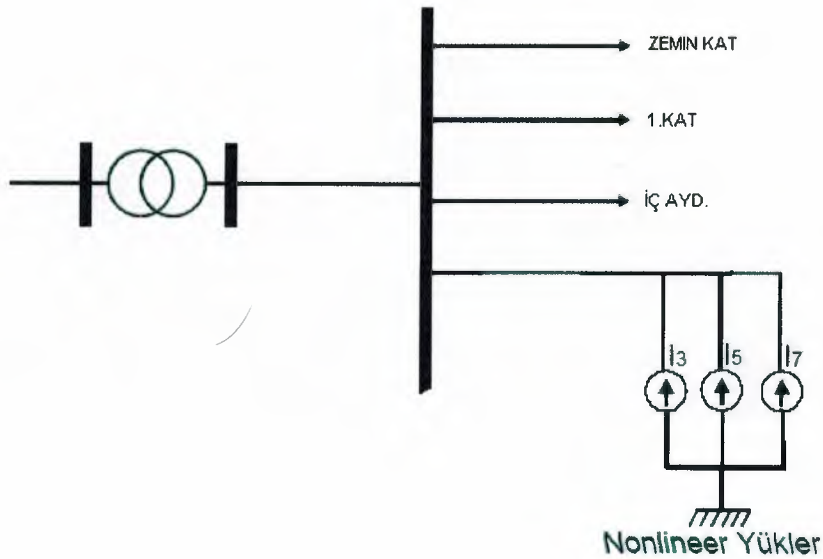


Fig 4.3.5 Transformator 2

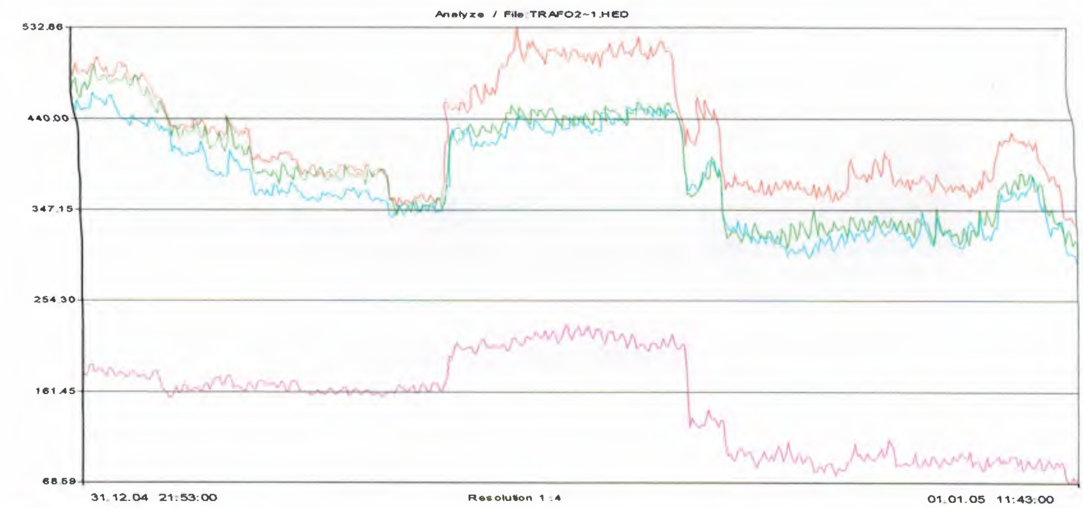


Fig 4.3.6 Transformer -2 Phaze currents & neutral current

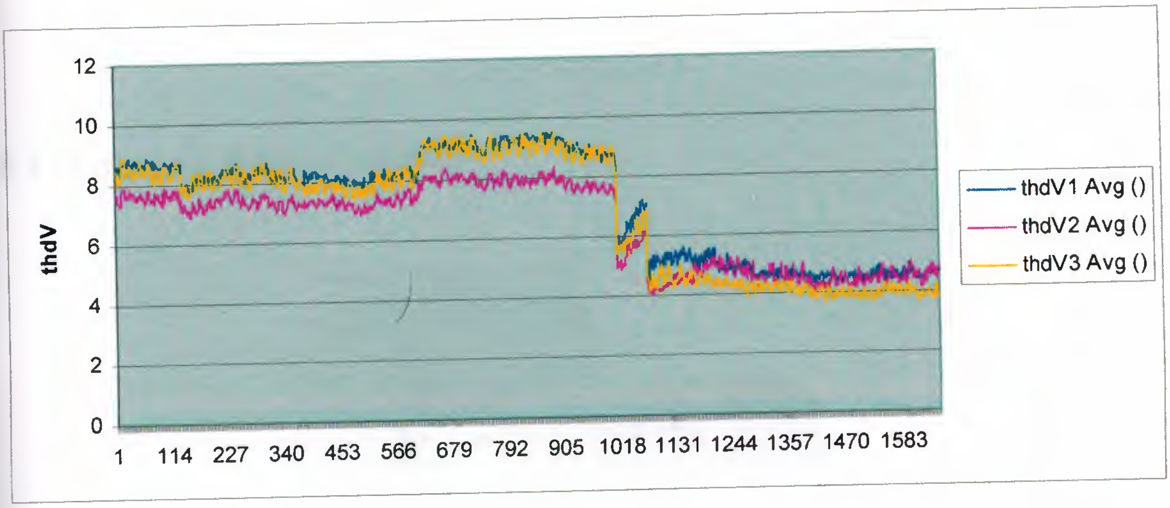


Fig 4.3.7 Transformer -2 total harmonic distortion of voltage

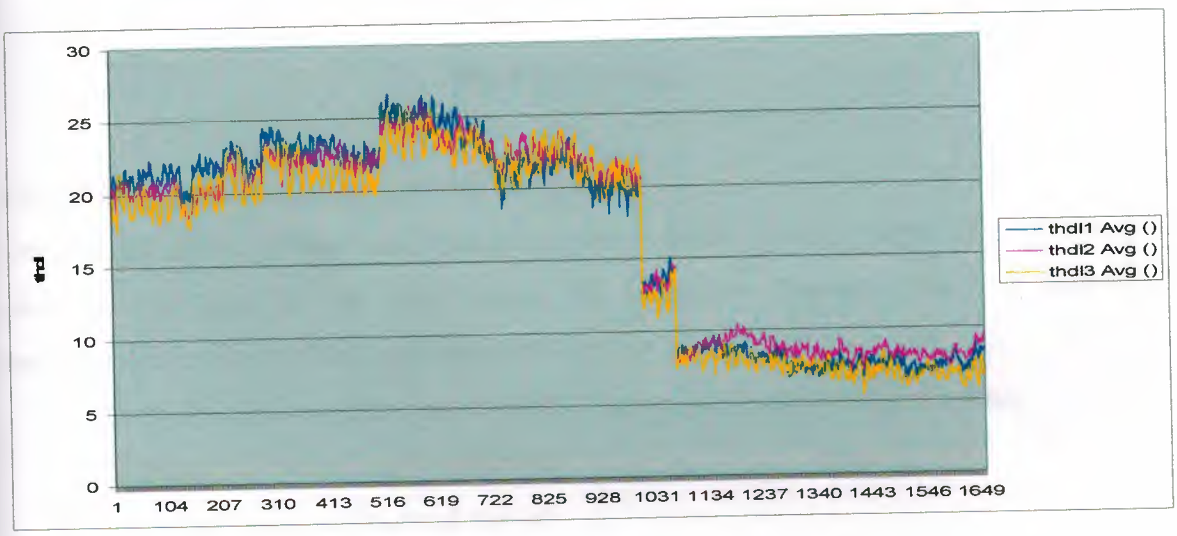


Fig 4.3.8 Total harmonic distortion of current

CHAPTER FIVE
COMPONENTS

5.1 Harmonic Filter Reactors

Fig 5.1.1 Reactor

Up to 36 kV 160 A Standard : EN 60289-IEEE 59

Detrimental effects of harmonic distortion can be stated in many ways. The following examples are some of the many causes for harmonic distortion that may lead to malfunctions in components and electrical systems :

- Excessive heating effect on electrical distribution equipment and cables
- Increased electrical insulation stress
- Increased noise in electrical motors
- Electronic mis-timings: Computers, fax machines etc.
- Capacitor overloads - premature failure
- Increased maintenance requirements and downtime
- Fluorescent light flickering
- Tripping of thermal circuit breakers unexpectedly.

Non-linear equipments generate harmonics as multiples of a fundamental frequency such as ac & dc motor speed controllers, arc furnaces, compact fluorescent lamps, uninterruptible power sources. As a result of this fact, besides the fundamental frequency, 3rd,5th,7th,9th harmonics are generated.

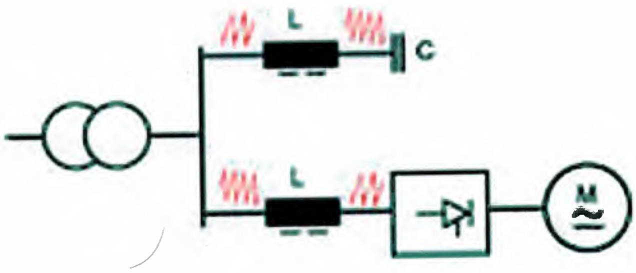


Fig 5.1.2 example for connection

1-36 Kv 160 A

Fr=134 – 189 – 210 Hz

Standard :EN 60289-IEEE 59

The distortion limits determined by IEEE19 standard are as follows :

Hospitals and Airports : % THD (V) = 3

Industrial Facilities : % THD (V) = 5

Speed Control Equipments : % THD (V) = 8

Power capacitors that are used in industrial plants in order to compensate reactive power can be used as filters besides being used for compensation. In order to prevent resonance in the systems, the value of inductance is selected approximately 20% different from the frequency of resonance which is

The most commonly used filters are 189 Hz and 210 Hz filters. In these frequencies, XL/ XC ratio which is called p factor is: %p = XL / XC = %7 and % 5.67 respectively. The harmonic values of six stroke loads are : I5 = 0.25I1, I7 = 0.13I1, I11 = 0.09I1.If an inductance is serial-connected to a capacitor (Qcn (kVAr)) in p value (XL/XC) at Ucn voltage, QC which is effective reactive power flows to the network that has Un voltage is:

$$Q_c = \frac{Q_{cn} \cdot U_n^2 \cdot (50/f_n)}{U_{cn}^2 (1-p/100)}$$

The inductance which resonates the Qcn capacitor at "fr"is:$L_n = f.U^2_{cn} / 2\pi f_r^2.Q_{cn}$

Example: In order to transfer Q_c = 50 kVAr to a U_n = 400V network for a serial - connected capacitor to a reactor (%p = 7), a U_{cn} = 440V capacitor must be Q_{cn} = 56.3 kVAr (or 525V, 80 kVAr). Thus, the terminal voltage of the capacitor is $U_c=U_n/(1\text{-}\%P)=400/(1-0.07)=430V$. Despite the fact that U_{cn} = 400V power apacitors are manufactured to resist 10% overloading and it is safe to use a 400 V capacitor for this example, $U_{cn} = 440$ V capacitors should be used in order to ensure complete safety. Using a bigger transformer than needed and using 12 stroke inverter instead of 6 stroke inverter reduces the harmonic distortion.

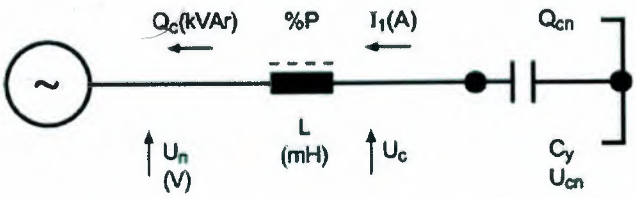


Fig 5.1.3 example for connection

Network voltage in addition to the harmonic current :

$$I_{rms} = I_{th} = \sqrt{I_1^2 + I_5^2 + I_7^2 + ...}$$

Power capacitors use 300-400 times more current than normal at the moment of switching. In order to reduce this to 100 times, a serial-connected reactor with a p=1% value can be used.

Specificatiions of harmonic filter reactors :

- Nominal Network Voltage $U_n = 400$ V
- Network Frequency $f_n = 50$ Hz
- Maximum harmonic current $I_h = 0.3\ I_1$
- Fundamental harmonic current $I_1 = ...$Ampere
- Heating limit current $I_{th} = 1.1\ I_1$ (p=7%) $1.2\ I_1$ (p=5.67%)
- Saturation current $I_{lin} = 1{,}8\ I_1$
- Maximum current $I_{max} = 2\ I_1$ (30 seconds)
- Insulation class B (130°C)
- Protection Degree IP00
- Insulation withstand voltage 3000 V_{ac} 1 minute
- Standards EN60289, IEEE59

Linearity of reactors

For harmonic and current limiting reactors.

For general purpose inductances.

Fig 5.1.4

Resonance Curves of Filters

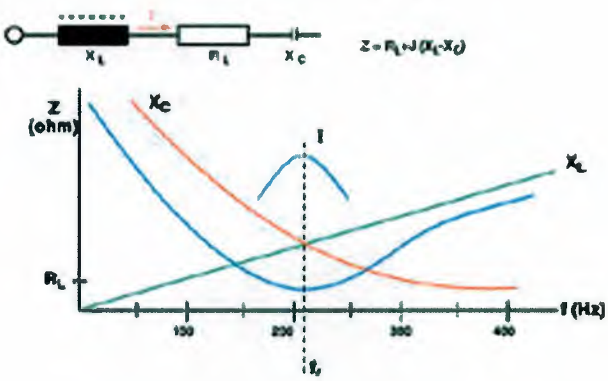


Fig 5.1.5 example the resonance curve of filters

An example of 400 V 150 kVAr filtered reactive power compensation

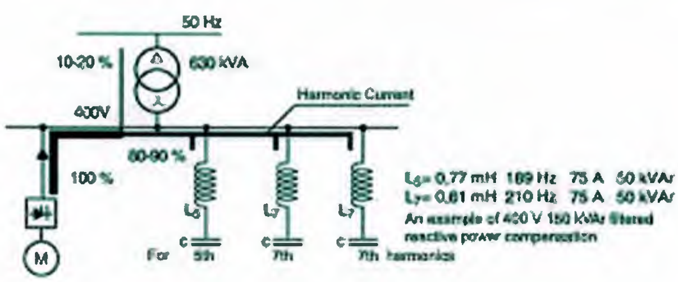


Fig 5.1.6 power compensation

Q_{cn}:Capacitor power at nominal voltage (kVAr)

C_y=3 (star equivalent capacity of delta connected capacitor)

Dimensions of 3 x 400 V Harmonic Filter Reactors

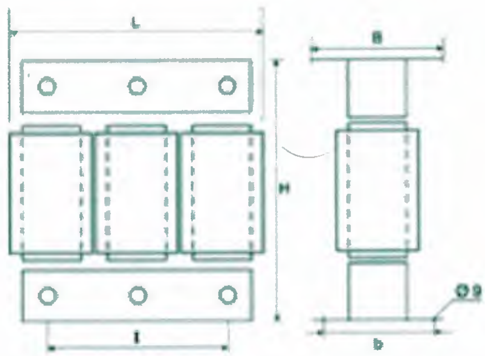


Fig 5.1.7

5.2 Neutral Earthing(Grounding) Resistors up to 36 kV

- Current Limiting Resistors for Introduction Furnaces
- Discharge Resistors for Capacitors
- Harmonic Filtering Resistors

Fig 5.2.1 Type test report

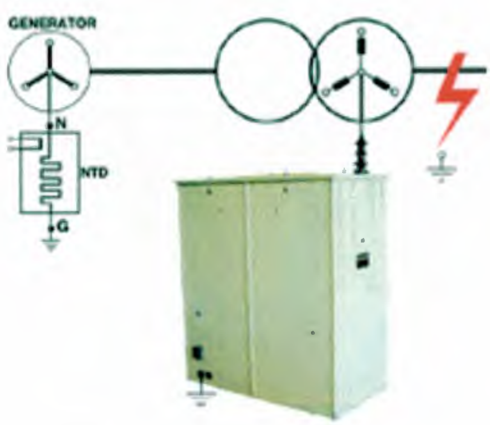


Fig 5.2.2 Example for earthing

Reducing single phase fault currents which occur in M.V. electrical networks to prevent damages on transformer and generators,decreasing temporary overvoltages occured by breaking instanteneous fault current,providing long - life for switching devices,decreasing the step voltages to a less harmfull level for personnel can be considered as the four most common reasons for using NERs.

Although it is possible to limit fault currents with high resitance NERs,high resistance earth short circuit currents can be extremely reduced.As a result of this,protection devices cannot sense the fault,For this reason,it is the most common application to limit single phase fault currents appox,20% of three phase short circuit currents.

Routıne tests :

1. Measurement of total dc resistance.
2. Over - voltage test to the network frequency for one minute

Technical Specifications	
Operating Voltage (kV)	36 / $\sqrt{3}$ and down
Rated Current (A)	Up to2000
Temperature Rise (°C)	< 600
Ambient Temperature (°C)	< 50
Resistance Alloy	Stainless - Stell
Protection Degree	IP23-IP54
Standards	TS595-TS3033-TS914-IEEE32

Type tests (on demand) :

1. Temperature rise test.
2. Definition of protection degree.
3. Impulse voltage test (1,2 / 50).
4. Measurement of ac resistance.

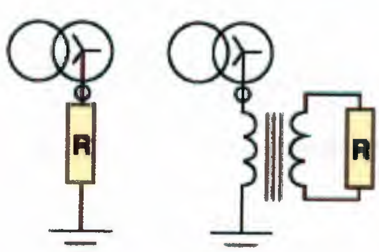


Fig 5.2.3 Examples of earthing the neutral point throuhg the resistance.

TYPE	Dimensions of sample ners				
	U_n(kV)	I_n(A)	t (s)	Weight (kg)	W x Lx H (cm)
NTD XL	$36/\sqrt{3}$	1000	5	800	100 x 200 x 200
NTD L	$36/\sqrt{3}$	300	5	350	100 x 100 x 200
NTD L	$17/\sqrt{3}$	1000	5	450	100 x 100 x 200
NTD M	$12/\sqrt{3}$	1000	5	320	100 x 100 x 150
NTD M	$12/\sqrt{3}$	300	5	150	100 x 100 x 150
NTD S	$12/\sqrt{3}$	200	10	100	100 x 100 x 80

R: NTD **I:** Protection Relay, **CT :** Current Transformer

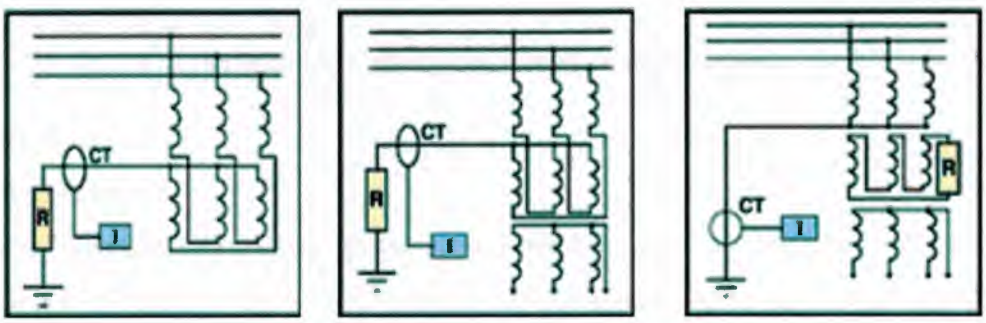


Fig 5.2.4 Examples for neutral point

Examples of providing artificial neutral point if is impossible to reach the neutral point of the source or if delta is connected.

5.3 Serial and Shunt Reactors

Air Core Harmonic Current Limiting Reactors.

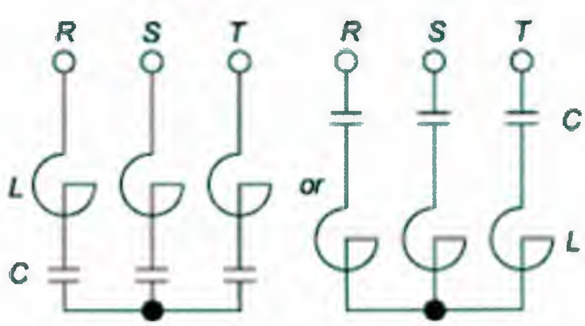


Fig 5.3.1 reactors

Technical Specifications :

- For outdoor and/or indoor
- Up to 36 kV from DC to 60 Hz
- Dry-insulated,air cooled
- Aluminium or copper winding
- Tolerance of rated inductance: For filtering reactors : (+ ,-) 3%
- For damping,series and natural earthing reactors : -0% +20% or due to specifications
- Class of temperature range B(130°C) or F(155°C)
- Ambient temperature range -40°C / +55°C
- Colling method : AN (air - naturel)
- Standards : EN60289 - IEEE59
- Color : Gray RAL7038
- Insulation of the winding : epoxy resin reinforced fiberglass

Applıcatıons :

Damping reactors :

Damping Reactors for power shunt capacitors.Are used in order to limit transient switching of a capacitor banks this reactors have low inductance values (appox. X_l=0.01 X_c) They are connected to power capacitors in series.

Harmonic filter reactor :

Harmonic filter reactors are designed to have low impedance at the required frequency.Harmonic currents flow into the LC instead of the network.

Current - limiting reactors :

The reactors are mainly used to limit short - circuit current to prevent fault current from rising to values dangerous for the devices.

Shunt reactors :

Shunt reactors,are used to compensate for capacitive reactive power produced by long lighty loaded transmision lines.Up to 36 kV and 18 MVAr.

Neutral grounding reactors :

Neutral grounding reactors,are used to limit short - circuit current between phase and earth.

5.4 True RMS – The Only True Measurement

Many commercial and industrial installations suffer from persistent so-called nuisance tripping of circuit breakers. Often these trips seem random and inexplicable but, of course, there is always a reason and there are two common causes. The first possible cause is inrush currents that occur when some loads, particularly personal computers and other electronic devices, are switched on; this issue is discussed in a later section of the Guide. The second likely cause is that the true current flowing in the circuit has been under measured – in other words, the current really is too high.

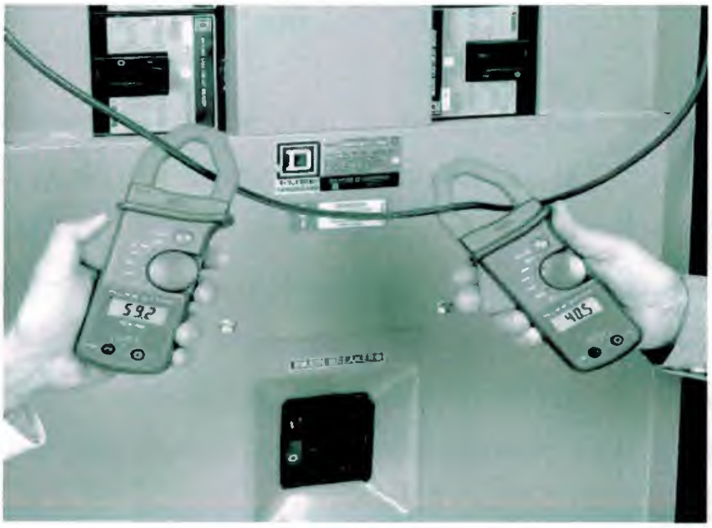


Figure 5.4.1 One current - two readings.

Under measurement occurs very frequently in modern installations – but why, when digital test instruments are so accurate and reliable? The answer is that many instruments are not suitable for measuring distorted currents - and most currents these days *are* distorted.

This distortion is due to harmonic currents drawn by non-linear loads especially electronic equipment such as personal computers, electronically ballasted fluorescent lamps and variable speed drives. Figure 5.4.3 shows the typical current waveform drawn by a personal computer. Obviously,this is not a sinewave and all the usual sinewave measurement tools and calculation techniques no longer work. This means that, when troubleshooting or analysing the performance of a power system, it is essential to use the correct tools for the job – tools that can deal with non-sinusoidal currents and voltages.

Figure 1 shows two clamp-meters on the same branch circuit. Both the instruments are functioning correctly and both are calibrated to the manufacturer's specification. The key difference is in the way the instruments measure.

The left-hand meter is a true RMS instrument and the right-hand one is an averaging reading RMS calibrated instrument. Appreciating the difference requires an understanding of what RMS really means.

Which do you trust? The branch circuit above feeds a non-linear load with distorted current. The True RMS clamp (left) reads correctly but the average responding clamp (right) reads low by 32%.

5.4.1 RMS

The 'Root Mean Square' magnitude of an alternating current is the value of equivalent direct current that would produce the same amount of heat in a fixed resistive load. The amount of heat produced in a resistor by an alternating current is proportional to the square of the current averaged over a full cycle of the waveform. In other words, the heat produced is proportional to the mean of the square, so the current value is proportional to the root of the mean of the square or RMS. (The polarity is irrelevant since the square is always positive.) For a perfect sinewave, such as that seen in Figure 2, the RMS value is 0.707 times the peak value (or the peak value is $\sqrt{2}$, or 1.414, times the RMS value). In other words the peak value of 1 amp RMS pure sinewave current will be 1.414 amps. If the magnitude of the waveform is simply averaged (inverting the negative half cycle), the mean

value is 0.636 times the peak, or 0.9 times the RMS value. There are two important ratios shown in Figure 5.4.2:

$$\textit{Peak factor} = \frac{\textit{Peak value}}{\textit{RMS value}} = 1.414 \quad \text{and} \quad \textit{Form factor} = \frac{\textit{RMS value}}{\textit{Mean value}} = 1.111$$

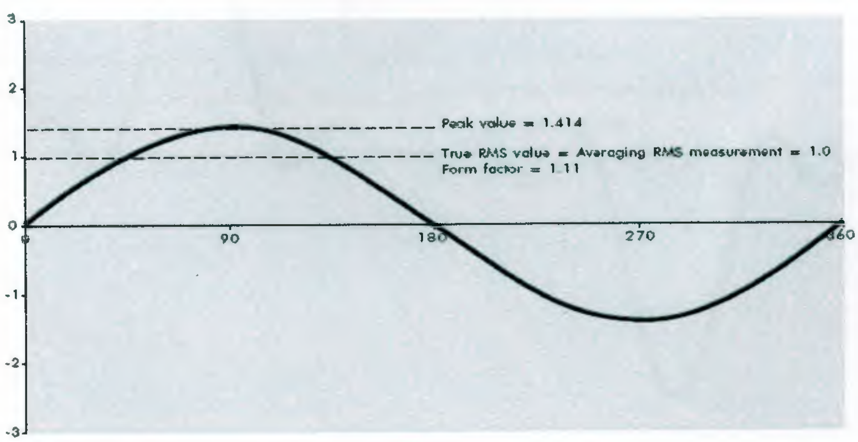


Fig 5.4.2 A pure somewave

When measuring a pure sinewave – but only for a pure sinewave - it is quite correct to make a simple measurement of the mean value (0.636 x peak) and multiply the result by the form factor, 1.111 (making 0.707 times peak) and call it the RMS value. This is the approach taken in all analogue meters (where the averaging is performed by the inertia and damping of the coil movement) and in all older and most current, digital multimeters. This technique is described as 'mean reading, RMS calibrated' measurement.

The problem is that the technique only works for pure sinewaves and pure sinewaves do not exist in the real world of an electrical installation. The waveform in Figure 3 is typical of the current waveform drawn by a personal computer.

The true RMS value is still 1 amp,but the peak value is much higher, at 2.6 amps, and the average value is much lower, at 0.55 amps. If this waveform is measured with a mean reading, RMS calibrated meter it would read 0.61 amps, rather than the true value of 1 amp, nearly 40 % too low. Table 1 gives some examples of the way the two different types of meters respond to different wave shapes.

A true RMS meter works by taking the square of the instantaneous value of the input current, averaging over time and then displaying the square root of this average. Perfectly implemented, this is absolutely accurate whatever the waveform. Implementation is, of course, never perfect and there are two limiting factors to be taken into account: frequency response and crest factor.

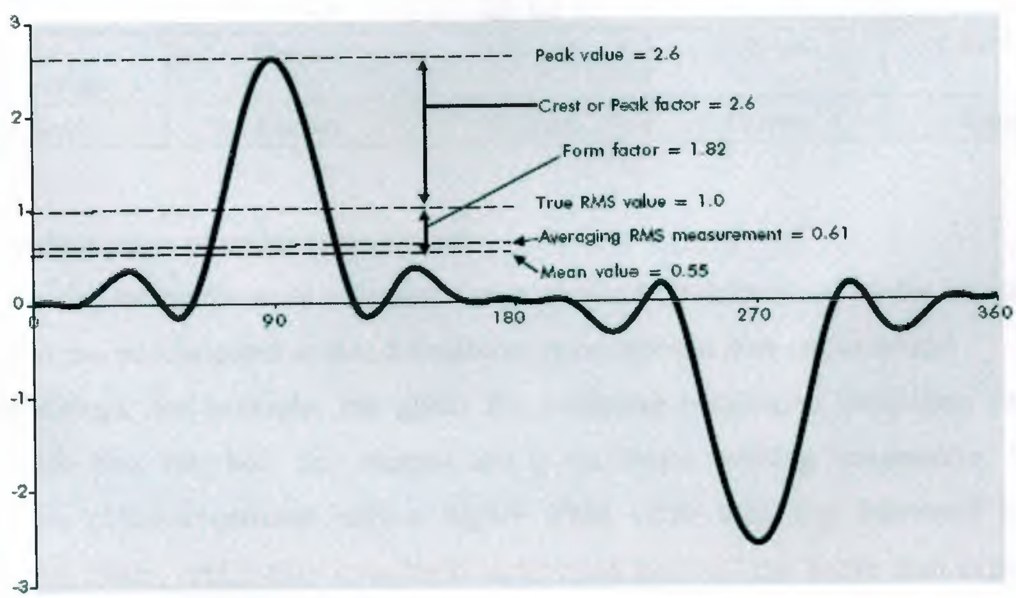


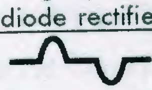
Figure 5.4.3 - Typical waveform of current drawn by a personal computer

For power systems work it is usually sufficient to measure up to the 50th harmonic, i.e up to a frequency of about 2500 Hz. The crest factor, the ratio between the peak value and the RMS value, is important; a higher crest factor requires a meter with a greater dynamic range and therefore higher precision in the conversion circuitry. A crest factor capability of at least three is required.

It is worth noting that, despite giving different readings when used to measure distorted waveforms, meters of both types would agree if used to measure a perfect sinewave. This is the condition under which they are calibrated, so each meter could be certified as calibrated – but only for use on sinewaves.

True RMS meters have been available for at least the past 30 years, but they used to be specialised and expensive instruments. Advances in electronics have now resulted in true RMS measurement capability being built into many handheld multimeters. Unfortunately, this feature is generally found only towards the top end of most manufacturers' ranges, but they are still cheap enough to buy as ordinary instruments for everyone and everyday.

Table 5.4.1 - A comparison of average responding and True RMS units

Multimeter type	Response to sinewave	Response to square wave	Response to single phase diode rectifier	Response to 3ø diode rectifier
				
Average responding	Correct	10% high	40% low	5-30% low
True RMS	Correct	Correct	Correct	Correct

The consequences of under measurement

The limiting rating for most electrical circuit elements is determined by the amount of heat that can be dissipated so that the element or component does not overheat.

Cable ratings, for example, are given for particular installation conditions (which determine how fast heat can escape) and a maximum working temperature. Since harmonic polluted currents have a higher RMS value than that measured by an averaging meter, cables may have been under-rated and will run hotter than expected; the result is degradation of the insulation, premature failure and the risk of fire.

Busbars are sized by calculating the rate of heat loss from the bars by convection and radiation and the rate of heat gain due to resistive losses. The temperature at which these rates are equal is the working temperature of the busbar, and it is designed so that the working temperature is low enough so that premature ageing of insulation and support materials does not result. As with cables, errors measuring the true RMS value will lead to higher running temperatures. Since busbars are usually physically large, skin effect is more apparent than for smaller conductors, leading to a further increase in temperature. Other electrical power system components such as fuses and the thermal elements of circuit breakers are rated in RMS current because their characteristics are related to heat dissipation. This is the root cause of nuisance tripping – the current is higher than expected so the circuit breaker is operating in an area where prolonged use will lead to tripping. The response of a breaker in this region is temperature sensitive and may appear to be unpredictable. As with any supply interruption, the cost of failure due to nuisance tripping can be high, causing loss of data in computer systems, disruption of process control systems, etc. Obviously, only true RMS instruments will give the correct measurements so that the ratings of cables,busbars and breakers can be determined properly. An important question is, 'Is this meter a true RMS meter?' Usually, if a meter is a true RMS meter it will say so quite prominently in the product specification but this is often not available when needed. A good idea of the answer can be obtained by comparing measurements with a known averaging meter (usually the cheapest to hand!) or a known true RMS meter while measuring the current in a non-linear load such as a PC and the current drawn by a filament lamp.Both meters should read the same current for the filament lamp load. If one instrument reads significantly (say more than 20 %) higher for the PC load than the other then it is probably a true RMS instrument, if the readings are similar, the meters are of the same type.

Conclusion

True RMS measurement is essential in any installation where there is a significant number of non-linear loads (PCs, electronic ballasts, compact fluorescent lamps, etc.). Averaging reading meters will give an under measurement of up to 40 % which can result in cables and circuit breakers being under rated with the risk of failure and nuisance tripping.

CONCLUSION

In ideal power system, current and voltage should be sinusoidal. Because of existence of non linear loads this is not possible. Distortion of sinusoidal voltage and current and harmonics perform as a result of non linear loads on the power system. This harmonics cause some problems on the system like overload, over heating and resonance. The most important method to prevent the occurrence of harmonics are precaution during design process and using harmonic filters. Precaution during design depends on designing a system structure that does not produces harmonics or at least produces a tolerable amount. The other method depends on detecting the harmonic load and level of each harmonics in the system assign suitable filters accordingly to be mounted to the system.

The system parameters and need assign the choice of suitable filter and mount them to the system. Later the parameters of the system are measured again and see whether the filter is suitable for the system or not. However this method may not always provided a suitable solution. For this reason depending and the measured system parameters, a simulation of the system is carried out and the system reaction to filter is recorded according to the simulation. The system simulation provides observation of possible problem that may occur and provides also for precaution prevent this problems.

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