

THE OUTPUT FREQUENCY SPECTRUM OF A
THYRISTOR PHASE-CONTROLLED CYCLOCONVERTER
USING DIGITAL CONTROL TECHNIQUES.

by

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A thesis submitted to the C.N.A.A.
in partial fulfilment for the award
of the degree of Doctor of Philosophy.

Sponsoring Establishment:

Plymouth Polytechnic.

Collaborating Establishment:

Bristol University.

August 1985.

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DECLARATION

I hereby declare that I am not registered for another degree. The following thesis is the result of my own investigation and composed by myself. It has not been submitted in full or in parts for the award of any other C.N.A.A. or University degree.

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Date : 31-7-85.

ABSTRACT

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Various methods of analysing the output waveform have been considered. A Fortran 77 program employing a modified Fourier series, making use of the fact that the input waveforms are sinusoidal, was used to compute the individual harmonic amplitudes. A six pulse three phase to single phase cycloconverter was built and a Z-80 microprocessor was used for the control of firing angles. Phase locked loops were used for timing, and their effect upon the output with changing input frequency and voltage were established. The experimental waveforms are analysed by a FFT spectrum analyser.

The flexibility of the control circuit enables the following investigations not easily carry out using traditional analog control circuit. The phase relationship between the cosine timing and reference wave in the cosinusoidal control method was shown to affect the output waveform and hence the harmonic content. There is no clear optimum value of phase and the T.H.D. up to 500Hz remains virtually constant. However, the changes of individual harmonic amplitudes is quite significant. In practice it may not be possible to keep the value of phase constant but it should be considered when comparing control strategies.

Another investigation involves the changing of the last firing angle in a half cycle. It shows that the value of firing angles produced by the cosinusoidal control method is desirable. Operation at theoretical maximum output frequency was also demonstrated.

ADVANCED STUDIES

During the research program I undertook a course of advanced studies. These included the reading of selected articles associated with the field of study. I attended several conferences and a colloquia on power electronics as well as the 1982 IEE Vocation School on Power Electronics. I also became an Associate Member of the IEE.

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List of symbols.

D, N, n, m integers

A_0	}	Fourier coefficients
A_n		
B_n		
C_n		

F_f repetition frequency

F_i input frequency

F_o output frequency

P pulse number

r modulation depth

t time

T period

V_{max} maximum instantaneous voltage

V_{mean} mean load voltage

V_{out} output voltage of converter

ω angular frequency

α firing angle

θ, ϕ phase angle

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CHAPTER ONE
INTRODUCTION.

1. Introduction.

The development of high power semiconductor devices in the past three decades opens up new areas of application for these devices. The more successful devices are thyristors, triacs, bipolar and recently field-effect transistors. The areas of application cover both domestic and industrial environments. Light dimmer, switching mode power supply, d.c. and a.c. motor controller, high voltage d.c. transmission link are just some examples. The functions they perform are voltage control and frequency conversion. (d.c. being considered as zero Hz). An application that usually requires a large number of thyristors is the cycloconverter.

A cycloconverter is a frequency changer. It converts directly a.c. of one frequency to a.c. of another frequency in a single stage. The power rating of a cycloconverter is usually high i.e. up to and above hundreds of kilowatts. The output voltage waveform of a cycloconverter consists of different sections of the usually three phase input waves as shown in Fig. 1.1. In a modern cycloconverter the switching is done by a group of thyristors connected between the input and output. As cycloconverter loads are designed to work with sinusoidal waveforms e.g. induction motors, the distortion in the output of a cycloconverter can cause extra losses in the load and may also affect the stability of the drive system. This prevents more widespread application of cycloconverters.

This dissertation addresses the application of a microprocessor in the control circuitry of a cycloconverter and the flexibility of control thus offered. Using such a control circuit, it is shown that the traditional cosinusoidal control method does not uniquely define an output waveform for a certain output frequency and voltage. The phase relationship between the cosine timing and reference waveform also affects the output waveform and hence its harmonic content. This dissertation also includes the harmonic content at high output frequencies and effects of changing firing angles produced by the cosinusoidal control method.

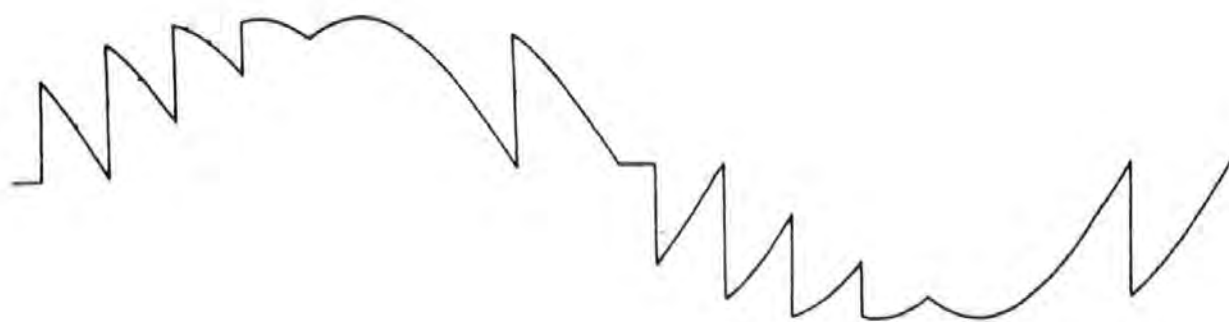


Fig. 1.1 Output voltage waveform of a cycloconverter.

1.1 Applications of cycloconverters and alternative methods.

The first application of the cycloconverter was for traction in Europe in the 1930's. A three phase 50 Hz supply was converted to single phase 16.66 Hz using grid controlled mercury arc rectifiers. To date, the applications of cycloconverters can be divided into two groups :

(A) Those which convert the mains supply of fixed frequency to a range of different, usually lower, frequencies for variable speed drives. For example, cycloconverters supplying synchronous motors were used for mill drives (1),(2),(3) and supplying induction motors for mining applications (4).

(B) Those which convert the variable frequency output of a generator driven by a variable speed prime mover to a fixed frequency such as some 400 Hz supplies on aircraft.
(5) (P.431) (6) (P.10-14)

The cycloconverter is, of course, not the only solution to the above areas of applications. For the variable speed constant frequency (V.S.C.F.) application, the alternative can be a hydraulic speed regulation system to drive the generator at constant speed and hence obtain constant frequency output. Hydraulic speed control system tends to be heavy and have a relative low efficiency.

Cycloconverters have the advantage of small size and light weight, but usually both output voltage and current waveform contains harmonics.

For variable speed drives, the situation is more complicated. The subject is very lively at present and there is a large number of publications. An annual conference is devoted to Drives /Motors /Controls (7), (8) and (9). This topic is also covered by conferences on electrical machines (10) and power electronics (11). A full review of the subject is published in recent IEE Proceedings (12).

In the past the Ward-Leonard set has been widely used for variable speed drives. Then the thyristor phase controlled converter as a direct substitute of the Ward-Leonard system became popular. Variable frequency and voltage for A.C. motor speed control can be provided by phase control rectifier followed by a force commutated thyristor quasi square wave inverter. Recently there is a lot of interest shown on pulse width modulated inverters. Some of them use newly developed devices such as single or darlington transistors, gate-turn-off thyristors (GTO) and field effect transistors (FET).

The disadvantages of d.c. motor systems are associated with the commutator of the d.c. motor. Apart from the absolute speed limit imposed by the mechanical

construction of the commutator, at high power level commutation problems also dictate operation at low speed. The first cost of a d.c. motor is also higher than a squirrel cage induction motor with equal ratings and enclosure, and requires more frequent maintenance.

The attraction of A.C. motors, specially squirrel cage induction motors, are that they are cheap, robust and easily available. For induction motors, a small reduction of speed can be obtained by regulating the supply voltage or by increasing the rotor resistance. Eddy current coupling can be used if efficiency is not an important consideration e.g. if the system only runs at low speed for a short length of time. Inverters and cycloconverters with wide output frequency range, e.g. 2 to 60 Hz and 1 to 20 Hz respectively, are used for a.c. drives requiring a wide speed range (13). These converters require more complex control functions than phase controlled converters used in d.c. drives. The cycloconverter tends to occupy the high power end of the market due to the large number of thyristors required and its potentially low losses because conversion is done in a single stage.

The co-existence of so many systems is not surprising if we consider the number of units used by each system. A unit is defined here as an electrical machine or power electronic circuit that can handle the full load

power. This is not to suggest that one unit is equivalent to another in terms of performance, size and cost etc. This simple comparison is a guide to the requirements of different systems in comparison with a cycloconverter. Table 1 shows that the total number of units used from mains input to mechanical output is three except for a cycloconverter. However a cycloconverter generally requires more thyristors in its power electronic circuit than other systems. (See next section for the cycloconverter circuit). The characteristics of the different units are reflected in the choice of system for an application. For example the price fall of thyristors in terms of performance/cost in the last decade tends to favour phase controlled converters more than Ward-Leonard sets. But until there is a significant improvement in one of these units, there is no overall winner.

Table 1. Comparison of the number of units used by
variable speed drive systems
(from electrical input to mechanical output)

Types of Drives	Electrical Machine(s)	Power Electronic	Extra for four quadrant operation	Total
Ward-Leonard	3	0	0	3
Phase controlled converter	1	1	1	3
Inverter	1	2 *	0	3
Cyclo-converter	1	1	0	2 **

* For systems supplied by batteries, a charger is usually required.

** Cycloconverters require more thyristors than power electronic units used in other systems.

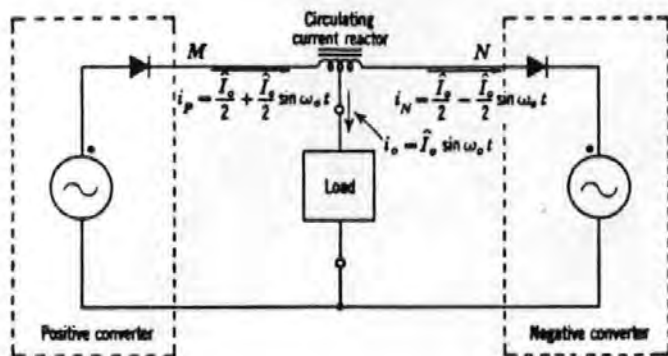
1.2 Cycloconverter circuits.

The operation of Naturally Commutated Cycloconverters (N.C.C.) can be described as modulating the output of a phase controlled a.c. to d.c. converter to produce an a.c. output. As a phase controlled converter can supply current in one direction only, two back-to-back converters are required for a balanced single phase a.c. output.

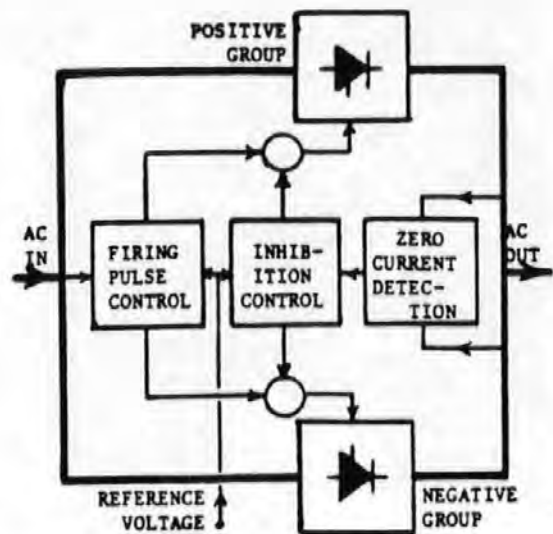
As the two converters produce current in opposite directions, it is necessary to control the magnitude of current flowing from one converter to the other. There are two methods to prevent a large current flowing. Fig. 1.2 show the two methods in block diagram form.

(A) A large circulating current reactor is connected between the converters to limit the maximum current flowing - this is the circulating current cycloconverter. The reactor needs to carry the full load current as well as the circulating current flowing between the converters. Therefore the inductor is usually large and expensive.

(B) Only one converter is allowed to conduct at any time by monitoring the current in the conducting converter and wait until the current in the conducting converter reaches zero before supplying firing pulses to the other group and allowing it to conduct - this is the inhibit cycloconverter



(a)



(b)

Fig 1.2 Block diagram of cycloconverter

(a) circulating current cycloconverter

(b) inhibit cycloconverter

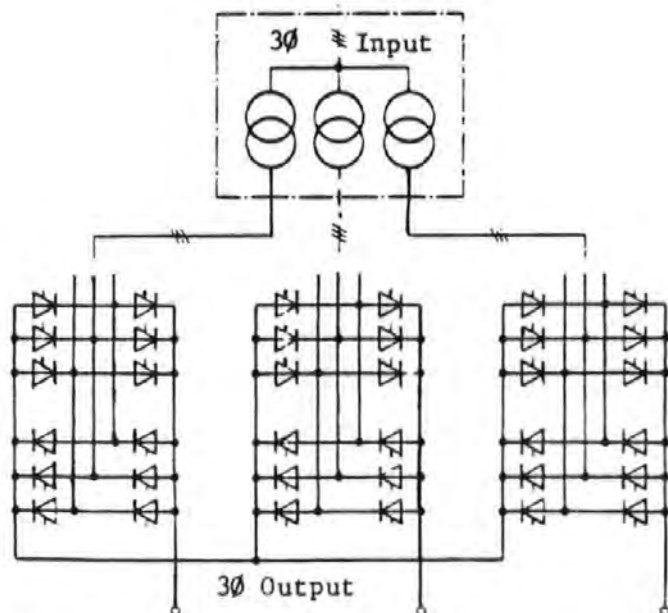


Fig 1.3 Connection of thyristors
in a six pulse cycloconverter
(Three phase input
and output)

(14). Extra sensing and control functions are required in the control circuit to perform the function described above for this type of cycloconverter.

Similar to the phase controlled converter, cycloconverters can have either single or multiple (usually three) phase input, and single or multiple phase output. Also, a better output waveform can be obtained by increasing the pulse number. However, more thyristors are required for cycloconverter with a higher pulse number. A typical three phase input and output, six pulse cycloconverter consists of 36 thyristors, see Fig. 1.3. Note that for bridge circuits with multiple phase output, isolation between different output phases is required. It can be in the form of isolated phases in the input transformer as indicated in Fig.1.3 or employing a motor with six connections.

As the cycloconverter employs natural commutation, the switching of current from one thyristor to the next relies on the incoming thyristor having a higher supply voltage at the switching instant and thus reverse biases the conducting thyristor and turns it off. A gate pulse applied to the incoming thyristor, at the instant when that thyristor is due to start conduction is the only control signal required. The pulse should be of sufficient length so as to allow current in the usually inductive load to build up to above the latching current of the thyristor. Apart

from the voltage and current handling capability, there are no other special requirements for the thyristors used, and converter grade thyristors are sufficient. Hence there is no essential difference in the power part of the circuit between a cycloconverter and a dual phase controlled converter for reversible d.c. drive.

There are other, not so common cycloconverter circuits. For example, if the three phase load is connected in delta, only two separate single phase outputs are required. This is the open-delta cycloconverter (15) (P.216-220). Recently, cycloconverters have been used to produce outputs that have frequencies higher than the input (16). Natural commutation is possible because a high frequency resonant (tank) circuit is connected to the output. High frequency is desirable because small storage elements, i.e. the capacitor and inductor, can be used. With two cycloconverters connected together, with a high frequency link, the system can convert input power of any frequency and voltage to output power of any other frequency and voltage (17), (18). The term generalised transformer is sometimes used to describe such a power conversion system.

1.3 Cosinusoidal control method.

Control of cycloconverters is achieved by modulating firing angles. Therefore to simulate a sine wave the thyristors are, at first, triggered to produce a low voltage. Then the firing angle is advanced to increase the output voltage and is again retarded at the end of the half cycle. The firing angle is thus, in general, changing from one pulse to the next.

For a P pulse phase controlled converter with continuous current, the average output voltage over a period between firing of thyristors with constant firing angle is given by Lander (19) (P.77).

$$V_{\text{mean}} = V_{\text{max}} * \frac{P}{\pi} * \text{SIN}\left(\frac{\pi}{P}\right) * \text{COS}(\alpha)$$

where P is the pulse number of the circuit

α is the firing angle $0 \leq \alpha \leq \pi$

Hence with $\alpha = 0$ the converter produces maximum output voltage and by definition is the earliest possible moment for natural commutation. With the desirable output voltage for cycloconverter in the form of $V_o * \text{SIN}(\theta)$, if the converter is to produce this voltage

$$\text{Then } V_o * \text{SIN}(\theta) = K * V_{\text{max}} * \text{COS}(\alpha)$$

$$\Rightarrow \alpha = \cos^{-1}(r * \sin(\theta)).$$

To approximate the desired sinusoidal output for the cycloconverter, the cosine of the firing angle (α) must vary sinusoidally. For maximum output voltage i.e. $r=1$, the firing angle is actually increasing and decreasing linearly (20), (21). As the firing angle is changing, the average output voltage given above is not strictly applicable to cycloconverter but the output as shown in Fig. 1.4, roughly follows the desired sinusoidal waveform. To search for better output waveforms the firing angles produced by the cosinusoidal control method is varied as described latter.

Traditionally with analog control circuits, a series of cosine timing waves are derived from the supply (see Fig. 1.4 and 1.5). For the three pulse example shown, these are just the input phases inverted and scaled down to appropriate scale. For the six pulse case, the cosine timing waves are the sum of two input waves and their inversion. These cosine timing waves are compared with the output of a free running oscillator - the reference wave to produce the firing instants. At the crossing point of the two waves, the instantaneous value of the cosine timing wave equals the instantaneous value of the reference wave and firing the next thyristor at that instant would produce an output voltage proportional to the reference wave. The rate of

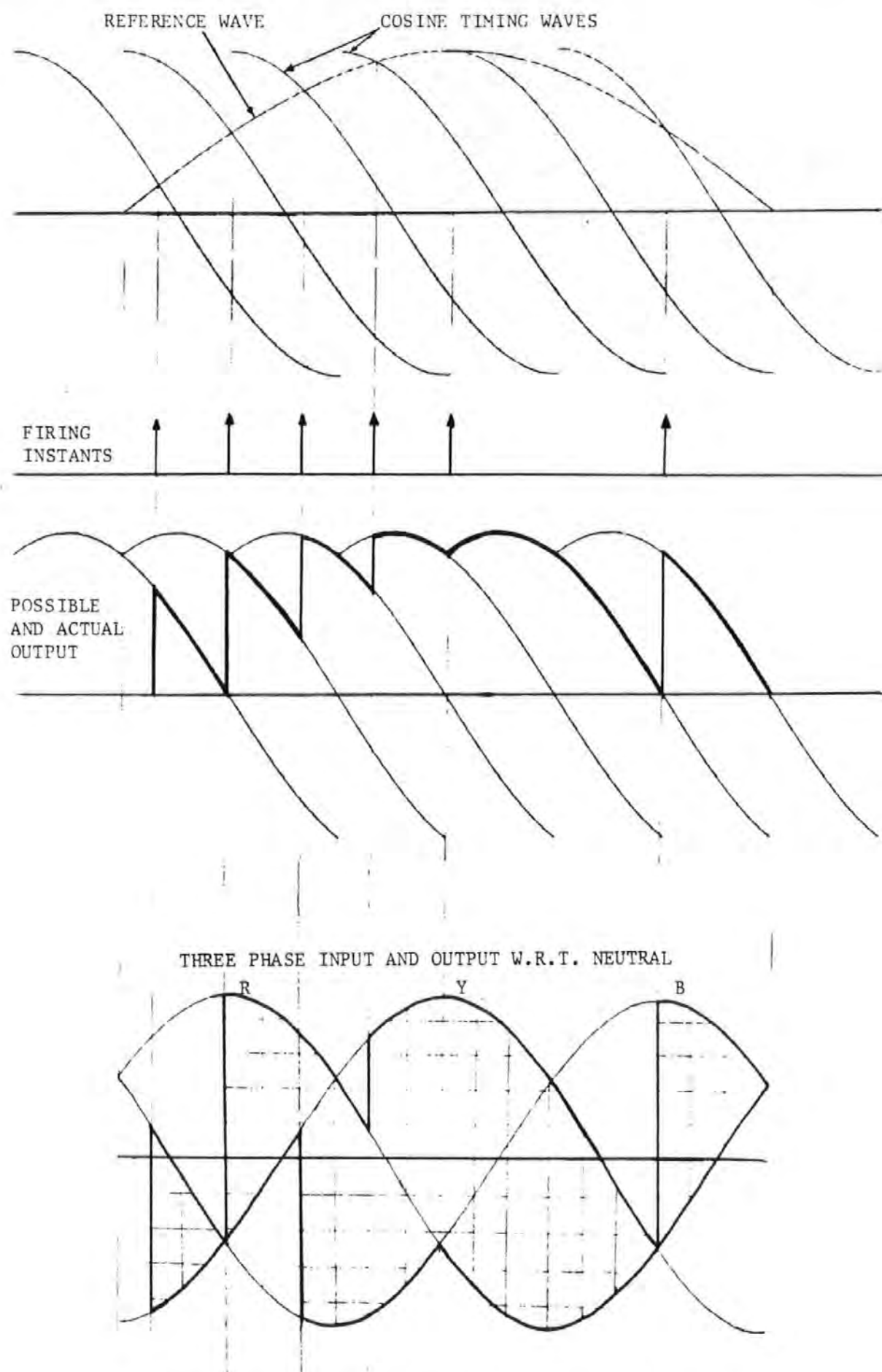


Fig. 1.4 Cosinusoidal control method for a cycloconverter.
(6 pulse)

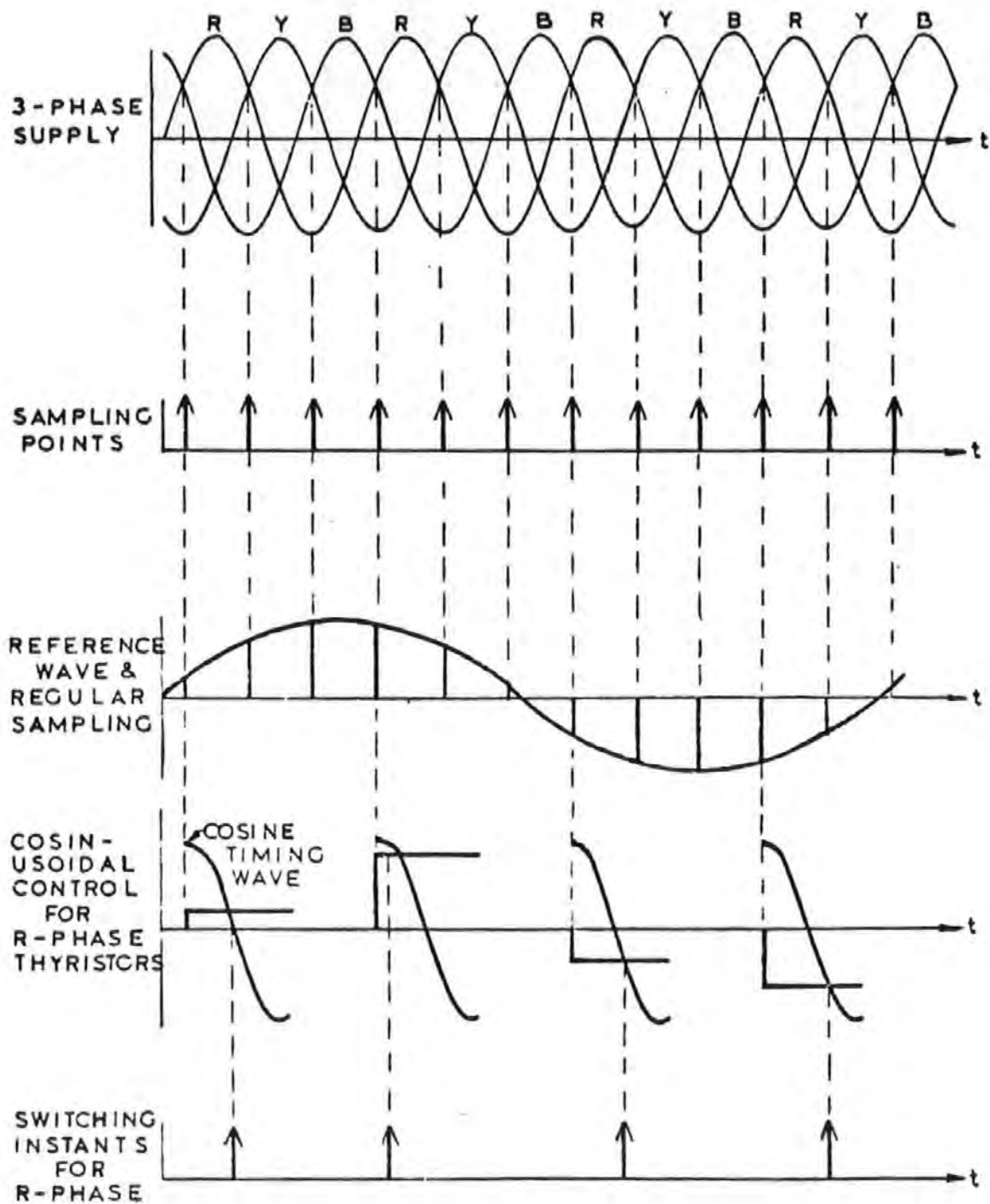


Fig. 1.5 Cosinusoidal control method for a cycloconverter with regular sampling

change of the reference wave determines the rate of change of output and therefore its frequency. The relative amplitude of the reference wave determines the magnitude of the output voltage and is defined as unity when they are of equal amplitude.

With the cosinusoidal control method, the firing angles depend on the value of the two waves as they cross each other. The values at other parts of the wave do not affect the result. Also the crossings occur at irregular intervals and the method is called the Natural Sampling Method. A modification of the method that attempts to produce the desired output at more regular intervals, at the start of the cosine timing wave, is called Regular Sampling Method (22) as shown in Fig.1.5. The output is supposed to synthesis the reference wave at regular intervals, i.e. at the start of the cosine timing waves, but the firing instants are, of course, still at irregular intervals to produce an A.C. output.

The actual number of cosine timing waveforms in a input cycle equals the pulse number of the converter circuit. For a 50 Hz supply, 3 pulse circuit, the spacing between the cosine timing waves is 6.6 ms or 120 degrees. The reference wave is the controllable signal and can change in frequency and amplitude independently and hence vary the output voltage and frequency of the cycloconverter. In fact the

reference wave need not be sinusoidal and the cycloconverter can be considered as a 'linear' amplifier of the reference wave if cosine timing waves are used. A trapezoidal reference waveform is sometimes used to improve the input power factor and increase the magnitude of the wanted component. Timing using other waveforms e.g. triangular was considered by Bland (23) and shows that the use of cosinusoidal control signals is desirable for sinusoidal outputs.

1.4 Power factor of cycloconverter.

Cycloconverters can handle both capacitive and inductive loads. In fact the phase angle between the fundamental component of voltage and current can be greater than 90 degrees. With displacement angles greater than 90 degrees or power factor negative, the power flow is reversed. Therefore one cycloconverter at each end is sufficient for a high frequency bi-directional power transfer link. In drive applications, this results in regenerative braking.

The input power factor of a cycloconverter is, however, always lagging. This is due to the fact that at most parts of the output cycle, the firing angle is greater than zero to produce a less than maximum output voltage and consumes current in the latter part of the input cycle. The

input current is not sinusoidal but only the fundamental of the input current produces real power. The maximum input power factor, shown by Pelly (15)(P.359), is 0.843 and occurs at maximum output voltage with unity load displacement factor. This could be an important consideration as more and more thyristor circuits are used and could lead to restriction in their use or higher tariff on thyristor equipments. This is an even more important consideration if supply comes from a weak link or a dedicated generator as is almost certainly the case for VSCF applications.

CHAPTER TWO

ANALYSIS OF CYCLOCONVERTER OUTPUT WAVEFORM

2. Analysis of cycloconverter output waveform.

As the output voltage waveform of a cycloconverter is not a pure sine wave but made up of sections of sine waves at supply frequency, assessment of the distortion is important. It helps to evaluate the merit of different control strategies and to enable comparison to be made with other converter systems. It also helps to deduce the acceptability of such a waveform to the load and obtain the desirable filter characteristic if it is required.

2.1 Fourier component method.

The usual method of analysis is to transform the output waveform into the frequency domain using Fourier series. The waveform to be analysed must be periodical. If the waveform to be transformed is $F(t)$ with period T . Then

$$F(t) = A_0 + \sum_{n=1}^{\infty} A_n \cos(n\omega t) + \sum_{n=1}^{\infty} B_n \sin(n\omega t)$$

$$\text{where } A_0 = 1/T \int_0^T F(t) dt$$

$$A_n = 2/T \int_0^T F(t) \cos(n\omega t) dt$$

$$B_n = 2/T \int_0^T F(t) \sin(n\omega t) dt$$

$$\omega = 2\pi/T$$

n is a positive integer.

alternatively

$$F(t) = \sum_{n=0}^{\infty} C_n \cos(n\omega t + \phi_n)$$

$$\text{where } C_n = \sqrt{A_n^2 + B_n^2}$$

$$\phi_n = \tan^{-1}(B_n/A_n)$$

This is an infinite but convergent series of sinusoidal functions. The frequencies of these are multiples of the repetition frequency of the waveform (plus a d.c. term if the waveform is not balanced). A method for determining the fundamental repetition frequency (F_f) of a cycloconverter output waveform was described by Hanley (24). This involves calculating the quantity $P \cdot F_i / F_o$ (In most cases this is the number of sections of input wave in a output cycle, see section 2.1.2), and reducing the numerator and denominator to the smallest integers. If D is the smallest integer denominator, then it requires D cycles of the wanted output before the waveform repeats itself. Therefore the fundamental repetition frequency is $1/D$ times the wanted output frequency i.e. F_o/D . The wanted output component becomes a harmonic of the repetition frequency. However, the amplitude of the fundamental could be very small compared with the wanted output component.

2.1.1 Graphical Methods.

The harmonic content of some waveforms, e.g. the magnetising current of a transformer, can be obtained by graphical methods. To obtain the third harmonic component, three output waveforms, displaced by one third of the period of the complete waveform are added together and then divided by three. Higher order harmonics can be evaluated using numerical integration of the Fourier series. These methods can be useful for evaluating a few of the harmonics of experimental waveforms.

Graphical methods are not very useful for analysing cycloconverter waveforms because the amplitudes of the low order harmonics, e.g. third and fifth harmonics in the working range, are usually low. The dominant harmonics are centred around the average switching frequency $P \cdot f_i$.

2.1.2 Modifications of Fourier Series.

For a cycloconverter with ideal source and semiconductor switch, i.e. assuming no commutation overlap and conduction losses, one phase of the output voltage of a cycloconverter can be represented by:

$$F(t) = \left\{ \begin{array}{ll} \text{SIN}(wt + \phi_1) & T_0 < t < T_1, T_p < t < T_{p+1}, \dots T_{(m-1)p} < t < T_{(m-1)p+1} \\ \text{SIN}(wt + \phi_2) & T_1 < t < T_2, \quad \quad \quad \cdot \quad \quad \quad \cdot \\ \cdot & \cdot \quad \quad \quad \cdot \quad \quad \quad \cdot \\ \cdot & \cdot \quad \quad \quad \cdot \quad \quad \quad \cdot \\ \text{SIN}(wt + \phi_p) & T_{p-1} < t < T_p, T_{2p-1} < t < T_{2p}, \dots T_{mp-1} < t < T_{mp} \\ F(t + T_{mp}) \end{array} \right.$$

where T_{mp} is the period of the waveform.

This equation shows that the sinusoidal input waveforms are connected to the output in turn. Therefore the output can also be described as a phase modulated sine wave. The step change of phase is equal to $2\pi/P$ but at irregular intervals.

There is usually a large number of sections, e.g. 30 for a six pulse cycloconverter with 10 Hz output. Therefore analysis by the normal Fourier series is very laborious. A large number of terms need to be integrated and the limits substituted to obtain the Fourier coefficients. Using the fact that the input waveforms are sinusoidal, Hanley (25) reduced the equations of integration to algebraic summation. The product of the two sinusoidal functions to be integrated is rearranged to become the sum of two sinusoidal functions. The integration is then performed on the two terms, leaving the times of switchings to be substituted. The derivation is shown in Appendix I.

The calculation of Fourier components using the modified method is ideally suited for a digital computer. A Fortran 77 program was written to perform the calculations. Section 4.5 will give a more detail description of the program.

Another modification of the Fourier series, in complex form, was published by Slonim (26). By performing integration by parts twice on the product of the two sinusoidal functions, the integration is once again reduced to summation.

2.1.3 The Fast Fourier Transform.

The Fast Fourier Transform (F.F.T.) is a numerical analysis technique that also eliminates the need to perform integrations. A periodic waveform is sampled at 2^N points and transformed into (2^N-1) frequency components. The algorithm for the transform itself is very efficient compared with a numerical integration implementation of the Fourier series. But for cycloconverter output waveforms not much computation time may be saved by using this method as all 2^N points of the output need to be evaluated. There may be a thousand points to give an accurate representation of the waveform, compared with the evaluation of just the switching instants in the last method.

2.1.4 Existence functions.

Existence functions (5) are mathematical descriptions of the operation of switching power converters. An existence function has a numerical value of one or zero to describe the two states of a mechanical or semiconductor switch (all types - thyristor, transistor etc.). Operation of the whole converter is described by a set of existence functions defining the actions of all switches. The usefulness of existence functions in converter analysis is that the output of a converter can be considered as the sum of the outputs produced by the individual switches - hence described as the total sum of the existence functions multiplied by the input waveforms. To obtain the frequency component, the input waveforms and existence functions are transformed separately into the frequency domain and then multiplied together to obtain the contributions from every input phase. These are added together to obtain the spectrum of the complete waveform. The procedure is efficient because usually the input can be considered as steady d.c. or pure sine wave a.c. These can be represented by a single term in the frequency domain. The procedure is efficient also because the existence function having value of either one or zero is relatively easy to transform into frequency domain. This eliminates the need of integrating complicated terms.

The particular advantage of this approach for

frequency analysis is that it can be applied to all switching converters and it gives a universal approach to the subject - that is, the function and behaviour of switching converters are controlled by the switches and depends on the switching pattern or existence function.

2.1.5 Spectrum analyser.

Practical measurement of harmonic distortion involves the use of a spectrum analyser. The spectrum analyser resolves the input waveform into its harmonic components and displays amplitude of harmonic against frequency rather than the usual display of amplitude against time on an oscilloscope. During the evaluation of the harmonic amplitudes, only a finite number of output cycles can be considered by the spectrum analyser. There are two types of spectrum analyser. One type employs a tunable analog filter, and the display is built up point by point along the frequency scale. Therefore if the waveform is not repeatable, the harmonics are strictly from different waveforms. Also it takes a long time (in the order of tens of minutes) to produce the spectrum of a waveform at mains frequency. As it requires a very stable but turnable reference frequency, a warm up period after switch on is also required to eliminate thermal drift before accurate results can be taken.

The Fast Fourier Transform type analyser digitizes the input waveform and performs the transformation digitally. Only one set of successive samples of the input is taken and is used to calculate the amplitudes of all the harmonics. The process can be separated into the sampling of the waveform and the transform calculation. The sampling process is exactly the same as that in a data logger or digital storage oscilloscope. Indeed, a recent trend is to link a digital storage oscilloscope, often with a IEEE-488 or GPIB bus, to a micro-computer and to perform the calculations in the computer. Thus the frequency analysis is performed by two low cost general purpose equipments. As the sampling frequency in the F.F.T. type analyser can be kept constant easily at low frequency, this eliminates the problem of frequency drift in the analog filter type analyser. The experimental results were obtained on a H.P. 3582A F.F.T. spectrum analyser.

Phase information is generally not available on an analog filter type analyser. The F.F.T. type analyser usually uses the trigger point as reference. Therefore, if the phase between reference wave and fundamental output component is required, the analyser must be triggered by the reference wave. (The reference wave is not always available as it could be just numbers stored in the computer's memory.) Also the phase of all harmonics depends on the triggering point.

2.2 Assessment of Harmonic Content.

The production of harmonics in the output of a cycloconverter as a result of the inherent process of synthesis is unavoidable. The output of a cycloconverter will always contain some harmonics. Strictly, due to the modulation of a fixed input frequency by a changing wanted output frequency, the frequencies of distortion components are generally not integer multiples of the wanted output frequency, but the term harmonic is used to refer to frequencies other than the wanted frequency. It has been shown by Pelly (15) (P.397) that the frequency of the distortion components of a three pulse cycloconverter with cosinusoidal control can be expressed as

$$3(2P-1)*F_i + 2nF_o \quad \text{and} \quad P*F_i + (2n+1)*F_o.$$

This shows that with output frequencies much lower than the input frequency the harmonics are in families centred around frequencies at integer multiples of the average switching frequency $P*F_i$. It also shows that some of the distortion components can have frequencies lower than the wanted output frequency. For example the $(3F_i - 4F_o)$ term, with input frequency of 50 Hz and output frequency of 35 Hz, has a frequency of 10 Hz. In general, all components with frequencies lower than the wanted output frequency are

termed sub-harmonic. The condition for having sub-harmonic components is related to the repetition frequency which is discussed in section 2.1. There is no sub-harmonic at output frequency of $P \cdot F_i / n$ (where n is a integer) - i.e. for a three pulse cycloconverter, the output frequency is 150, 75, 50, 37.5, 30, 25 etc Hz. because the repetition frequency is equal to the wanted output frequency. It should be made clear that the use of Fourier series can be still completely valid even for^{an} output with sub-harmonics because the fundamental frequency of the Fourier series is set to the repetition frequency rather than the wanted output frequency.

Another implication of the general harmonic frequency expressions is that certain frequencies can have contributions from more than one family. For example, with input frequency of 50 Hz and output frequency of 30 Hz, the $(3F_i - 4F_o)$ term has a frequency equal to 30 Hz i.e. the wanted output frequency. The $(9F_i - 14F_o)$ component also has a frequency of 30 Hz. The harmonic having a frequency of 270 Hz has contribution from the $(3F_i + 4F_o)$ term as well as the $(6F_i - F_o)$ term. This makes the analysis of harmonic amplitudes at specific values of frequencies in close form very difficult because theoretically there are a large number of contributions from many families. It is uncertain how many of these components are significant.

The acceptability of distorted output waveforms as discussed above depends on the application. For V.S.C.F. applications the specification can be stated in very simple terms such as T.H.D. less than a certain value say ten percent and with each individual harmonics less than say five percent. For drive applications the situation is more complicated as the current harmonic is affected by the load. The limiting factors being the manifestation of harmonics as torque pulsation and extra losses. As some loads cannot tolerate torque pulsation while others are heavily damped, there is no absolute limit of harmonic amplitude above which the output becomes unacceptable.

It has been shown by Pelly (15) (P.236-238) that the R.M.S. distortion is minimum, for a circulating current cycloconverter with continuous current, when the cosinusoidal control method is employed. This is not to say that the cosinusoidal control method is the optimum control method. The current can be discontinuous in some part of the cycle. The losses in the load may not be proportional to the value of R.M.S. distortion because of factors such as winding inductance and saturation of the magnetic components in the load. For an induction motor, the inductance of the motor winding tends to reduce the amplitude of higher order current harmonics. Therefore harmonics with low frequency tend to have greater effect on the motor. The interaction of the supply harmonics and the space harmonics produced by the

motor slots is complex. A good index should reflect either torque pulsation as a percentage of the load torque or the additional losses due to harmonics in the waveform or a combination of the two manifestation of harmonics depending on application. But unless a motor as well as the nature of the mechanical load is specified, the losses in the motor as well as the effect of different harmonics cannot be determined. However, in general, harmonics with low frequency, especially sub-harmonics are to be avoided.

Although Fourier series imply there is an infinite number of harmonics, in practice when studying harmonics a finite range must be established. It can be solved by assuming that a low pass filter with a known cut off frequency is available or the harmonics having frequencies above that range have negligible effect. The cut off frequency for this work, which was chosen to be above the average switching frequency, is 500 Hz. This value is of the right order because the amplitudes of harmonics above that frequency are rarely greater than few percent and the frequency is a lot higher than the wanted component so a simple filter can be used to attenuate harmonics above that frequency.

2.3 Time domain method.

The methods of analysis described so far are all in the frequency domain. The distortion of the cycloconverter output waveform can also be quantified in the time domain.

Criteria such as the total absolute volt-sec error of the waveform may be a useful parameter. The advantage of the time domain method is that it is a single value parameter which is very convenient for comparison. Pelly's proof of minimum R.M.S. distortion with the cosinusoidal control is an example. However it does not account for the rapid changes in the voltage waveform or the high frequency component. Hence the time domain method requires further development to take that into account.

CHAPTER THREE

METHOD OF INVESTIGATION OF FIRING ANGLE CHANGES.

3 Method of Investigation of firing angle changes.

3.1 Modification of firing angles produced by the cosinusoidal control method.

As shown in section 1.3, the cosinusoidal control method is the natural choice for the generation of firing angles. This is used as a starting point of investigation to explore the possibility offered by the experimental microprocessor based control system.

Any changes of firing angle produced by the cosinusoidal control method alters the output waveform and hence alters the harmonic content. However systematic methods must be used because usually there are a large number of firing angles. Each firing angle, in theory, is a continuous variable (in the experimental system it can be adjusted by $1/512$ of the supply period i.e. less than one degree) and there are a large number of combinations of changes that can be made to the firing angles. With analog control circuitry, the modifications that can be made are limited to changing the shapes of the timing and reference waveforms to that which can be generated accurately and easily in practice - for example, having one or both of the waves triangular instead of sinusoidal or adding second harmonic to the reference wave. With microprocessor based control, it can change the value of one or any combination of the firing angles.

As suggested in section 2.2 it is not possible to eliminate all harmonics in the output of cycloconverters. Considering Pelly's proof of the desirability of the cosinusoidal control method (15) (P.236-238); i.e. it produced the minimum R.M.S. distortion, then the best that can be achieved is a better compromise of harmonic amplitudes. One or more of the more harmful low order harmonics may be suppressed in exchange for an increase of the high order harmonics. Further, Miyairi's proof (27) of the equivalence in harmonics between cycloconverters and bridge converters state that the R.M.S. voltage of the $6m\cdot F_1$ component of the bridge converter is equal to the total R.M.S. voltage of the $6m\cdot F_1$ family of the cycloconverter if the firing angle of the bridge converter (α) and the modulation depth r of the cycloconverter are related by $\cos(\alpha) = r/\sqrt{2}$. This implies that the best trade off that can be done is further restricted to within one harmonic family.

One possible method used is to change one of the firing angles produced by the cosinusoidal control method. The last pulse in the output cycle can be the longest and thus gives more scope of investigation. This is selected to study the effect of its variation upon harmonics. The last pulse in both half cycles were changed together by the same amount otherwise second harmonic will be introduced.

For an output frequency of 25Hz, which is close to the limit of the practical operating range, the firing of each pulse was varied over the possible range to study the effect upon harmonics. The theoretical possible value of firing angle is between 0 and 180 degrees. However, if the changes are restricted to affect one firing angle only, then there is a further restriction. The firing angle can not decrease by more than the conduction period of the last pulse i.e. $(2\pi/p + \phi_n - \phi_{n-1})$ and increase by more than $(2\pi/p + \phi_{n+1} - \phi_n)$. See Fig. 3.1 below.

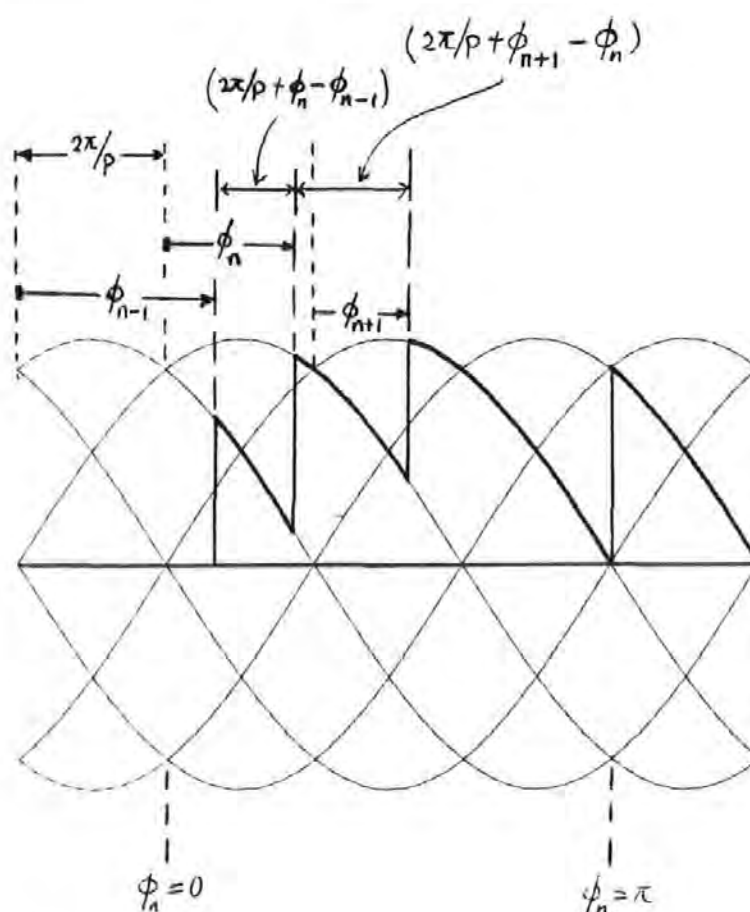


Fig. 3.1 Range of firing angle adjustment.

3.2 Variation of phase of the reference wave in the cosinusoidal control method.

As discussed in section 1.3, the cosine timing waves are synchronized to the supply, and their spacings are fixed according to converter configuration i.e. $2\pi/p$. With analog control circuitry, the reference wave is produced by a free running oscillator. It is difficult to maintain an exactly fixed frequency ratio between the mains supply and output of the oscillator. Therefore, the phase angle between the two waves with changing ratio of frequencies is meaningless.

With digital control circuitry, the natural choice is to have the cosine timing and reference waves synchronized by using the same clock as reference. This clock signal is derived from and synchronized to the supply using P.L.L. circuitry which will be described in section 4.1. Therefore the timing and reference wave are in effect phase-locked to the supply. As pointed out by Tso (28), the timing and reference wave need not be generated physically. The method suggested by Matonka (20) stores the triggering pattern only, and there is no sinusoidal signal in analog or digital form. The firing angles can simply cycle through a sequence stored as a look up table in memory.

For output frequencies which have output to input frequency ratios that can be analysed by Fourier series, i.e. frequency ratio in rational number, there is a fixed phase relationship between the timing and reference wave. The waveforms in Fig.3.2 illustrate such a condition. These waveforms are plotted by a digital plotter which is controlled by RML 380Z microcomputer via a IEEE-488 interface. The BASIC program is listed in Appendix II. The input to output frequency ratio is two and the length of one half cycle of the output equals one cycle of the input. Therefore each half cycle of the output is of equal length, and the values of firing angle in corresponding pulses is identical. That is, the phase relationship between the timing and reference wave at the start of the period included for Fourier analysis is the same as at the end i.e. the start of the next cycle. The phase between the cosine timing and reference wave is defined as zero when the instantaneous value of the cosine timing wave is one while the reference wave is zero. The value of the phase is expressed in the input frequency scale. These definitions will be used in the rest of the dissertation. The phase between them can, of course, be varied over a range of values. Fig. 3.2 show three different phase angles between cosine timing and reference waves.

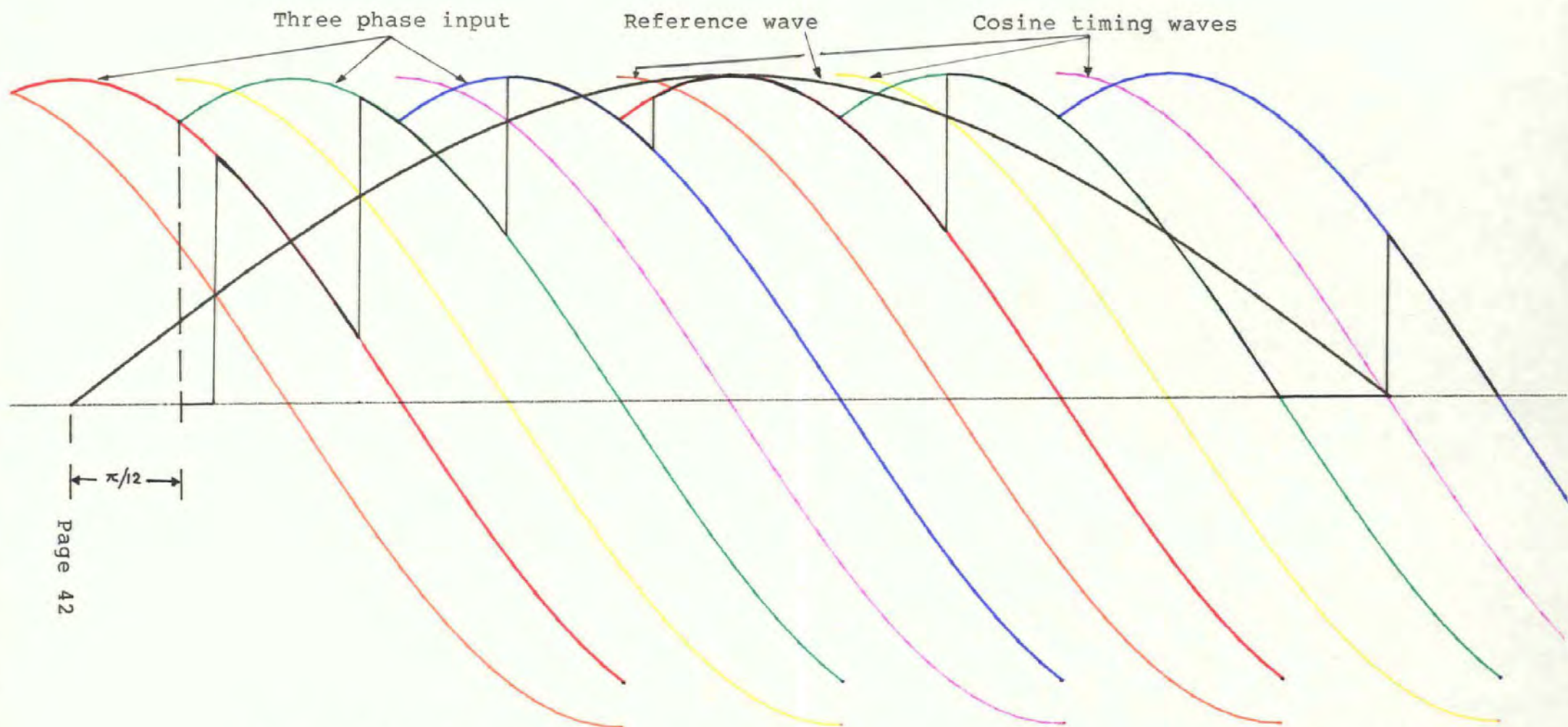


Fig. 3.2a Cosinusoidal control method with phase between start of cosine timing and reference wave equals to ϕ . ($\pi/12$)

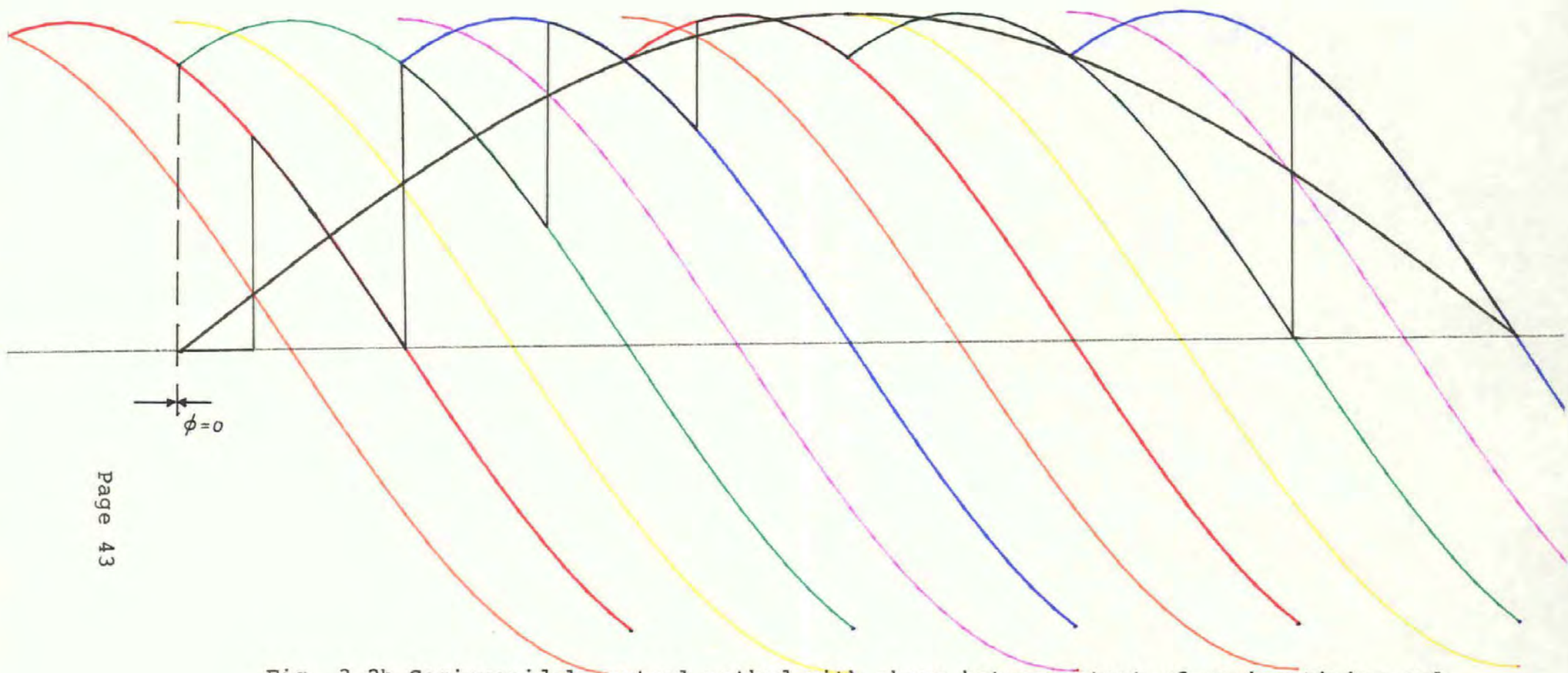


Fig. 3.2b Cosinusoidal control method with phase between start of cosine timing and reference wave equals to ϕ . ($\phi = 0$)

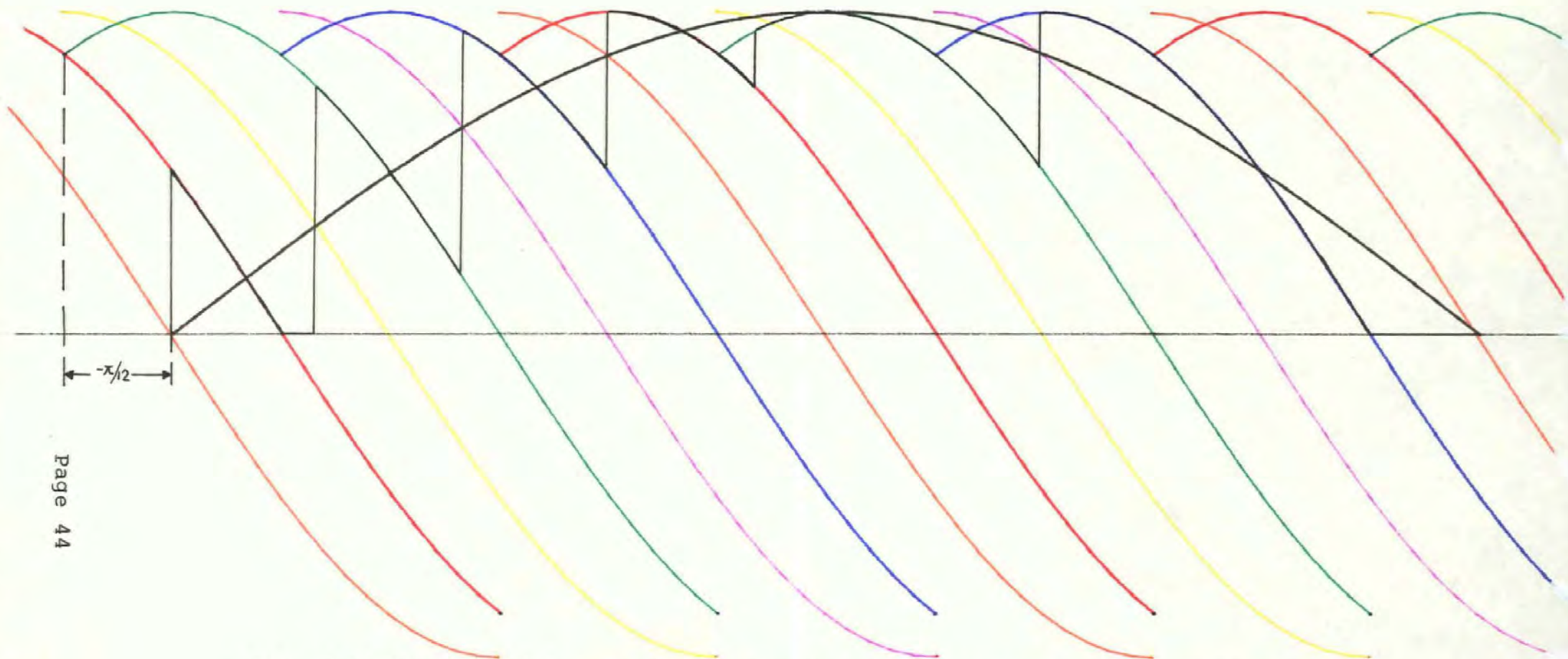


Fig. 3.2c Cosinusoidal control method with phase between start of cosine timing and reference wave equals to ϕ . ($-\pi/12$)

The first and subsequent cross-over points of the waves (the firing instants) varies with the phase between the two waves. Therefore the output waveforms with the same control method, circuit configuration, modulation depth and the same output to input frequency ratio can be different. It will be shown in chapter five that the spectrum contents change as well.

The phase shift described is not the same as the phase between reference waves in order to produce^a poly-phase output. In fact poly-phase reference waves can be shifted together with respect to the timing wave. For an output frequency of 25Hz the phase relationship between cosine timing and reference wave in the three output phases are identical. Power factor of the load does not affect the phase shift either.

Apart from the pulse number of the circuit, the range of phase angle or length of time the reference wave can be shifted before the same output waveform is produced depends on the ratio of the fundamental repetition frequency to the wanted output frequency of the output (or the quantity F_o/F_f). With six pulse operation and wanted output frequency equal to fundamental repetition frequency, the range is 60 degrees in the input frequency scale or 3.3 ms for a 50 Hz input.

If the fundamental repetition frequency is $1/n$ times the output frequency, the range is reduced by $1/n$ times. For example with fundamental repetition frequency equal to half the wanted output frequency, the range is reduced by half to 30 degrees or 1.66mS. Hence the second cycle of the reference wave is phase shifted by 30 degrees compared with the first cycle.

An undefined situation occurs if the crossing is at when the instantaneous values of both waves equal to zero. The cross-over can be included or left out in that half cycle. Fig.3.1c shows the case where the pulse is included while Fig.3.1a omitted the crossing point. The net results in the output waveform being shifted one pulse from the start to the end of the output half cycle. The problem is similar to that of the possibility of mis-fire when the converter is required to produce maximum output voltage but the end-stop control approach is not suitable as this would produce an unbalanced output i.e. with a mean d.c. level.

When the repetition frequency becomes much lower than the wanted output frequency, the range of phase becomes very small and has little effect upon firing angles on successive cycles. If the ratio of output to repetition frequency is an irrational number and the waveform never repeats itself, the phase between cosine timing and

reference wave can not be controlled. However, for a waveform that can be analysed by the Fourier series method, i.e. a waveform with finite repetition period, this phase has a finite value. This phase relationship between cosine timing and reference wave must be taken into consideration when comparing the amplitude of harmonics between different control strategies. There is no known publication that considers this variable and this is therefore a useful area of study. For some drive applications, with the output frequency fixed over a long period of time, it could well be profitable to phase shift the reference wave so that the harmonic losses produced in the load are a minimum at that frequency and load displacement angle.

CHAPTER FOUR
EXPERIMENTAL SYSTEM.

4. Experimental system.

The experimental system can be divided into four functional units: (i) the control circuit which contains the microprocessor, (ii) the synchronisation and timing circuit which enable the microprocessor to work in synchronism with the mains supply, (iii) the power circuit which performs the actual frequency conversion, and (iv) zero current detection and current waveform sensing circuit. These units will be described in the following sections. Fig. 4.1 below shows the complete cycloconverter setup.



Fig. 4.1 The experimental cycloconverter.

4.1 The microprocessor control circuit.

The basic requirement for the control circuit is to control the power circuit so that it produce the waveforms that are described in chapter 3. The waveform must be repetitive so that it can be analysed by a spectrum analyser. Hence the microprocessor is required to alter the firing angles cyclically for each pulse of the converter according to the required output. The system is required to make use of the flexibility of stored program control to investigate different control parameters and strategies. The potential saving in manufacturing using microprocessors with many peripheral functions on chip is not of importance in this project. Thus the possibility of using a single chip controller type microprocessor with program stored in ROM (Read Only Memory) or EPROM (Erasable Programmable Read Only Memory) is eliminated.

A microprocessor other than the single chip controller type usually requires some peripheral devices to perform properly. They can include ROM, RAM (Random Access Memory), PIO (Peripheral input/output) and clock signal generator. The development time to integrate such parts into a functional unit can be much reduced or eliminated altogether if a ready-built development system can be used. These systems usually provide some monitor functions with a key pad and some form of display so that the user can inspect and change the content of memory and registers. The execution of user program can be controlled by single stepping and setting up various break points in the program using facilities provided by the monitor so that the user can verify the logical operation at various stages in the program.

However, for this real time control application where there is only around a milli-second between successive firing pulses, the microprocessor control circuit is required to respond within that time. Therefore, it is desirable to have full control over the interrupt facility and not rigidly controlled by the monitor, at least when the user program is running so that the user program can respond to external events in the shortest possible time and assign priority to events according to their urgency. In this case the firing angle jittering due to slow interrupt response must be minimized.

A method of storing program and data in permanent form such as on cassette tape or floppy disc, or in semi-permanent form such as battery back-up RAM, is essential. The system should also be flexible enough so that it can be expanded at a later stage if required, for example to increase the amount of memory.

There is another type of microprocessor development system which emulates the action of a microprocessor in the control circuit during development. Instead of adding components to the development system, a separate self-contained control circuit - the target system is built. But the microprocessor is unplugged and the emulator is used in the target system instead. These emulators are very flexible in exercising parts of the control circuit and are suitable for the development of single chip controller type microprocessor systems. When all the development work is finished the microprocessor is then put back in place of the emulator.

Comparing with most eight bit microprocessors, sixteen bit microprocessors with more processing power are more suitable for the control of cycloconverters which require evaluation of sinusoidal functions. Although a sixteen bit microprocessor itself is not expensive, a sixteen bit microprocessor development is at least an order

of magnitude more expensive than an eight bit system. This may account for the absence of reports on developments using sixteen bit systems. The unit selected for this work is the MPF-1 'Microprofessor' single board computer manufactured by Multitech Industrial Corporation. The unit contain most of the essential parts including a Z-80 microprocessor running at 2MHz. Only one Z-80 PIO chip is added to the standard unit which is a straight forward plug in operation.

Some facility to assist program development is also desirable. Program development work can be performed on a separate computer (micro, mini or even main frame computer) as long as the codes developed can be transferred into the development system easily. This can be accomplished either by an electrical connection such as a RS-232 serial link, or by programming the codes into an EPROM and which is then plugged into the development system. A Research Machine 380Z microcomputer running a Z-80 assembler was used for the development of ^{the} assembly language program.

There are a large number of programming languages available for microcomputers. They range from low level machine code language to high level language such as BASIC. The trade-off is between ease of programming, good arithmetic and scientific functions for high level languages and fast execution speed for low level languages. Generally high level programming languages are more efficient to

program e.g. it is an easy matter to perform floating point multiplication in BASIC. The most common language for a microcomputer is interpreted BASIC which is slow compared with other compiled high level languages such as Pascal and compiled BASIC. However, these high level languages are incapable of dealing with interrupts efficiently which is essential for real time control. Forth is a threaded language which requires the programmer to define words which are treated in the same way as vocabulary in the language itself. The language is stack orientated and uses Reverse Polish Logic notation. Execution is fast compared with BASIC but usually offers integer arithmetic functions only. It is a good compromise if the control function is not too demanding.

A six pulse converter bridge, with supply frequency at 50Hz, has on average 3.3ms between each pulse. The control program is required to obtain the correct firing angle for the next pulse and place it in the correct channel, hence firing the correct thyristor at the desired time. It also has to decide if it is necessary to change converter bridge to allow current flowing in the opposite direction. There are other secondary functions which are desirable if it can be incorporated. For example, to sense if current is exceeding the limit or device temperature is too high. A display to indicate the state of the controller is also desirable. For the present generation of eight bit

microprocessors, the clock speed is usually between one and five mega-Hertz. Hence a basic instruction such as load accumulator takes around a micro-second. With high level languages there is a lot of overhead to deal with variable names, line numbers etc. and cannot handle the required operation within the 3.3mS available. The only viable option is using assembly language.

4.2 Synchronisation and Timing.

In order to modulate the firing angles and hence to control the output voltage waveforms of the cycloconverter, synchronisation of the control program and hence the firing circuit to the supply is essential. A useful timing reference is to establish the earliest instant of time where natural commutation can occur between the conducting thyristor and the incoming thyristor. The firing angle is customarily defined as zero at this instant. The desired firing angle or controlled time delay for the firing pulse can then be implemented using this timing reference. The circuit used was based on a design published by Bhat (29). The block diagram of the circuit is shown in Fig.4.2.

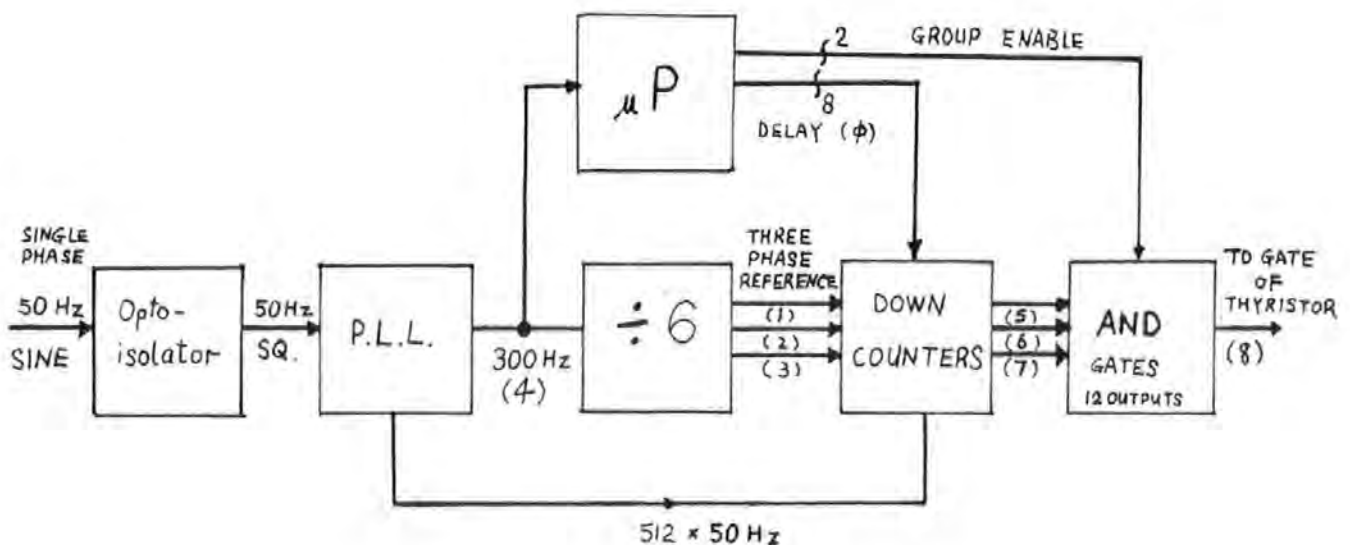


Fig. 4.2 Block diagram of the control circuit.

The function of synchronisation was performed by a Phase Lock Loop (P.L.L.) circuitry because of its inherent out of band signals rejection characteristic which minimise the timing error that may be introduced by noise in the supply. The requirement for this P.L.L. circuit is not demanding. As lock-in transients only occur at switch-on and the triggering of thyristors can be delayed until the P.L.L. circuit becomes locked, hence the pull-in time is not critical. The supply frequency is fairly constant and variation is well within the lock range of the P.L.L. if the centre frequency has been selected correctly. Using the CMOS 4046 P.L.L. circuit with its phase comparator II selected, the output is always in phase with the input when the P.L.L. is in lock independent of the working frequency.

In order to supply a suitable signal to the P.L.L. the supply voltage in one phase of the supply was stepped down, rectified and reshaped by amplification to saturation using the transistor in the opto-isolator to become a square wave. The mark space ratio of this signal is not exactly one to one but that introduces only a negligible phase error to the timing signal.

There is a divide by six counter between the output of the voltage controlled oscillator and the input of the phase comparator in the P.L.L.. That is the P.L.L. is

working in the frequency multiply mode. One of the outputs from the P.L.L. circuit, trace [4] of Fig.4.3, is a square wave which is at six times the supply frequency and synchronised to the supply. This signal which contain all the timing references required by a six pulse bridge, i.e. all the starting points of the cosine timing waves, is used to interrupt the microprocessor. The divide by six counter produces three signals, traces [1],[2] and [3], with 120 degrees phase shift between each pair of them. These three square waves of supply frequency are in phase with the three phase supply. Similar methods for obtaining a three-phase reference from one phase of the supply were reported by others in the field (21)(30).

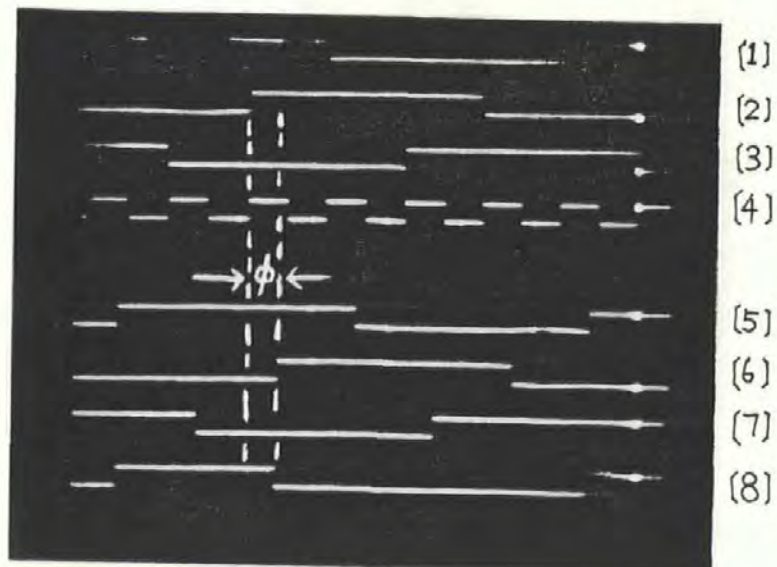


Fig. 4.3 Timing waveform of the timing circuit to generate firing angle (ϕ).

The sequence of firing angles used for a particular output was worked out on a programmable calculator using the cosinusoidal control method and then transferred into the microprocessor memory. The microprocessor then follows the machine code program to cycle through the sequence of firing angles stored in memory to produce the desired fixed frequency output. Any or all of the firing angles stored in memory can be altered using the development system monitor to study the effect upon harmonics. The assembly language program that controls the firing angle is listed in Appendix III. The program is under interrupt control as described in the next paragraph. As the microprocessor receives only timing signals that are derived from the supply, any variation in the magnitude of the supply voltage is not compensated by change of firing angles. This is in contrast with the self regulation of output voltage against variation of input voltage obtained by using analog control circuitry when the cosine timing waves are derived from and with amplitude proportional to the supply. This self-regulation effect was described by Pelly (15) (P.231).

Like the development of most control circuits using microprocessors, there is a lot of flexibility in choosing hardware or software to perform a particular function. The trade off is between work load for the microprocessor including the number of interrupts and possibility of

simultaneous occurrence while it is running and effort of programming against cost of extra components and construction effort. In this case, to reduce the work load of the eight bit microprocessor, the conversion of desired firing angles into actual time delays is performed by logic counters. The microprocessor is interrupted by the synchronisation signal via the strobe line of the P.I.O. The microprocessor then outputs the firing angle as an eight bit binary number into the P.I.O. This number is then loaded into the appropriate counter at the instant when the microprocessor is interrupted again. The counter starts to count down and produces a delayed version of the input. For example with constant firing angle trace [6] is trace [2] delayed by ϕ . The three delayed square waves, trace [5], [6] and [7] together with their inverses are combined using AND gates to produce six firing pulses. For example, trace [8] is in logic terms : trace [5] AND NOT trace [6]. There is also a third input to the AND gates for group enable.

The clock for the logic counters, with a frequency of 512 times the supply frequency, is derived from the supply by another P.L.L. circuit. The smallest change of firing angle is $1/512$ of the input cycle which is less than one degree. This is considered as accurate enough for the investigation.

If the supply frequency is shifted from its nominal value of 50Hz, a phase angle of ten degrees, say, will correspond to a different length of time. The delay produced by the counters, however, is a true function of phase angle of the supply and not a multiple of fixed unit of time. This is because the clock signal for the counters is a multiple of the supply frequency and not a fixed frequency signal produced by a crystal oscillator. Therefore the frequency of the wanted output component is not fixed solely by the parameters on which the firing angle is calculated but changes in proportion to the input frequency. This is an advantage when the amplitudes of harmonics are being investigated. For the amplitude of harmonics will remain the same but frequencies of the harmonics will be shifted by an amount proportional to the input frequency as the harmonic frequency is a function of the input and output frequency. For example an one percent increase in frequency of the supply will mean an one percent increase in frequency of the wanted output component and all its harmonics. On the other hand, if time delay produced by the counter is the same irrespective of supply frequency, the output voltage waveform is different and the harmonic content will not remain the same. Any changes of harmonic amplitude due to variation of input frequency may mask the changes of harmonic amplitude due to other effects under investigation.

For practical purposes, the frequency and voltage of the mains supply is fairly constant. The Central Electricity Generation Board (CEGB) is required by law to maintain the supply between 49.5 and 50.5 Hz and the operation limit set by CEGB itself is 49.8 to 50.2 Hz. Therefore the effects of voltage and frequency variation described earlier are negligible. The voltage regulation, however, may be significant when the supply is coming from the end of a long transmission line. When operating from a small generator, the voltage and frequency are both changeable, feedback may be required to produce a steady output at wanted frequency.

With the dual three phase six pulse converter bridges, the microprocessor is interrupted six times in an input cycle i.e. every 3.3 mS. As only one bridge is triggered at a time and the two thyristors connected to the same supply line in a bridge would not be triggered at the same time, only three sets of counters are required for the dual six pulse bridges. The outputs of the counters and converter bridge selection signals are combined together using AND gates. The firing sequence and the matching of firing pulse to the correct thyristor is fixed by the wiring. Therefore, at the time of construction the phase sequence and phase shift with respect to the synchronisation signal of the supply as connected to the converter bridges were measured to match up with the firing pulses.

With more powerful microprocessors that run at higher clock rate, e.g. a sixteen bit processor running at 12MHz, the external counter arrangement may become unnecessary. Chip count would be much reduced by using a single chip counter / timer that has three separate counter channels. The microprocessor can be interrupted by the counter directly and then control the firing circuit directly via a P.I.O.. As the firing sequence is not fixed by wiring, the microprocessor can identify the phase sequence and phase shift of the three phase supply with respect to the reference signal at power up. Then the firing pulse can be steered automatically to the correct thyristor and ^{thus} eliminating the need of connecting the three phase supply in a pre-determined order.

4.3 The Power circuit .

For isolation between control and power circuits, the firing pulses are applied to opto-isolated thyristors. See Fig.4.4. These were chosen to give maximum protection to the microcomputer. A pulse transformer with a high voltage spike across the secondary can induce enough energy to destroy the microcomputer and all other electronic circuits. As the current rating of the opto-isolated thyristor is not large enough to carry the main current, it is used to trigger the gate of the main thyristor i.e. a cascade connection. The signal going into the diode part of the opto-isolated thyristor is a continuous pulse for the whole duration in which the main thyristor is supposed to conduct. Unlike in the case of a pulse transformer, there is no need to modulate the pulse with a high frequency. However, once the main thyristor is on, the voltage between cathode and gate becomes low and the gate of the main thyristor is not driven by the opto-isolated thyristor. For loads with long time constant, or ^{if} the current becomes discontinuous, the gate of the main thyristor remains driven until it is completely on.

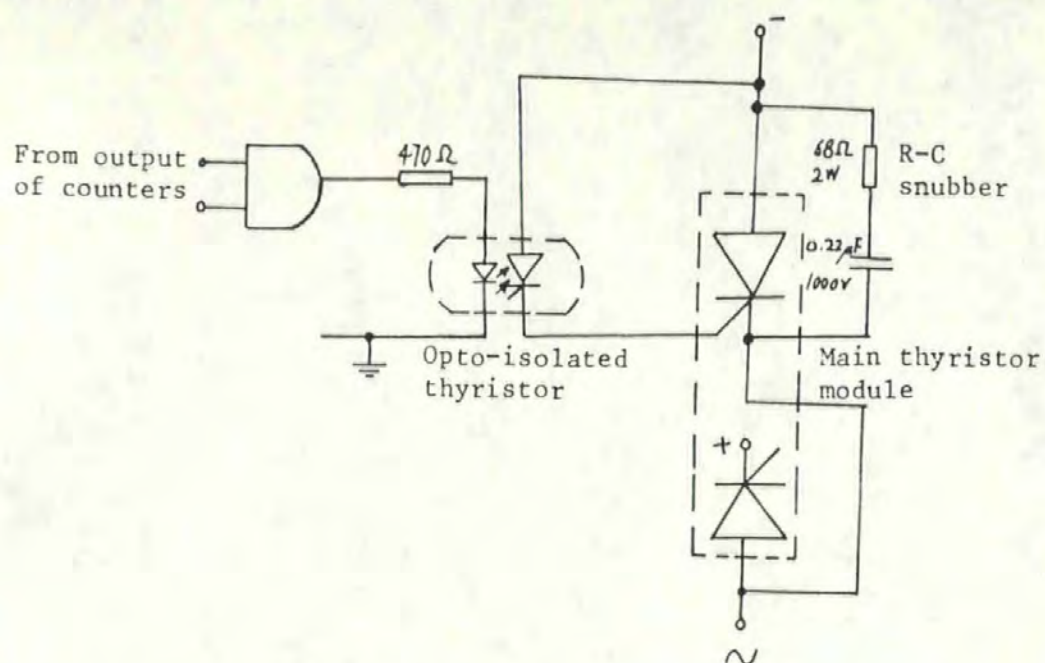


Fig. 4.4 Details of trigger circuit, R-C snubber and thyristor module.

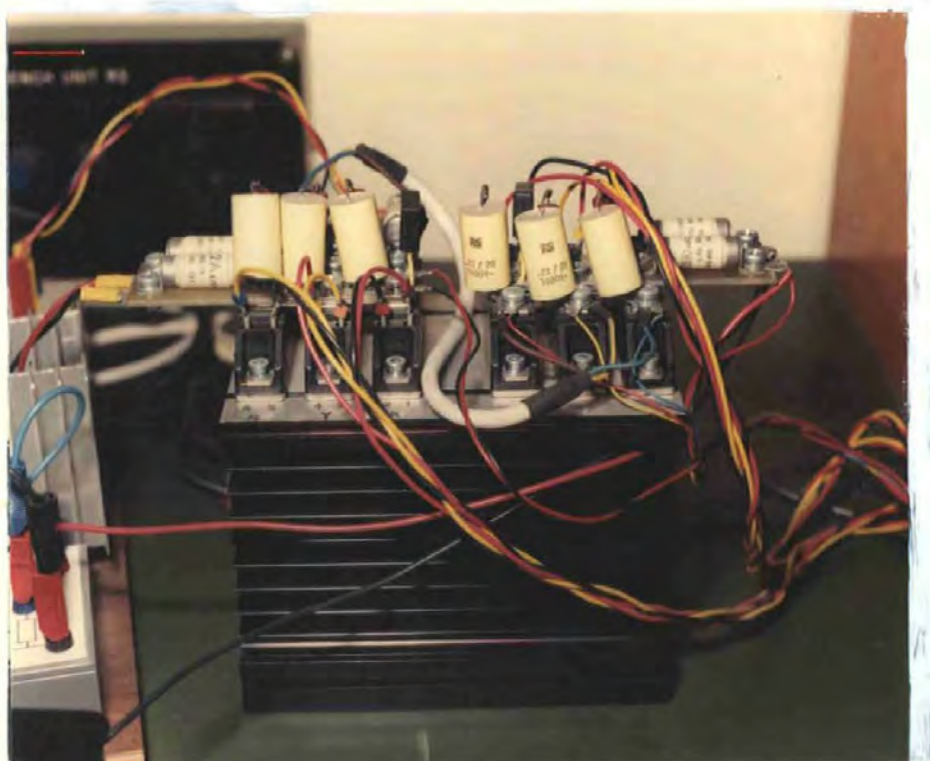


Fig. 4.5 Details of the power circuit construction.

The power part of the circuit is shown in Fig.4.6. The main thyristor bridges consist of six thyristor modules. Each module consists of two thyristors on isolated casings. These modules cost no more than conventional packaging but provide a very easy and cheap method of construction. All twelve thyristors of the three phase input, single phase output cycloconverter were mounted on the same heat sink. The connection of thyristor modules into ^{the} converter bridge is by two copper bars. The thyristors are rated at 18A and 800V. With the supply at 200V and maximum current of about 5A, this offers a large margin of safety and hence an easier task of protection. Also 6A ordinary HRC fuses instead of semiconductor fuses were used to reduce cost. The half cycle ten times over-current capability of the thyristor should give enough over-current protection. The R-C snubbers and transient suppressors of the voltage dependent resistor (VDR) type are mounted directly on top of the thyristor modules and thus reduce the length of conductor and its associated inductance. These R-C snubbers and transient suppressors reduce the rate of rise of the off-state voltage and magnitude of the voltage spikes, thereby also suppressing mains and radio interference. Fig.4.5 shows the construction detail including thyristor modules, heat sink, fuses and snubbers.

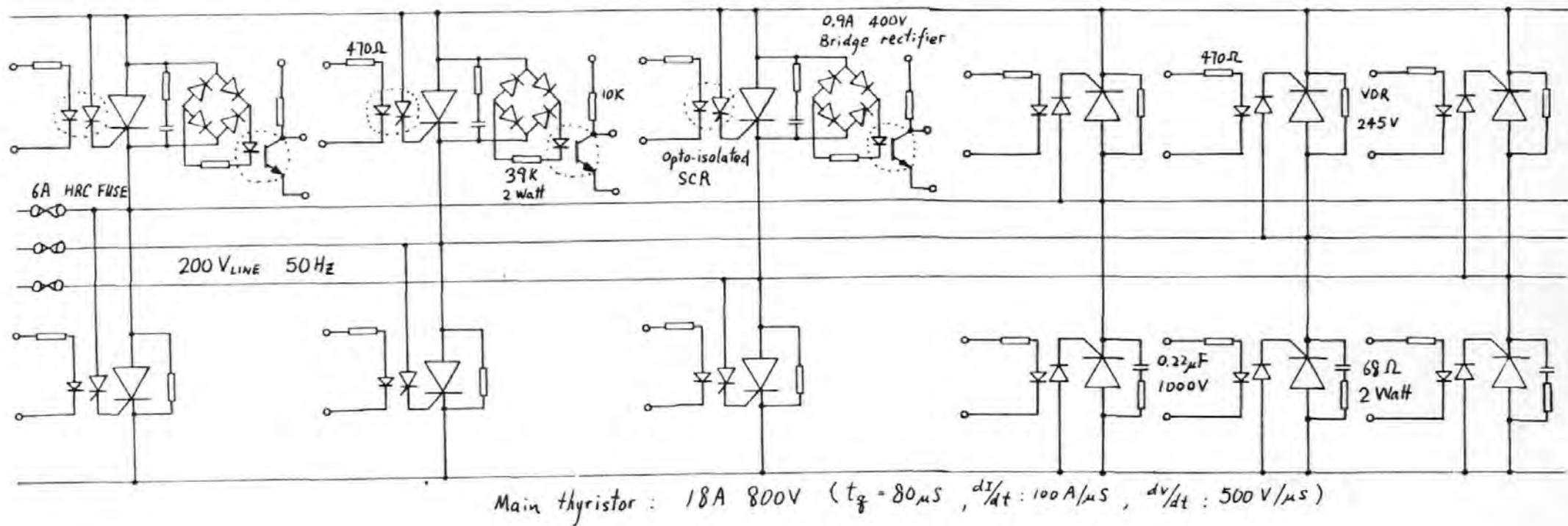


Fig. 4.6 Power circuit of the experimental cycloconverter.

4.4 Zero current and current waveform sensing circuit.

In order to operate the inhibition of the converter bridges, a zero current detection circuit is required. This is done by sensing the voltage across the thyristors. If a thyristor is conducting, there is only a small voltage across it. The thyristor would support a much higher voltage if it is not conducting. There is no current flowing if the voltages across all thyristors are higher than the conducting voltage drop across a thyristor. The draw back of this method is that when the supply voltage is near zero, the sensing circuit can not distinguish this from a conducting thyristor. One way to overcome this problem is to have a dead period, with neither bridge enabled, longer than the period of supply voltage that has voltage less than the voltage drop across a conducting thyristor. With a three phase supply, when one phase supply voltage is near zero, the other two phases are of course not near to zero. A zero current indication longer than the dead period is therefore an unambiguous zero current indication.

For a six pulse bridge circuit,^{since} the current flowing out of one thyristor connected to the positive rail comes back to another thyristor connected to the negative, only three voltages require checking. The two converter bridges of cycloconverter are connected back to back, i.e. the thyristors form anti-parallel pairs. Therefore the number of

voltages require checking remain three. The voltages in question are full wave rectified and fed into opto-isolators. The output from the opto-isolators are combined using a NOR gate. This one bit signal, ^{which} only shows if there is current flowing through the thyristors, is used as current feed back signal. The magnitude of the current is not available to the microprocessor. If current feedback is required a separate sensor is needed.

With the star point of the three phase supply earthed and an earthed C.R.O., the voltage of both output rails with respect to ground can be displayed. The total output voltage can be displayed on a C.R.O. with 'ADD' and 'INVERSE' facility.

The current waveform, however, can not be displayed on a C.R.O. using a dropping resistor. A Hall effect current sensor was constructed. The circuit diagram of the sensor is shown in Appendix IV. The current of interest is passed through a coil wound onto a 'C' shape ferrite core. The flux set up across the air gap, which is proportional to the current through the coil, is sensed by a Hall effect device. The air gap is just wide enough to accommodate the Hall effect sensor to create a maximum flux across the gap and hence maximum sensitivity for the sensor. The output voltage from the Hall effect sensor is amplified and calibrated against a known current. The frequency response

of this current transducer is from d.c. to more than 1 KHz.

A useful variation of the transducer circuit is to have two coils wound on the same core. The current flowing in each coil is completely isolated but the flux set up in the core is due to both coils and the waveform of the sum of current is produced by the Hall effect transducer. This enables experimentation of the group change over control without any danger of thyristors being damaged by a large current circulating between the two converter bridges. If loads with equal impedance are connected to each group, the resultant current waveform is the same as for the current flowing through a single load. If both bridges are conducting at the same time, this error would be shown up as an unexpected zero current in the current waveform and possibly shows current goes down to zero in a large step.

4.5 Computer simulation.

A Fortran 77 program running on a PRIME computer is used to simulate the experimental cycloconverter. This provides a means of cross checking the harmonic content. The program as well as a set of input and output data are listed in Appendix V . Figure 4.7 is the flow chart of the program.

The program is organised to get its input from a data file. The data file can contain more than one set of data so that the program can simulate the variation of one of the parameters over a range of values. The actual inputs of the program are : input frequency, number of pulses in a wanted output cycle, frequency ratio of wanted component and lowest component in the output, highest harmonic component to be evaluated, time constant of load, modulation depth. Also the phase between cosine timing and reference wave or the value of the firing angle for the last pulse in an output half-cycle can be included.

From these inputs, other constants such as output and repetition frequency and the number of wanted output cycle that need to be considered are calculated. Next, all the crossover points of the cosine timing and reference waves for both ^{the} positive and negative group are calculated. Then the group change over instant is determined and the

appropriate switching instants for positive and negative groups are selected. These switching instants are sufficient to completely define the output voltage waveform. (neglect overlap and device volt-drops) From these switching instants, the individual harmonic amplitude is calculated using the modified Fourier series described in 2.1.2.

The outputs of the program are the frequency and amplitude of the harmonics. Also some other information such as the group change-over point are printed to provide more information to check the results. The amplitudes of the harmonics are also written into an output data file. The data stored in these files can be plotted using library routines provided by the PRIME computer.

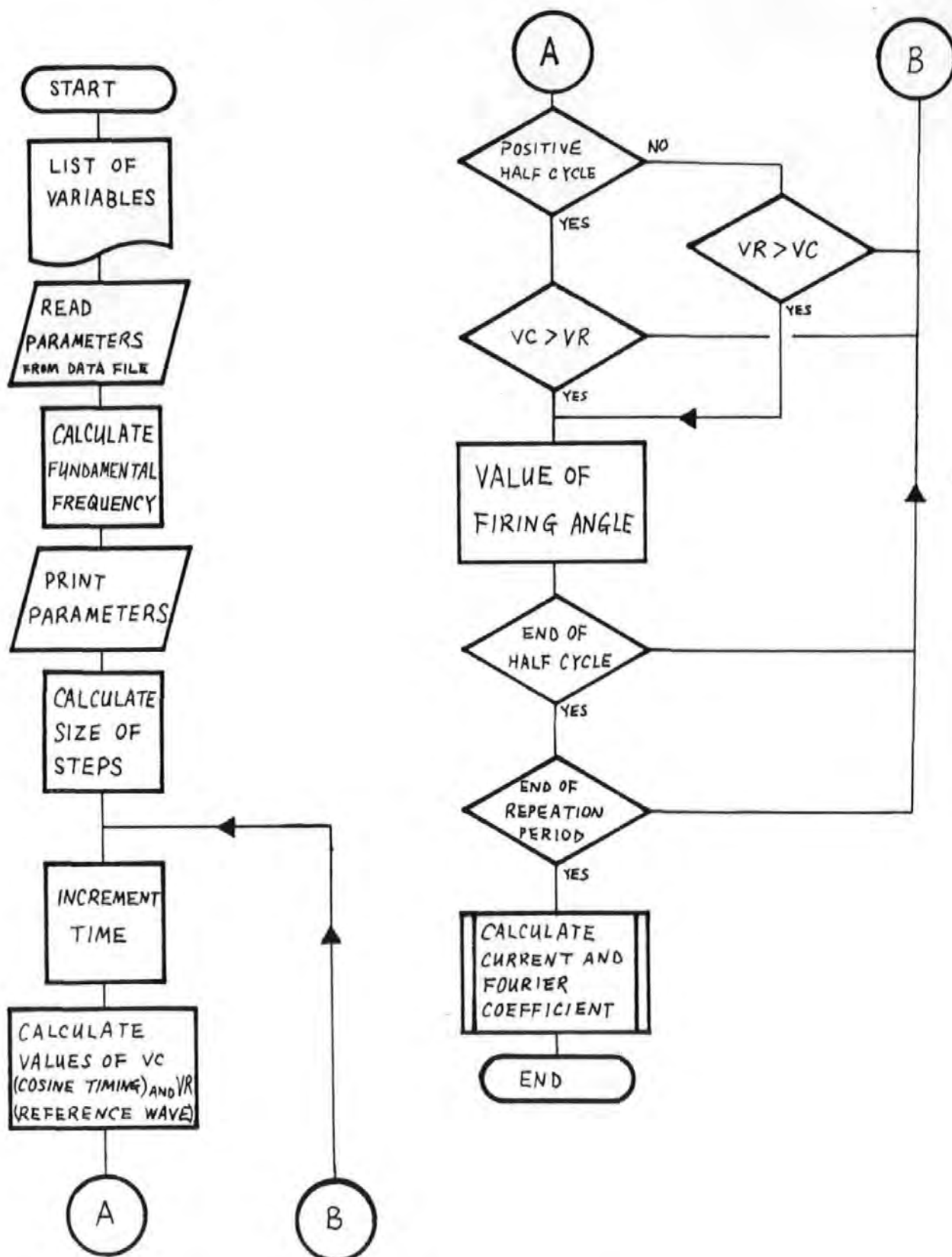


Fig. 4.7a Flow chart of firing angle calculation.

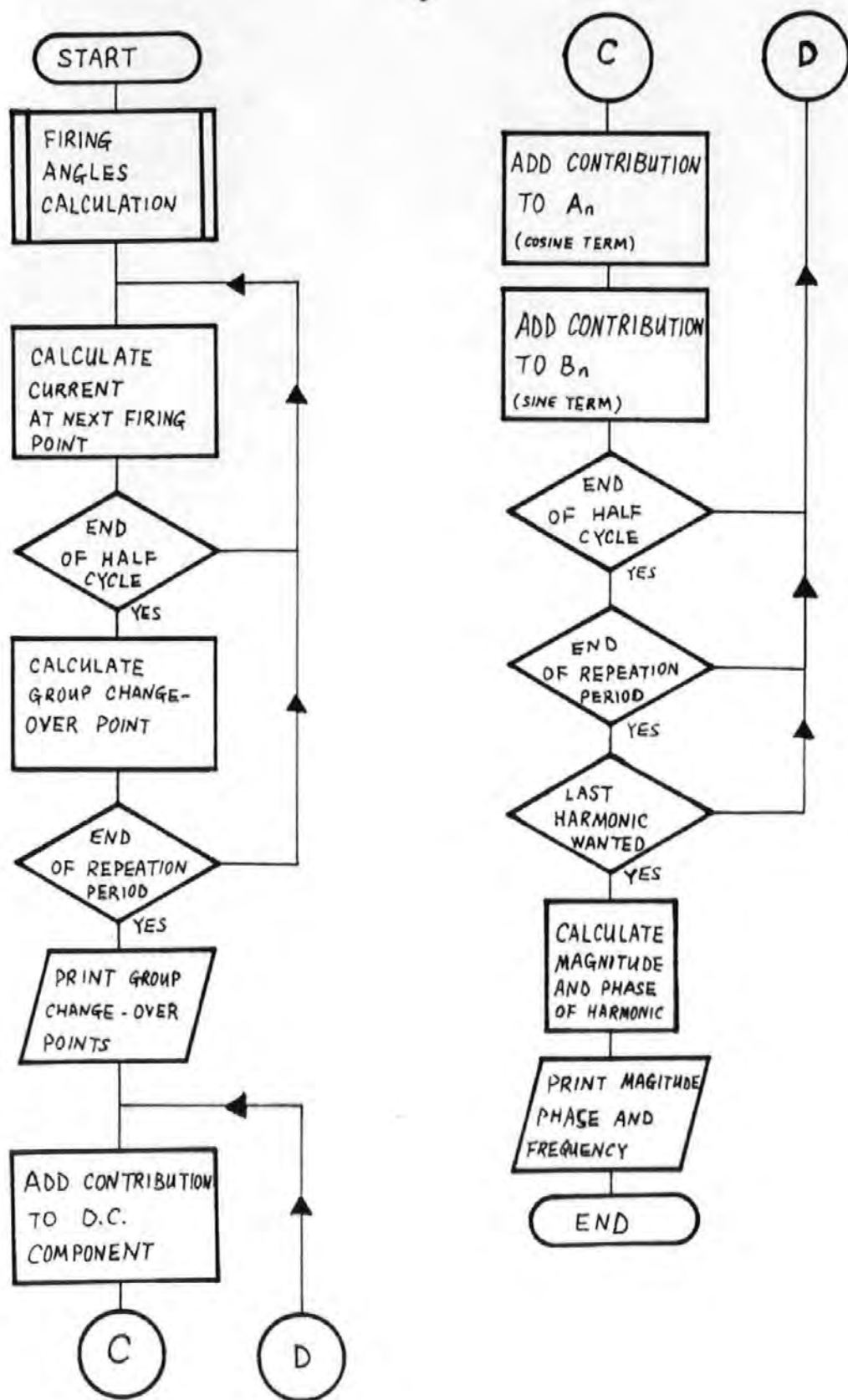


Fig. 4.7b Flow chart of current and Fourier coefficient calculation.

CHAPTER FIVE

EXPERIMENTAL RESULTS.

5 Experimental results.

5.1 General observation techniques.

The experimental cycloconverter performs satisfactorily and produces all the output waveforms as required. The supply for the experimental cycloconverter is 200V three phase at 50Hz. The load for the single phase output consist of two 50 Watts 1000 ohm resistors in parallel. Therefore the output voltage and current waveforms are of the same shape. The output from the Hall effect current transducer which is of suitable magnitude is used for harmonic analysis by the spectrum analyser and displayed by the oscilloscope.

When the experimental cycloconverter was first operated, the output waveform was proved to be stable by storing the waveform on an analog storage oscilloscope with continuous triggering over a period of time say two minutes. Therefore a large number of waveforms, around a hundred say, are stored one on top of another. Instability or malfunction of the control circuit within that period of time which produce a different output waveform would show up as extra traces. However, special attention must be given to the trigger signal. In general, for a cycloconverter output voltage waveform, a clean-up version of the actual display is desirable for triggering the oscilloscope in order to provide a stable display. Some oscilloscopes offer an a.c. coupled trigger input with high frequency rejection which provide satisfactory triggering. Otherwise a simple low-pass

R-C filter can be used to clean-up the trigger signal. In any case one must ensure that such filtering on the trigger signal does not eliminate the display of spurious output waveforms.

The time domain waveforms reproduced in the following pages are first recorded using a digital storage oscilloscope and then plotted on a flat-bed X-Y plotter. In contrast with the procedure adopted in the early stage, the waveform obtained is a single trace recording. The consistency of the output can still be checked by comparing one stored trace with another. The pre-trigger viewing and post-storage expansion facility of a digital storage oscilloscope also help to measure firing angles accurately. By using different amounts of pre-trigger, different sections of the output can be stored. This combined with the post-storage expansion enables the display of just one or two pulses on the screen to enhance resolution. The firing angle can also be measured quite accurately using a non-storage oscilloscope with a delayed time base. However, the display is difficult to read because there is a reduction of display intensity and a tendency to flicker with the low output frequency of the cycloconverter.

The harmonic spectrum produced by the spectrum analyser for each of the time domain waveforms is also plotted on the same graph paper which contains the time

domain waveform. As the two traces are plotted together at the same time, this technique eliminates the possibility of attributing a spectrum to a different waveform. The amplitudes of the individual harmonics shown are read-out from the spectrum analyser. They are normalised to the wanted component by setting the wanted component as reference on the spectrum analyser. Normalised harmonic amplitudes are suitable for comparing the cycloconverter against other frequency changers such as the quasi-square wave inverter. Normalised harmonic amplitudes are required for comparing the cycloconverter output waveform with a different input voltage specially for the 200V supply used which is not the 415V common standard. For VSCF applications normalised harmonic amplitude is a good starting point for calculating percentage T.H.D. and is a useful indication of the acceptability of the waveform. However for drive applications, extra losses in the motor are related to the absolute harmonic amplitude. With non-unity modulation depth, i.e. less than maximum output voltage, the process of normalisation is actually a retrograde step for determining extra losses in the motor. In this study almost all the waveforms investigated are with unity modulation depth.

5.2 Variation of phase between reference and cosine timing waves.

5.2.1 Practical results.

To investigate experimentally the effect of the variation of phase between cosine timing and reference waves upon harmonics, the three waveforms obtained in section 3.2 were reproduced using the appropriate sets of firing angles in the control program. The firing angles used are shown in Fig.5.2 against the appropriate pulse on the time domain waveform in hexadecimal form as stored in memory. (80 HEX = 90°) Table 2 below lists the firing angles used. In general, when the phase is different, all the firing angles require changing but not necessarily by such uniform steps as shown in Table 2. The change of firing angle is more irregular if the modulation depth is not unity. Fig. 5.1 shows the output of the timing circuit. Note that each pulse is the same length as the corresponding conduction period and there is a large overlap between pulses. It can be seen from the experimental results in Fig.5.2 that they have close resemblance to the expected waveform as drawn in Fig.3.2 (P.42-44). A useful method to identify different waveforms is to observe the firing angles near the maximum wanted output voltage. Using this method, the three waveforms are readily identified with the expected waveforms.

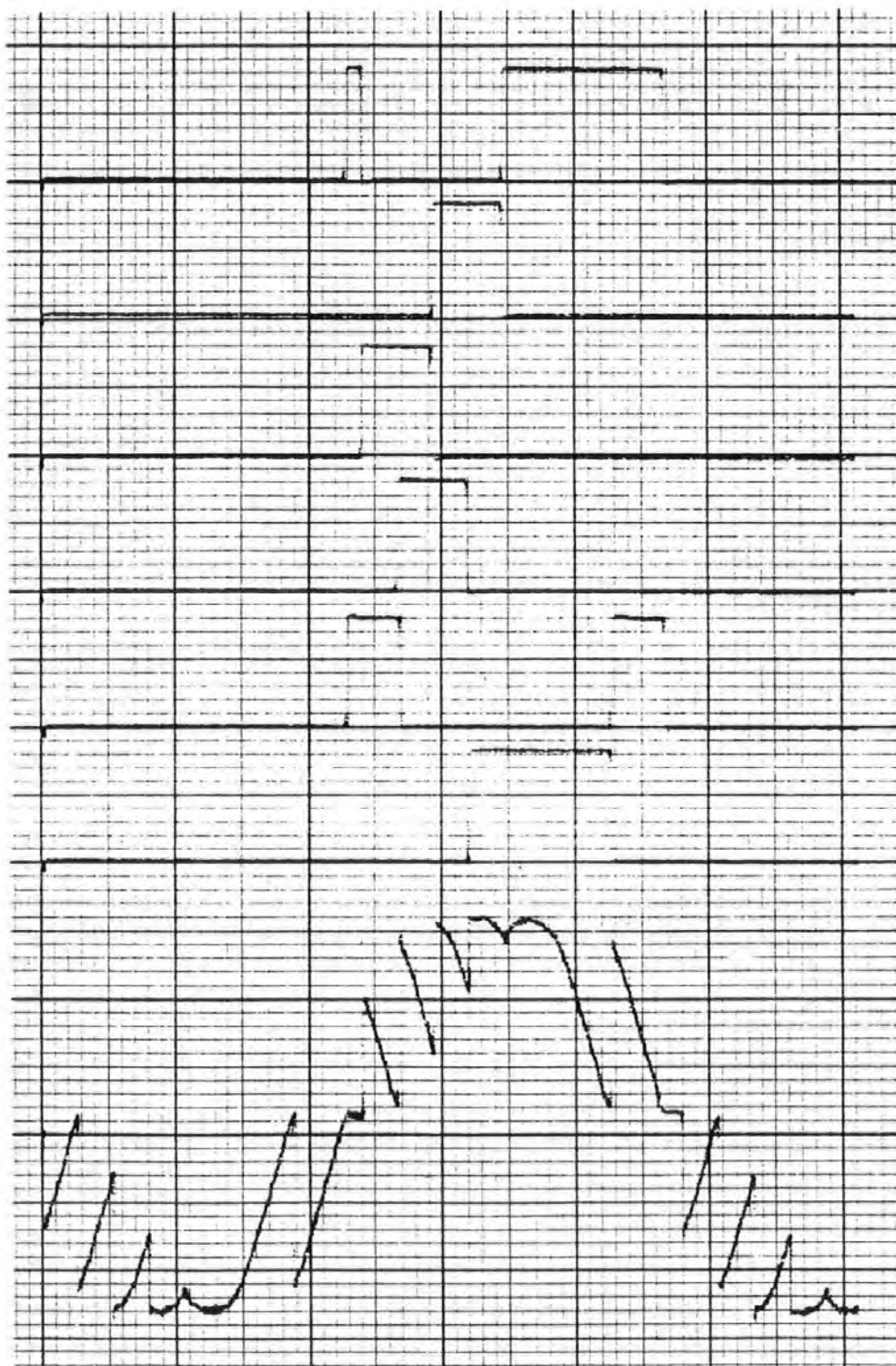


Fig. 5.1 Output of the timing circuit for a typical cycloconverter waveform.

Table 2 Firing angles used to produce the experimental waveforms.

Phase	Firing angles in Hexdecimals (80 Hex = 90°)					
$-\pi/12$	80	64	47	2B	0E	2B
0	72	56	39	1D	01	56
$\pi/12$	64	47	2B	0E	2B	80

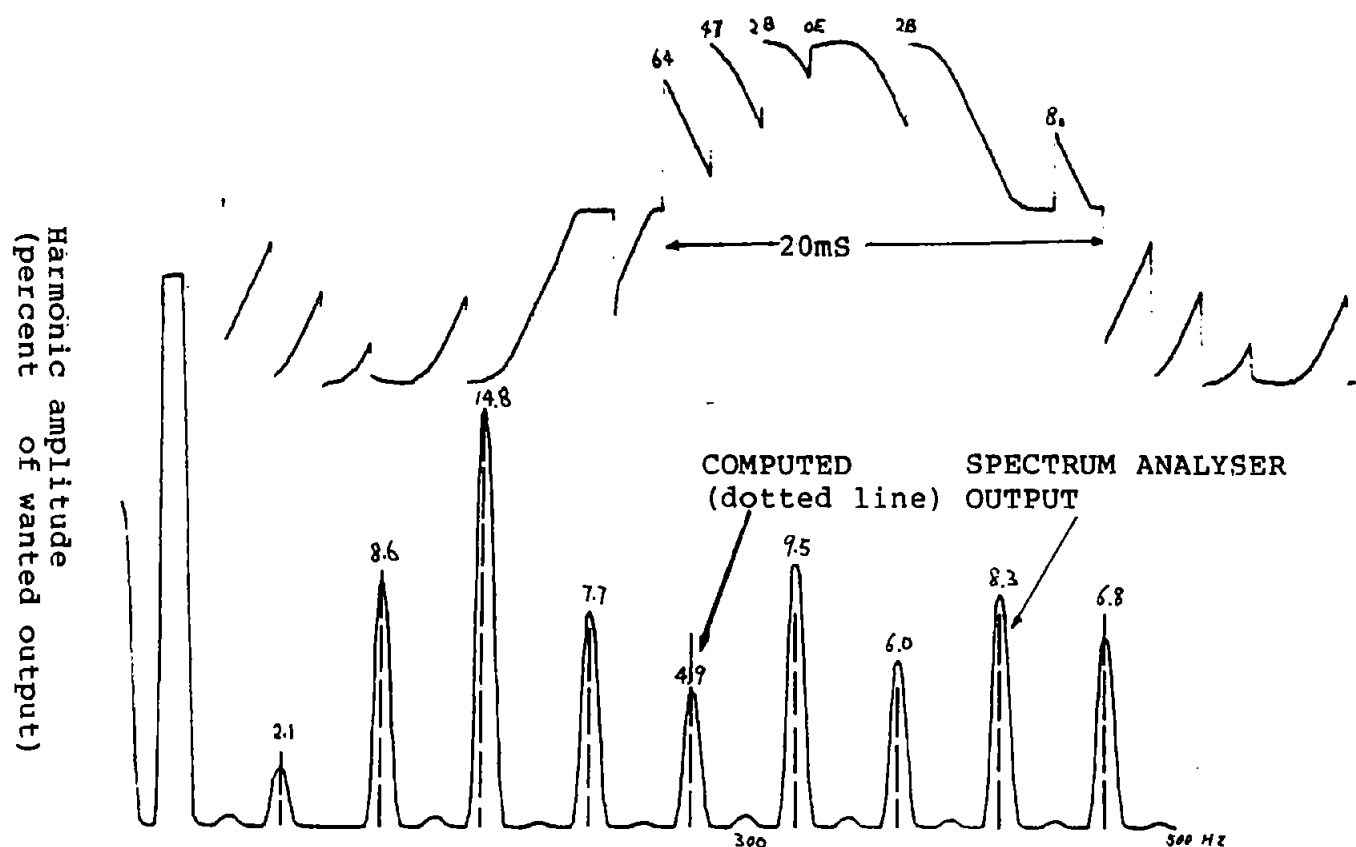


Fig. 5.2a Phase shifting between cosine timing and reference wave.
 $\phi = \pi/12$

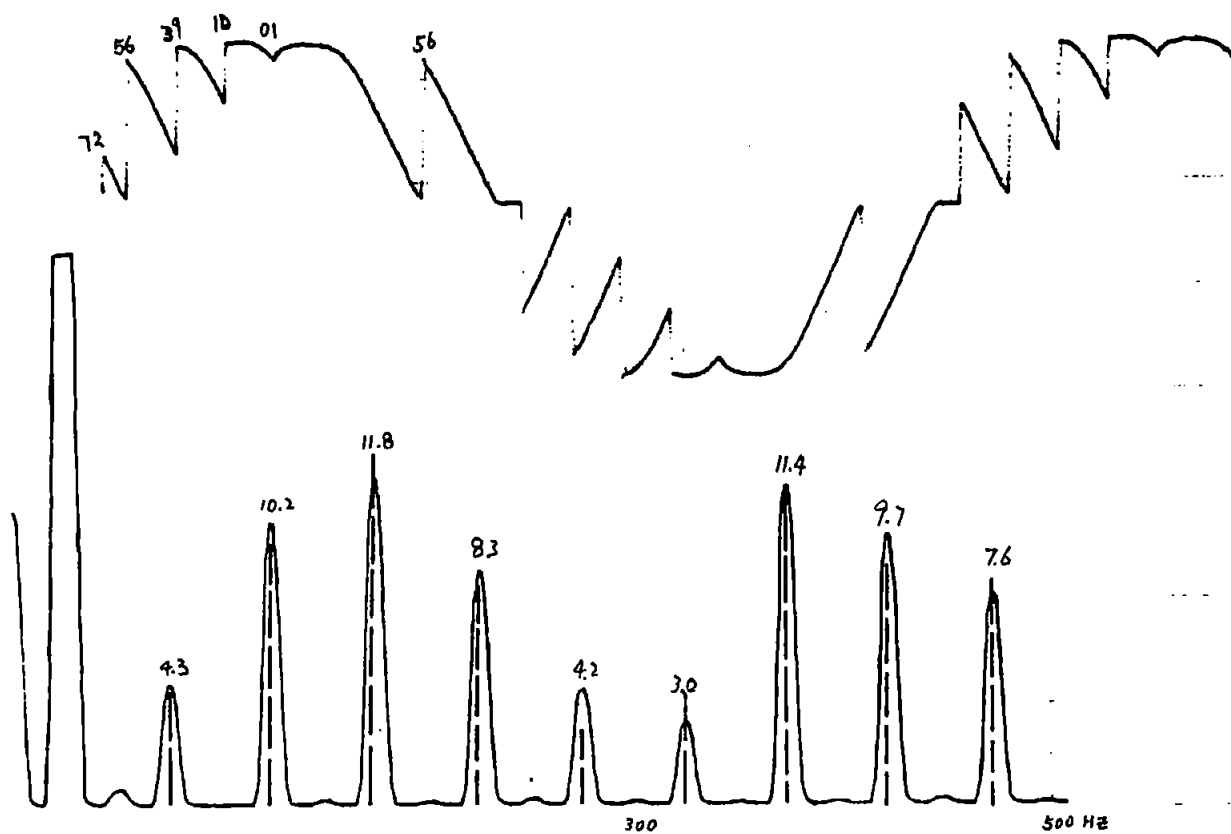


Fig. 5.2b $\phi = 0$

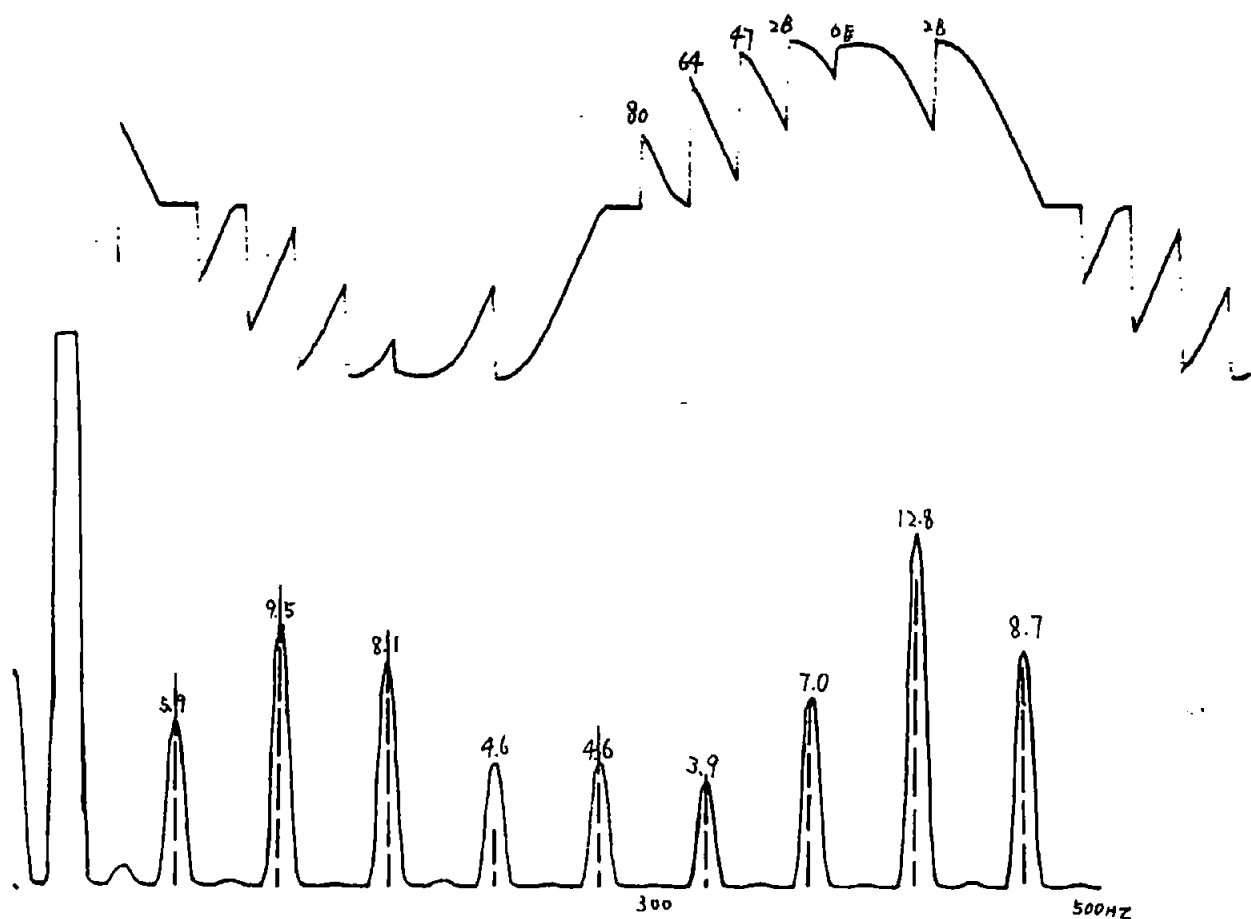
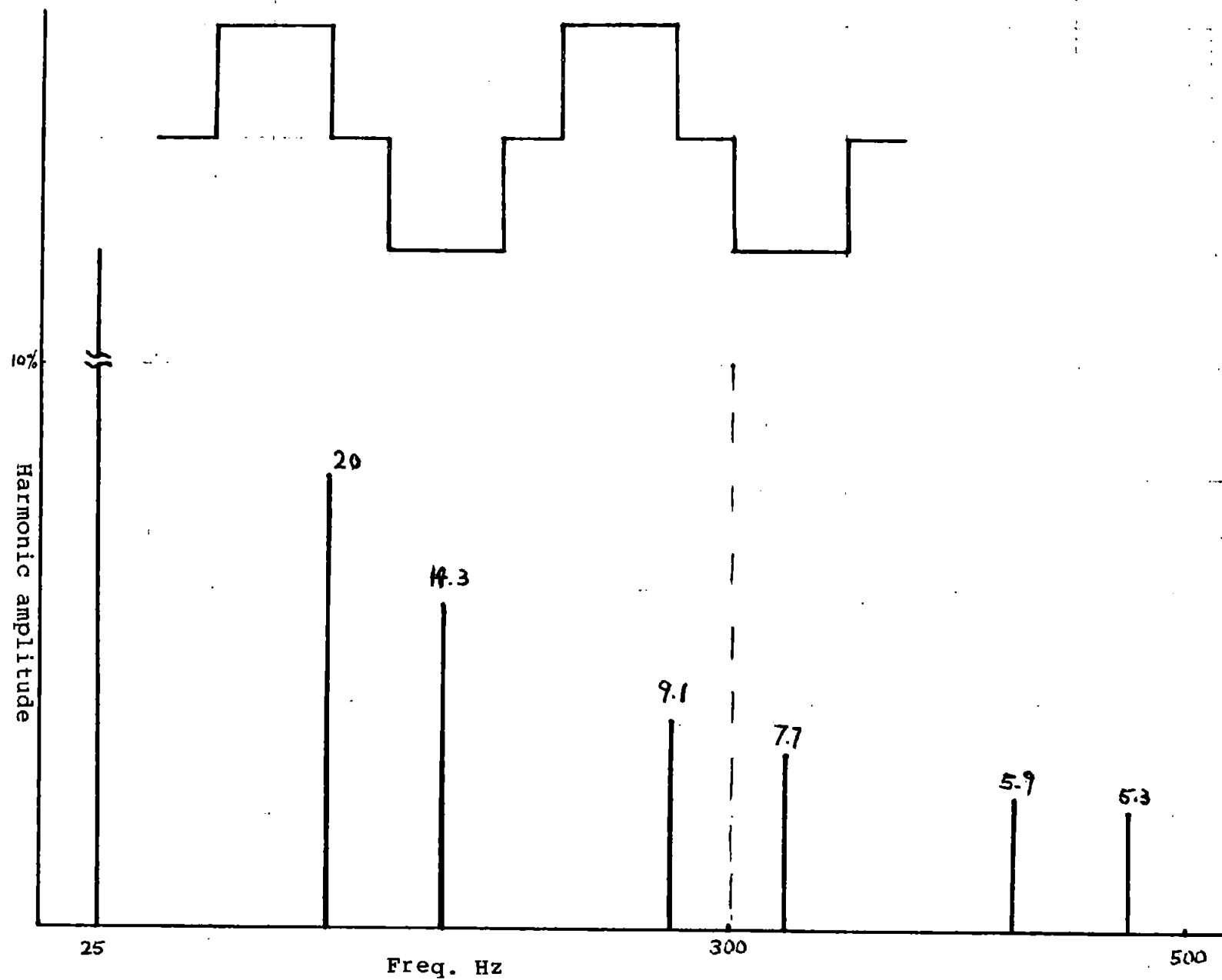


Fig. 5.2c $\phi = -\pi/12$

It can be seen that, in general, the spectrum is quite different from a phase controlled rectifier. The lowest unwanted harmonic component of a phase controlled converter with constant firing angle and hence steady output is at $P \cdot F_i$ - the average switching frequency (300Hz for the six pulse circuit), and all other unwanted components are at odd multiples of that frequency. For the cycloconverter in this case there are a few components below the average switching frequency $P \cdot F_i$, the lowest frequency component being the third harmonic of the wanted output component. The average switching frequency component itself is absent in this case and all harmonics are at frequencies of odd multiples of F_o from the average switching frequency. The largest harmonic component is around fifteen percent of the wanted component and at a frequency of F_o to $5F_o$ away from $P \cdot F_i$ i.e. $(P \cdot F_i \pm (2n+1) \cdot F_o)$ where $n=2$. Hence if a filter is required it should have significant attenuation at $P \cdot F_i$ i.e. 300Hz. A wide rejection band is required to attenuate the wide band structure of the unwanted component. This may be difficult to achieve as some unwanted components have frequencies close to the wanted component. Comparing with the quasi-square waveform in Fig.5.3, although the cycloconverter output has lower fifth and seventh harmonic, the third harmonic is present and therefore can not be judged as a better waveform.

Fig. 5.3 Quasi-square wave and its spectrum.



5.2.2 Computed results.

To compare the experimental results with the computed results, the computed harmonics are also plotted in Fig. 5.2 as dashed lines. They show good correlation with the experimental results. The largest deviation is around two and half percent of the wanted component. The discrepancy can be attributed to practical errors such as supply inductance, thyristor voltage drop, wave shaping of the snubber circuit and unbalanced three phase supply voltage not accounted for by the computer program. The two sets of results are also tabulated in Table 3 for more detailed comparison.

Table 3

TABLE OF EXPERIMENTAL AND COMPUTED RESULTS WHEN
THE PHASE BETWEEN COSINE TIMING AND REFERENCE WAVE IS CHANGING.

Numbers in bracket are experimental results.

Freq.	PHASE (π Radians)								
	1/12	1/16	1/24	1/48	0.0	1/48	1/24	1/16	1/12
75	7.7 (5.9)	4.9	3.4 (2.0)	3.5	4.2 (4.3)	4.4	4.9 (5.9)	4.0	2.8 (2.1)
125	11.0 (9.5)	9.8	9.8 (8.9)	9.2	9.7 (10.2)	10.9	11.6 (10.7)	11.2	9.2 (8.6)
175	9.2 (8.1)	10.1	11.3 (10.4)	11.7	12.5 (11.8)	13.2	13.8 (13.4)	14.6	14.8 (14.8)
225	2.0 (4.6)	4.2	5.6 (7.4)	7.2	8.1 (8.3)	8.4	7.2 (7.7)	7.5	7.9 (7.7)
275	6.1 (4.6)	5.5	5.1 (6.0)	3.6	2.5 (4.2)	4.7	5.4 (4.4)	6.8	6.8 (4.9)
325	4.0 (3.9)	4.7	5.7 (5.7)	5.4	4.3 (3.0)	3.4	5.2 (6.1)	6.5	8.0 (9.5)
375	6.6 (7.0)	4.3	6.4 (6.9)	9.4	10.8 (11.4)	8.5	5.4 (6.3)	4.4	5.1 (6.0)
425	11.0 (12.8)	13.2	11.2 (12.5)	7.7	8.7 (9.7)	12.8	11.7 (11.6)	6.9	7.6 (8.3)
475	8.0 (8.7)	6.0	8.8 (9.7)	11.3	8.0 (7.6)	4.6	9.6 (9.4)	11.2	7.5 (6.8)

5.2.3 The optimum value of phase between reference and cosine timing waves.

In principle there should be no essential difference in the output waveform of a cycloconverter with different values of phase angle between reference and cosine timing waves. The crossing points of the two waves are at different points on the reference wave cycle. The number of crossings is the same in each case. Therefore in each case the output is synthesising a different set of points on the reference wave and should all be equally good approximations of the reference wave. The T.H.D.'s for all cases are around 25 percent. However, to detect a possible best phase relationship, the computed harmonics are plotted against the phase angle between cosine timing and reference waves in Fig.5.4. The harmonics plotted are for discrete values of phase with straight lines joining the individual points. The 5th harmonic of 125Hz shows a relatively small change over the entire range. The 7th harmonic of 175Hz has a clearly increasing trend with increasing value of phase. However when the 7th is at its minimum the 3rd is at its maximum. This conflicting trend for the two adjacent harmonics means that there is no clear value of phase which produce the minimum amount of low order harmonic. At higher frequencies the same conflicting trend is also emerging. The harmonic with a frequency of 275Hz is minimum at a phase of zero

PHASE V HARMONIC AMP.

X - PHASE ($\pi/48$)

Y - % OF WANTED OUTPUT

F1 = 50 HZ , F0 = 25 HZ

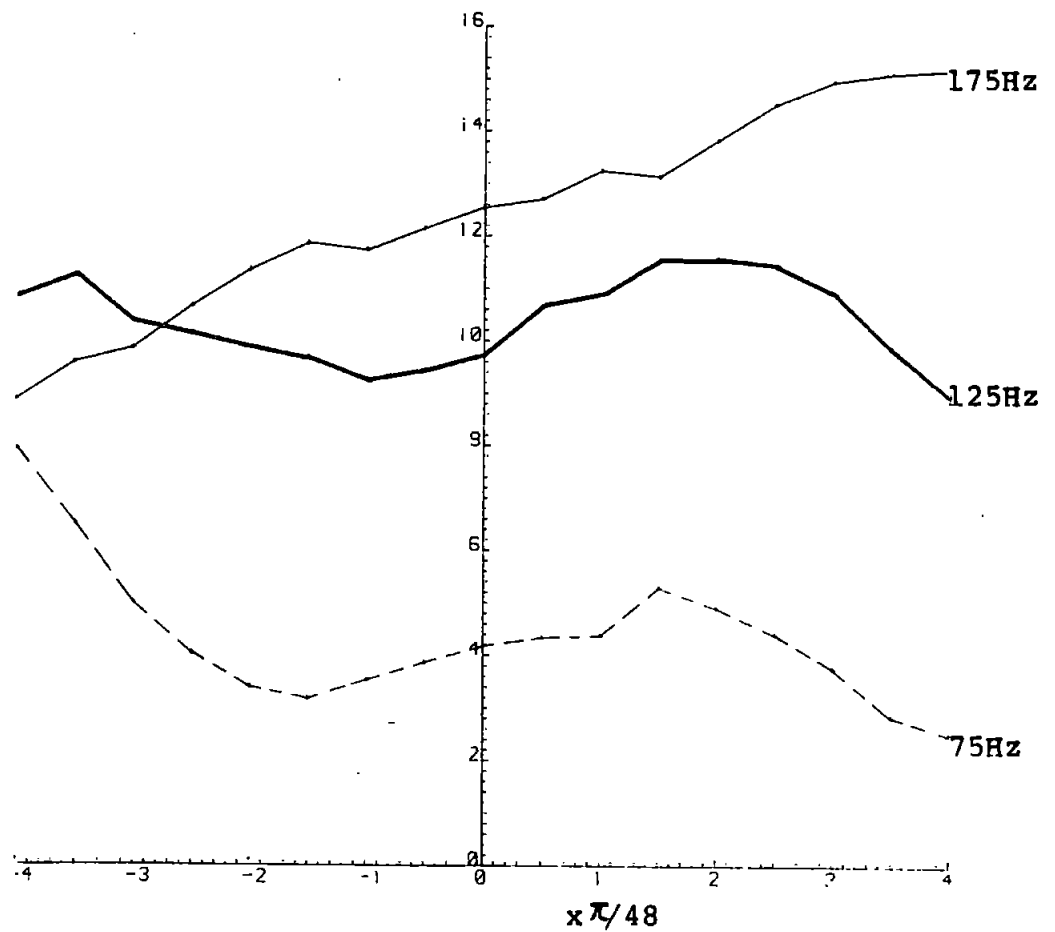


Fig. 5.4a Harmonic amplitude v phase shift
between cosine timing and reference wave.

PHASE V HARMONIC AMP.

X - PHASE (PI/48)

Y - % OF WANTED OUTPUT

F1 = 50 HZ . F0 = 25 HZ

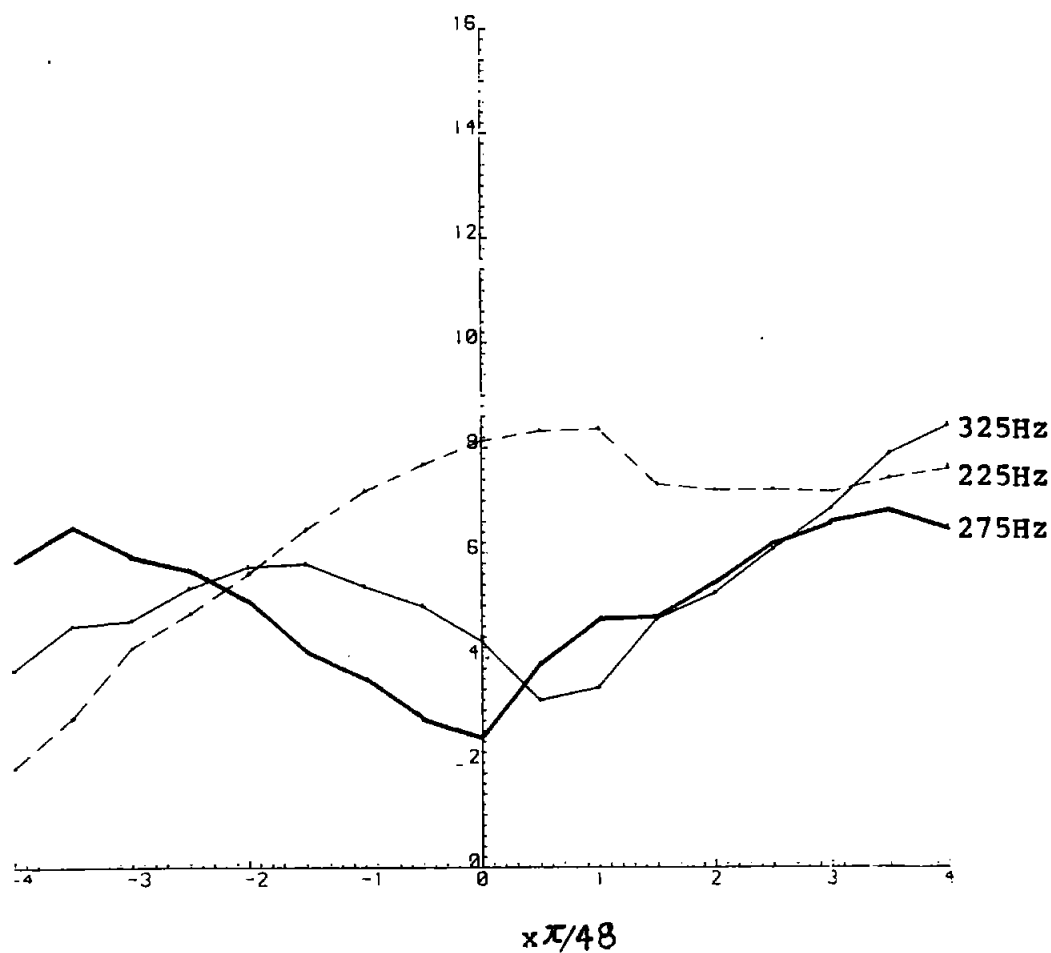


Fig. 5.4b Harmonic amplitude v phase shift
between cosine timing and reference wave.

PHASE V HARMONIC AMP.

X - PHASE ($P/48$)

Y - % OF WANTED OUTPUT

F1 = 50HZ , F0 = 25HZ

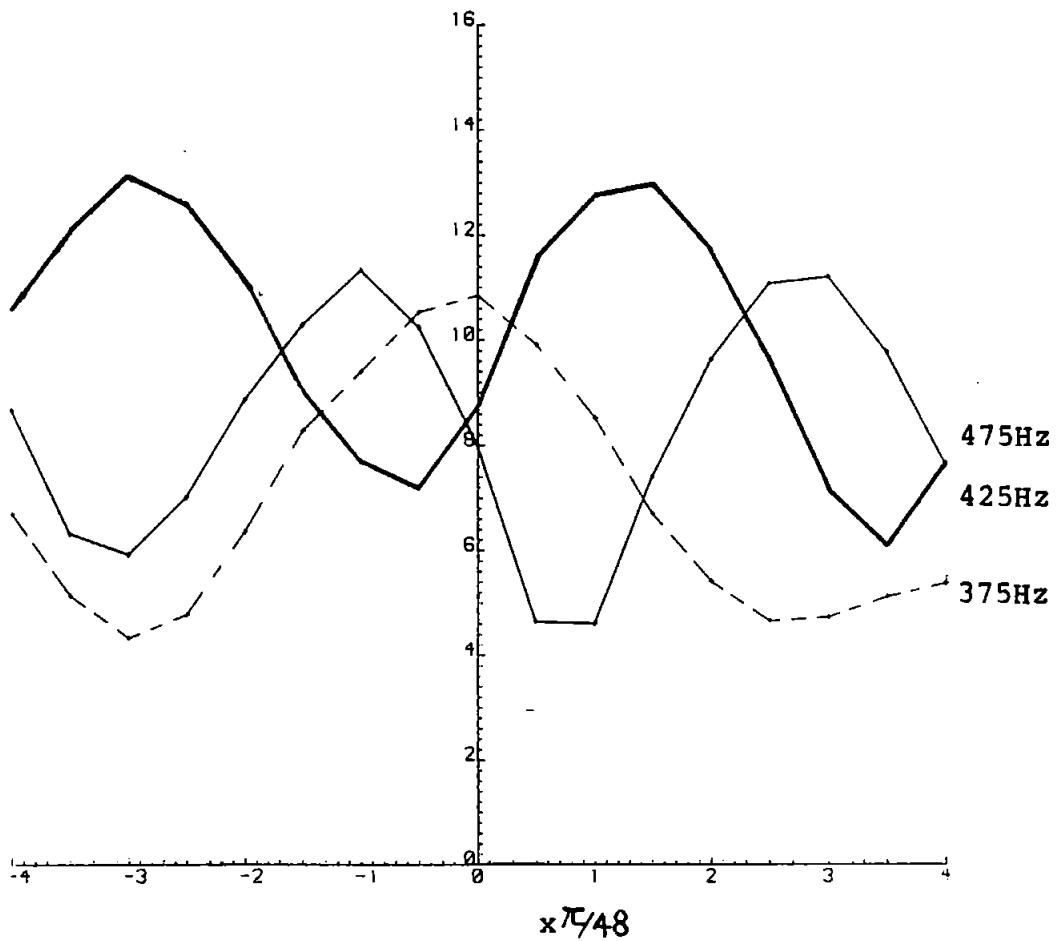


Fig. 5.4c Harmonic amplitude v phase shift
between cosine timing and reference wave.

while the 375Hz harmonic is at its maximum. The high order harmonics in Fig.5.4c behave quite differently and all show an oscillatory trend against phase. However the peaks are all at different values of phase.

As shown in chapter three, when a cosine timing wave crosses the reference wave at zero, this corresponds to either maximum or minimum value of phase angle depending on whether the crossing point is included. For this six pulse case the values of phase angle are either positive or negative $\pi/12$. As shown in Fig.5.4; 5.2a & c in general, the amplitude of a harmonic component at maximum and minimum value of phase is not the same. It means that there is a sudden jump in harmonic amplitude between the inclusion and exclusion of the crossing point when the reference wave crosses the timing wave at zero. At first sight, it is reasonable to assume a smooth transition as the value of phase is finite and cyclical. However, it should be noted that the magnitude the first output pulse tries to synthesise is zero when the crossing point is included but is larger if it is excluded. It is also interesting to note that, for this case - 25Hz output with a resistive load, the inter-group period is not the same for the cases of maximum and minimum value of phase. The inter-group dead period will be changed if the power factor of the load is different. But there are some other discontinuous current periods in the output waveform. The total number of current discontinuities

and their duration, including the inter-group period, for the case with resistive load is the same for maximum and minimum phase. Therefore in this case discontinuous current should not have a dominant effect upon harmonics and mask out the effect of changing phase.

5.2.4 The range of harmonic amplitudes specified by the cosinusoidal control method.

The effect of changing the phase between cosine timing and reference wave can be regarded as introducing an uncertainty to the output waveform and hence amplitude of harmonics. That is, a cycloconverter using the cosinusoidal control method does not produce a unique pattern of harmonics even for fixed output frequency with integer F_i/F_o ratio but instead a range is specified for each individual harmonic. The ranges in Fig.5.5 for the case of 25Hz output are obtained by plotting the maximum and minimum amplitude of each of the individual harmonics of the computed results against frequency. The range is generally between two to six percent of the wanted component. The amplitude with zero phase is also plotted as a dot on the line. This demonstrates that the case with zero phase is not necessarily the best because some of the harmonics produced are the largest over the range of phase, some are the smallest, as well as some in between the maximum and

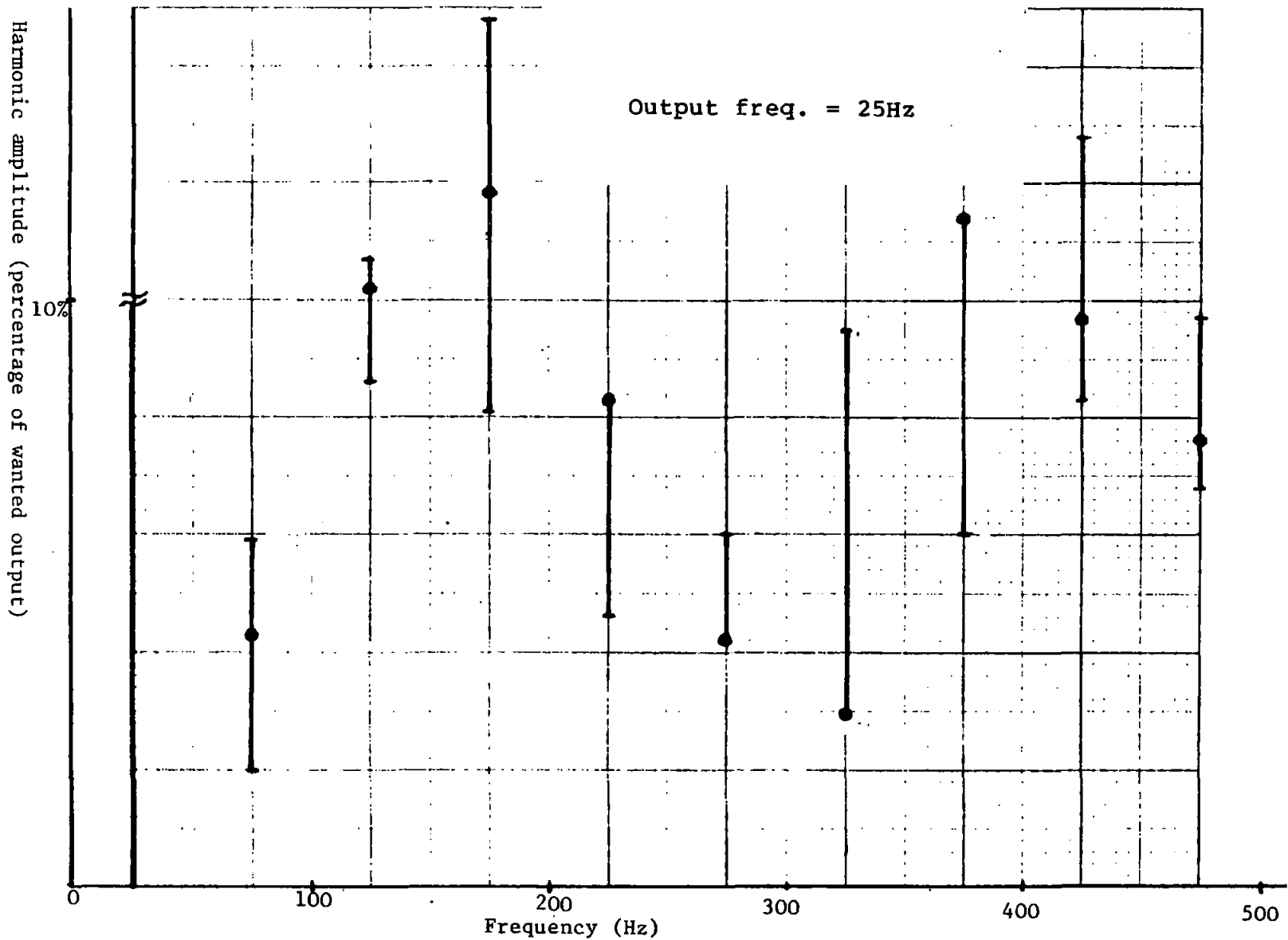


Fig. 5.5 Range of harmonic amplitudes specified by the cosinusoidal control method.

minimum. It should be noted that for an induction motor load there are other factors that influence the output voltage waveform. The mechanical loading on the motor would affect the power factor and amplitude of the current. Different power factors of the load would alter the group change-over point in the cycle. Therefore the group change-over dead period introduces a different amount of distortion to the voltage waveform. Also different current amplitudes would affect the duration and possibly the number of current discontinuities.

5.2.5 Results at other output frequencies.

All the results discussed so far are at an output frequency of 25Hz. Other frequencies studied are chosen to give a maximum range of phase. That means an integer ratio of $P \cdot F_i / F_o$. But this also implies the exclusion of sub-harmonics in the output spectrum. The actual frequencies includes 30, 25, 18.75, 15, and 12.5Hz. These correspond to a $P \cdot F_i / F_o$ ratio of 10, 12, 16, 20 and 24 respectively. The complete computed results are tabulated in Appendix VI.

At an output frequency of 12.5Hz, very similar results were obtained. The two largest components as shown in Fig.5.6 are the $(6F_i \pm 5F_o)$ component with nearly equal amplitudes of about eleven percent. However, with a lower output frequency, these are closer to the frequency of $6F_i$ and further away from the output frequency. That results in quite small low order harmonics :- 3rd, 5th etc. Hence the general conclusion that at low output frequency the waveform is better.

The uncertainty of harmonic amplitude introduced by the variation of the phase as shown in Fig.5.7a is again from two to six percent. The interesting feature here is that with a phase angle of zero, the low order harmonic produced is at or near to the minimum. However with a small change of around three percent, it is not necessarily a

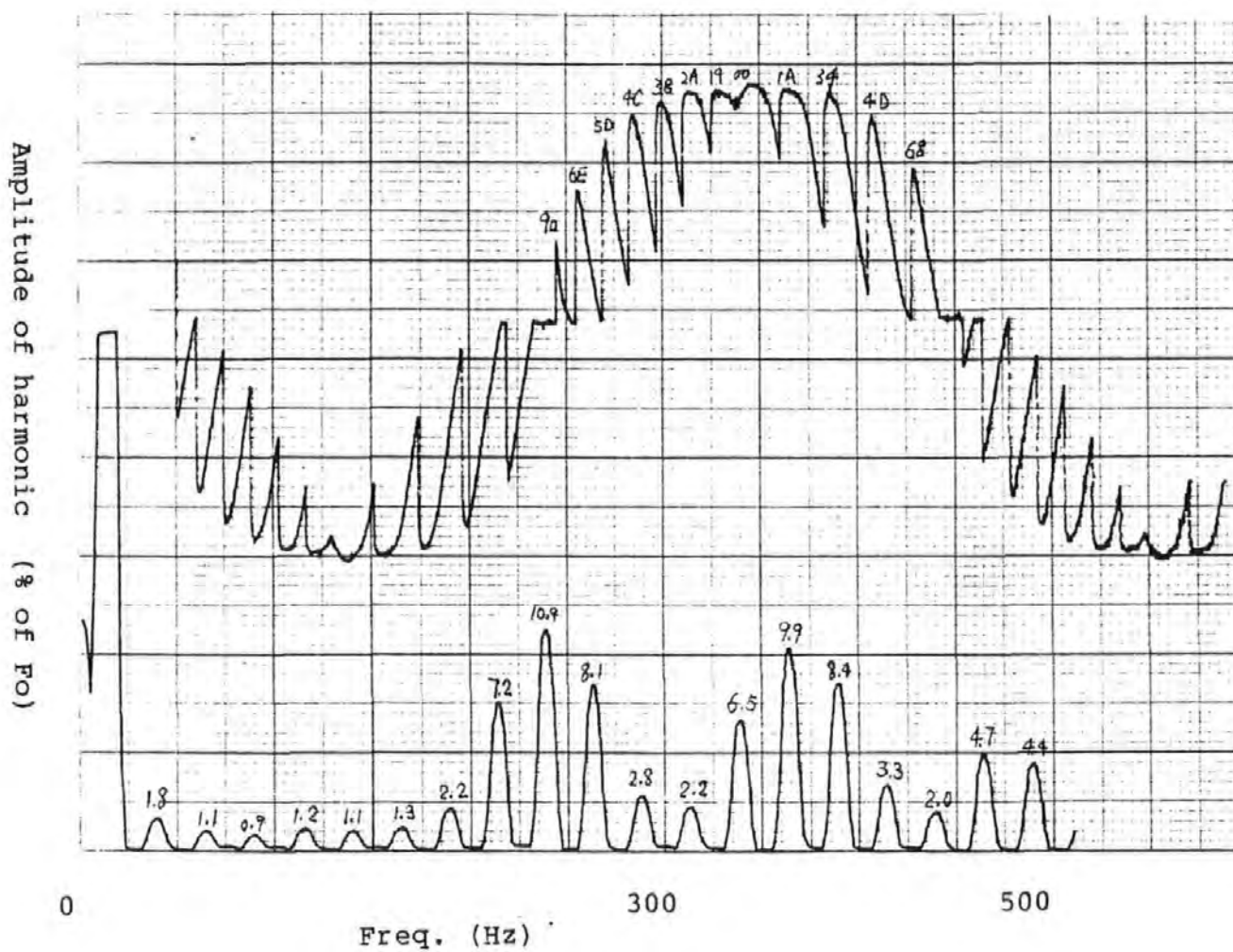


Fig. 5.6 12.5Hz output.

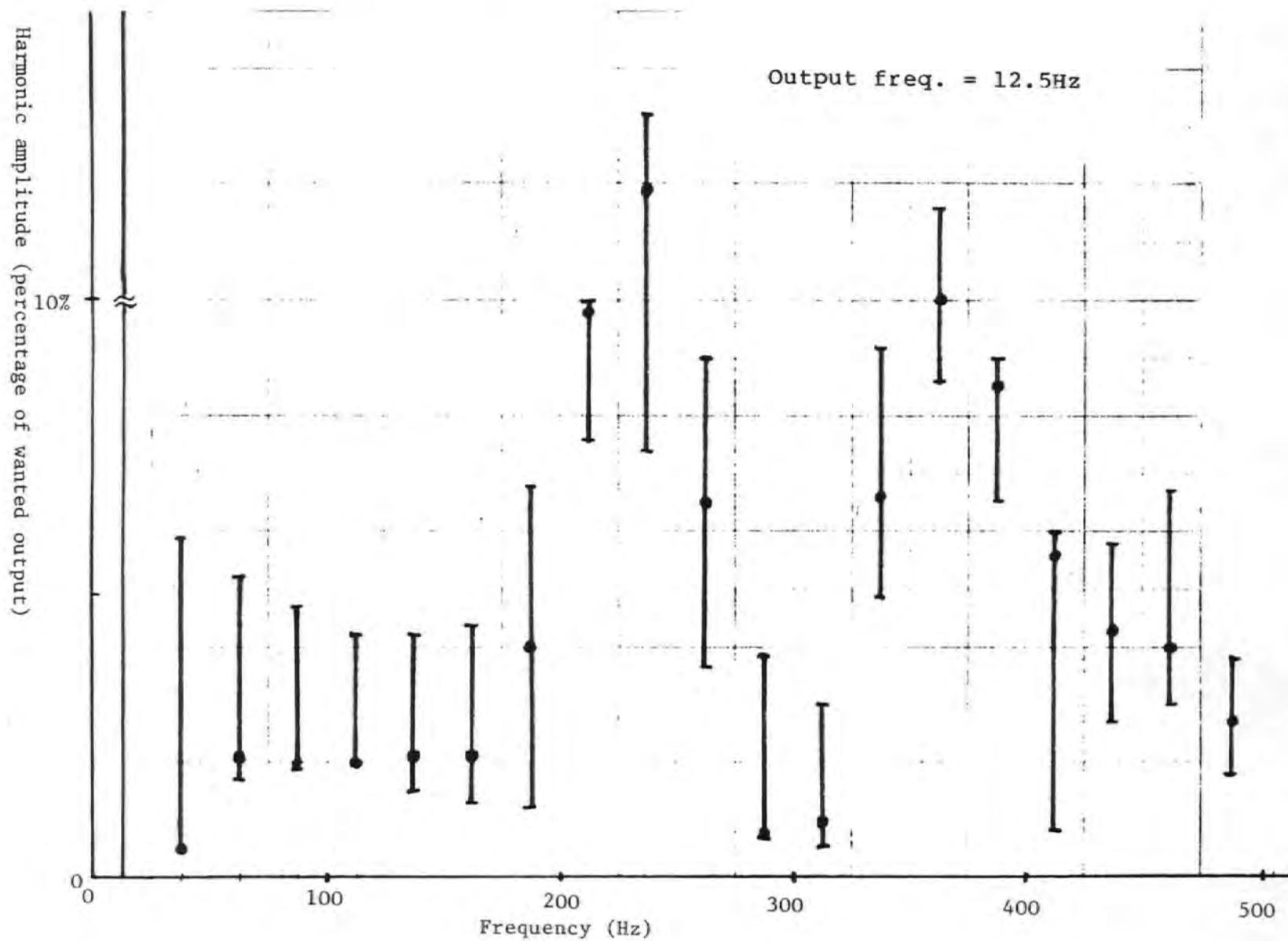


Fig. 5.7a Range of harmonic amplitudes specified by the cosinusoidal control method.

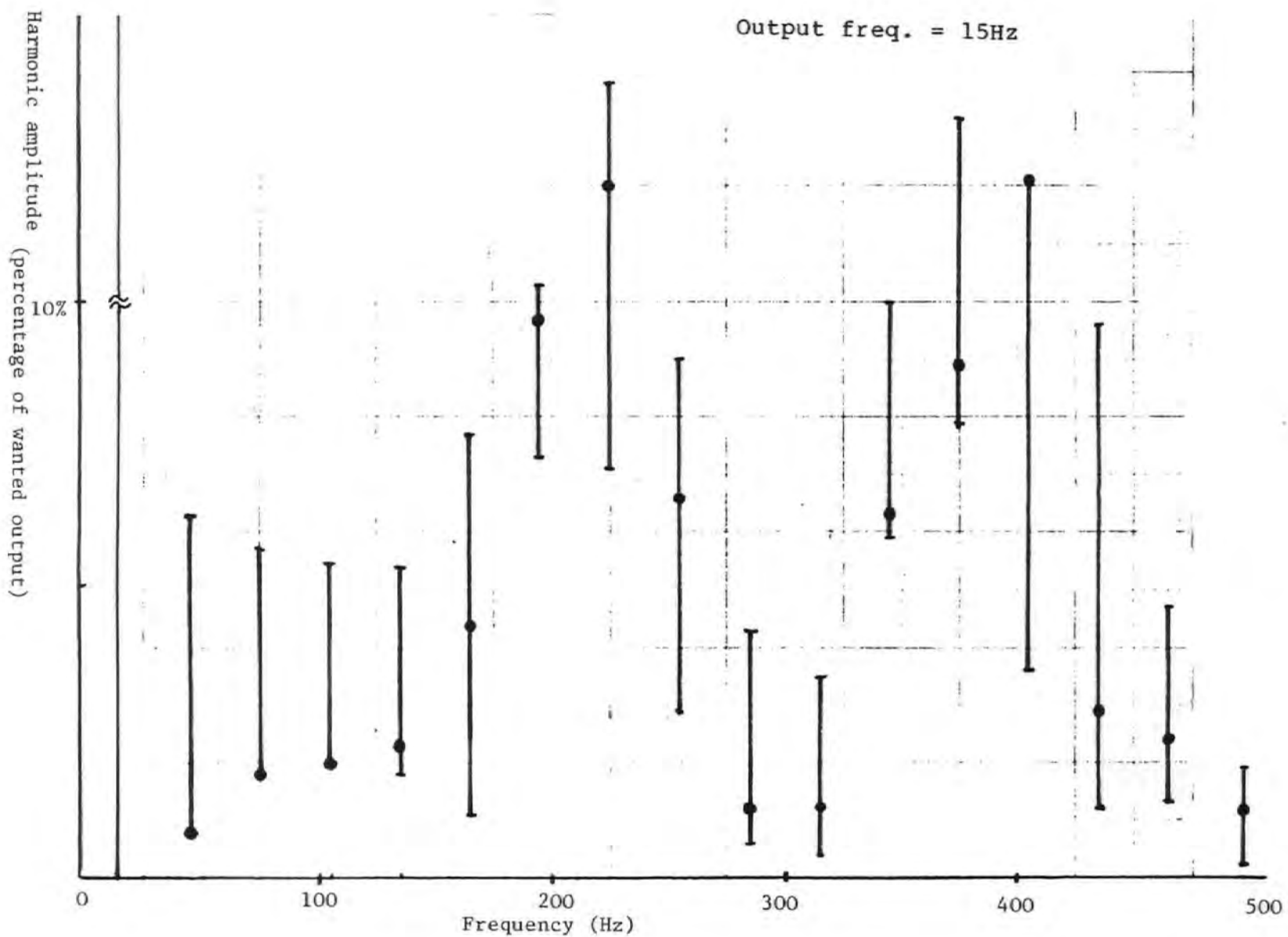


Fig. 5.7b Range of harmonic amplitudes specified by the cosinusoidal control method.

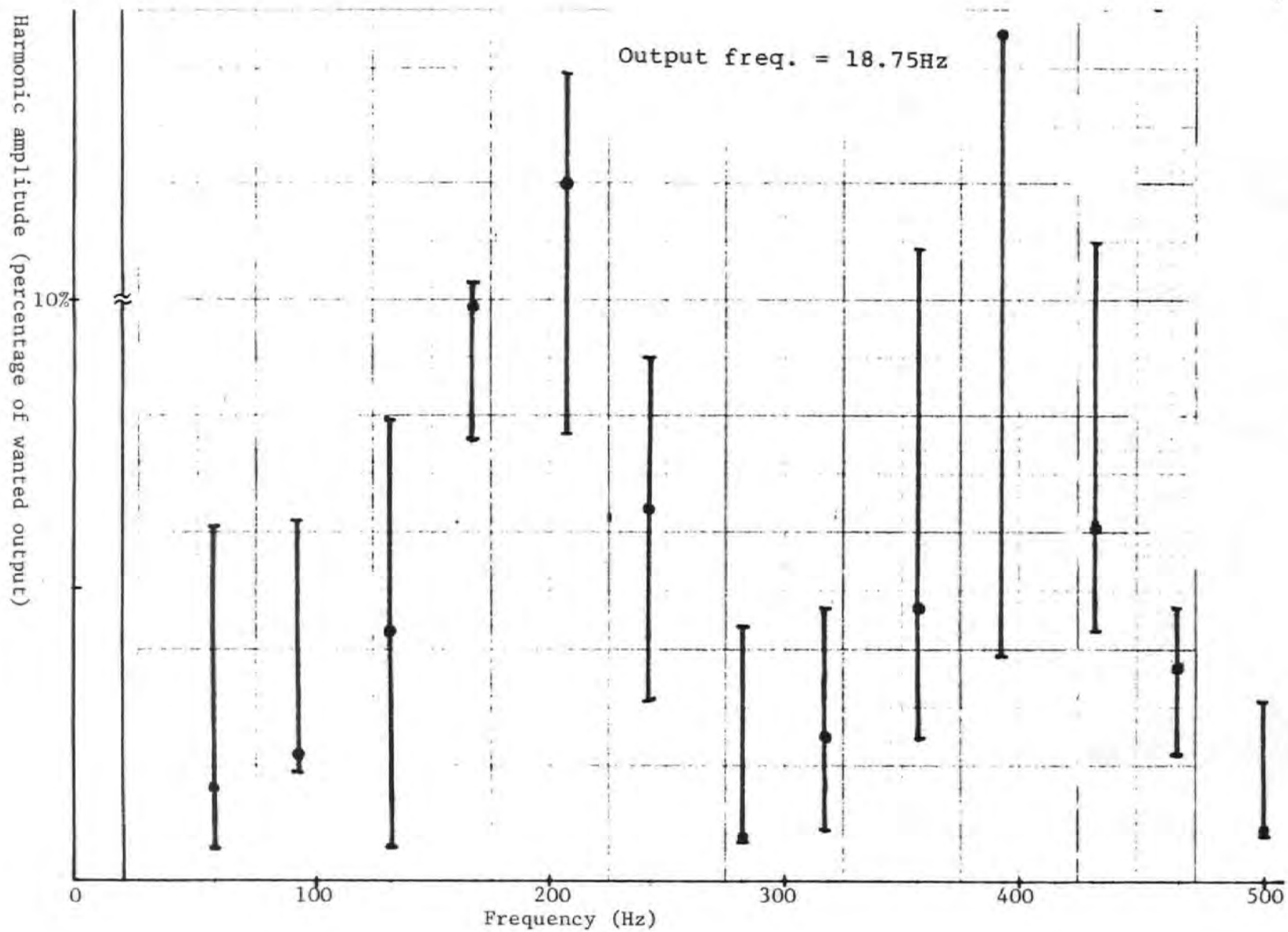


Fig. 5.7c Range of harmonic amplitudes specified by the cosinusoidal control method.

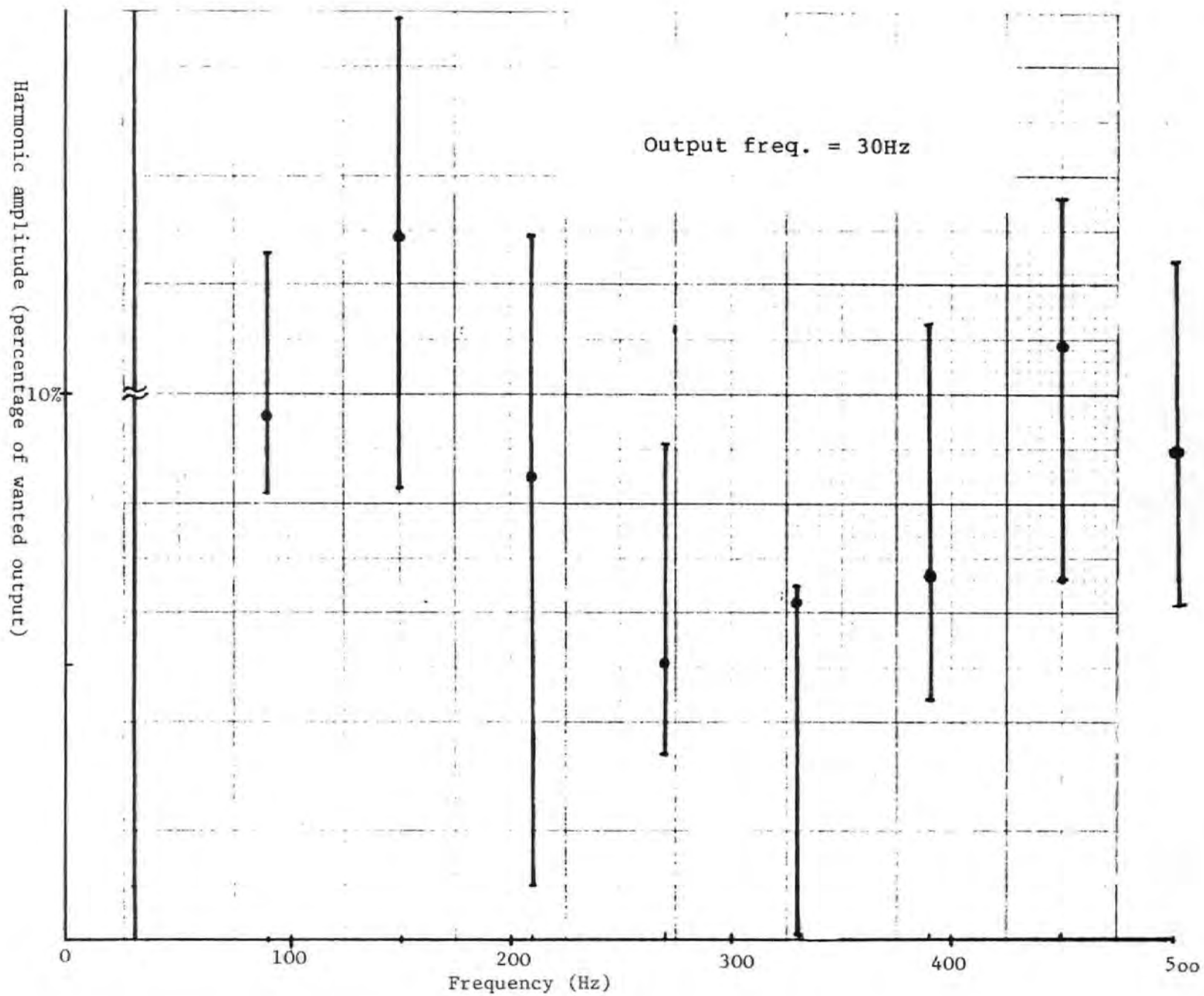


Fig. 5.7d Range of harmonic amplitudes specified by the cosinusoidal control method.

significant change as far as the load is concerned. As shown in Fig.5.7b, at an output frequency of 15Hz the low order harmonics are again at a minimum with a phase of zero.

However, at higher output frequencies of 18.75Hz, 25Hz and 30Hz, shown in Fig.5.7c, 5.5 and 5.7d respectively, there is no evidence of minimum low-order harmonic amplitudes for a phase of zero. The range of some of the harmonics is a little larger at around eight percent, for example, the 405Hz and 435Hz harmonic components at an output frequency of 15Hz. Also at output frequency of 30Hz, the amplitude of the low order harmonic is large. It is these large, low-order distortion components at high output frequency which determine the upper practical limit of the output frequency of the cycloconverter. The variation of phase seems unable to offer any improvement.

The data obtained indicate that there is no clear best value of phase for all the cases studied. It seems that there is a slight advantage in using phase of zero at low frequencies. However this would do little to extend the practical range of cycloconverter which is limited by the large harmonic components at high output frequency.

Although the frequencies studied only represent a few points on a continuous useable output frequency range of between 1 and 30Hz say, it is anticipated that more study at

other frequencies with a smaller range of phase is unlikely to point to a stronger evidence of best value of phase. Even if a preferable phase is identified for a number of output frequencies, in a practical system, the set speed may not remain constant for a sufficient long time for the controller to adjust the value of phase. For drive applications, even for fixed wanted speed, the required frequency for the cycloconverter is further complicated by the feed-back speed control loop. If the load on the induction motor is increased, it is necessary to increase the output frequency to compensate for the increase in slip. Further, the set or wanted frequency is not necessarily set at an integer ratio of $P \cdot F_i / F_o$. It may even be an irrational number which render the range of phase to zero.

5.2.6 Implication for comparing different control strategies.

The results obtained have important implications for the comparison of different control strategies. First, if reference and timing waves are used, regardless of their shape, the phase relationship between them must be considered because of the variation of harmonic amplitudes it can introduce. This applies to both analog control circuitry where real reference and timing waves are generated, and microprocessor-based control circuitry where these waves are just models for calculating the firing angles. Second, when comparing any control strategy with the traditional cosinusoidal control method, it is unlikely to be conclusive because of the uncertainty in the exact magnitude of harmonics. Reduction of amplitude in one of the harmonics together with increase in some other harmonics is not necessary an improvement. Unless every harmonic, or all harmonics within a range of frequency of importance, are all higher or lower than that produced by the cosinusoidal method, it is not certain. However, if Pelly's proof of the desirability of the cosinusoidal control method is taken into consideration, then it is reasonable to deduce that it is unlikely that any control strategies will produce an output waveform with harmonic amplitudes all lower than the minimum depicted by the cosinusoidal control method.

For example, consider cosinusoidal control with regular sampling. Fig 5.8 shows the time domain waveform and harmonic spectrum for the case with output frequency of 25Hz and phase of zero. The computed harmonics and experimental harmonics from the spectrum analyser are plotted as before. The notable feature here is the large third harmonic. At 14.5 percent, this is somewhat larger than the natural sampling case. However, not all the harmonics are higher than that with natural sampling and the relative importance of the third harmonic must be taken into account in comparing control strategies.

Some of the information was presented to The 20th Universities' Power Engineering Conference and are reproduced in Appendix VII.

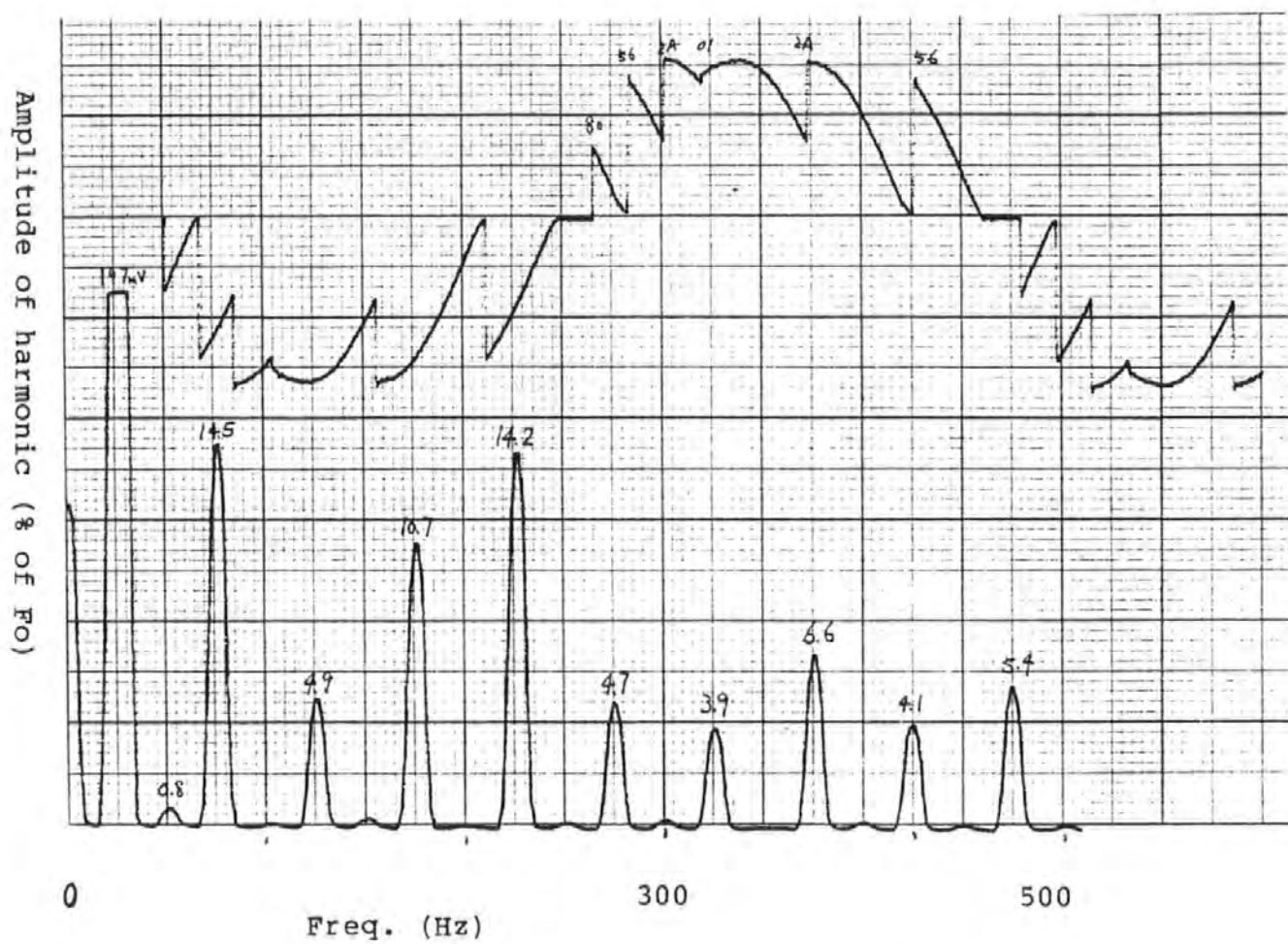
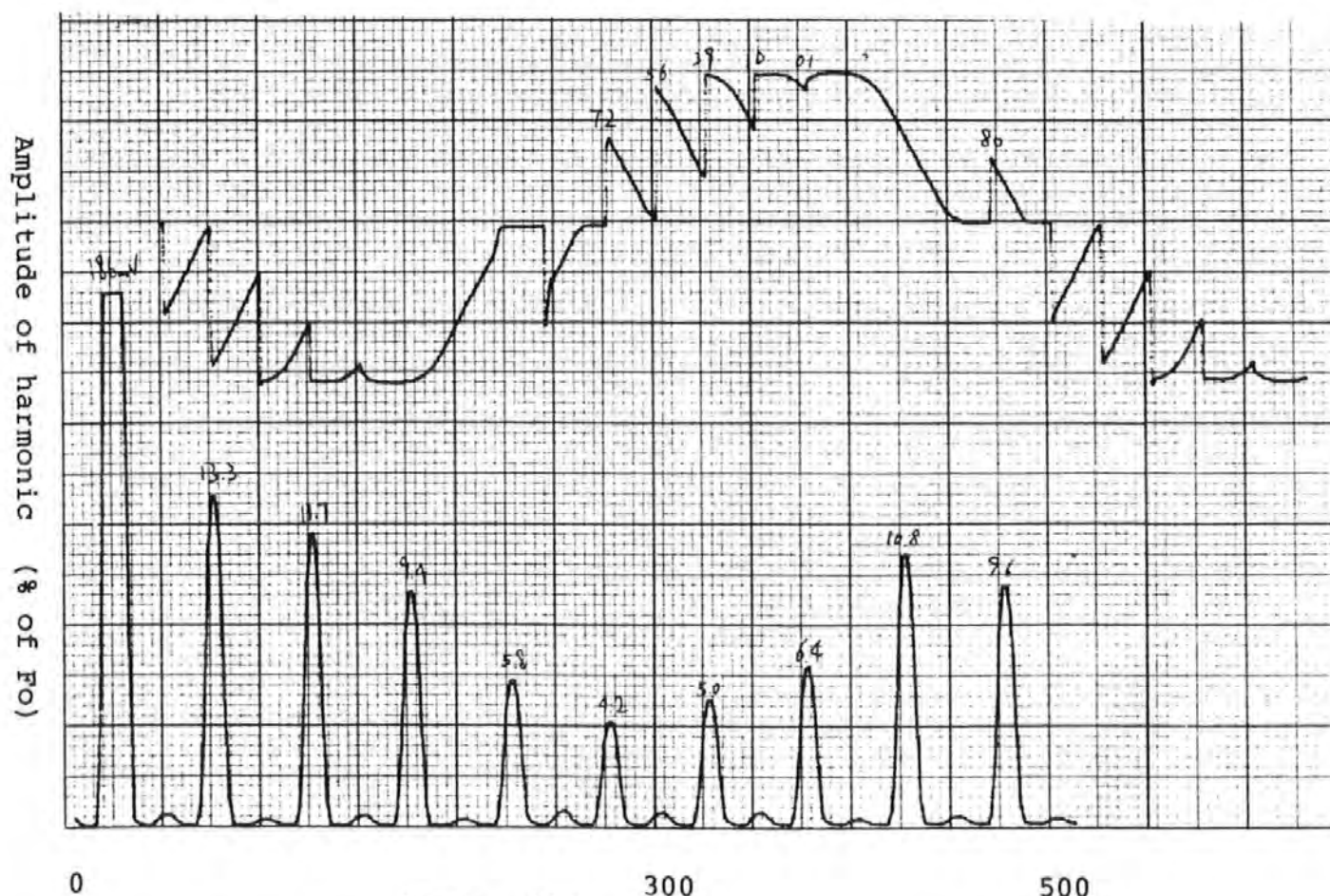
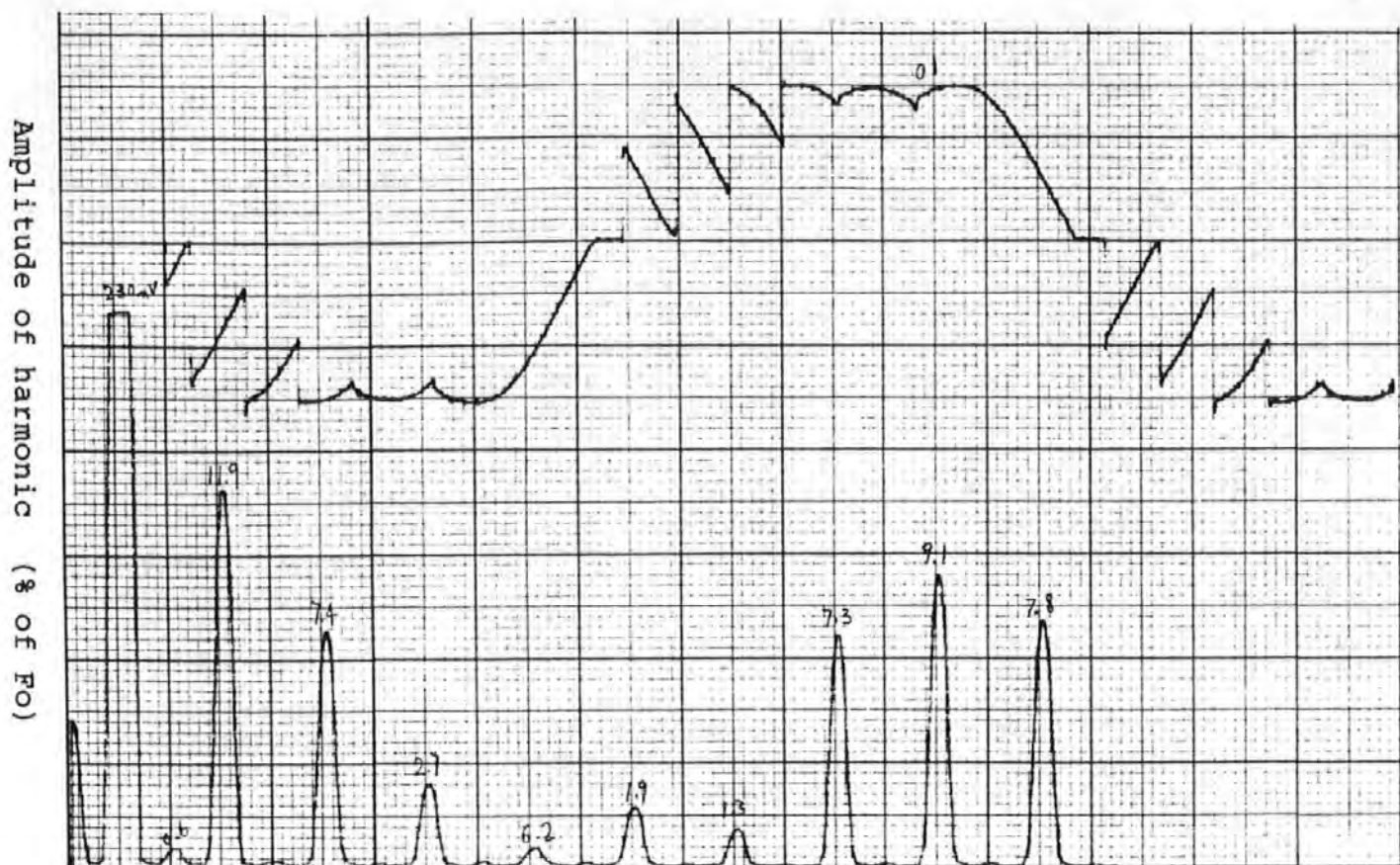


Fig. 5.8 Regular sampling with $F_0 = 25\text{Hz}$.

5.3 Variation of firing angles produced by the cosinusoidal control method.

In principle, changing one of the firing angles produces a different output waveform and causes a variation of harmonic amplitudes. This may offer a method of reducing the amplitude of some of the harmonics. This also offers a method of simulating the effect of control circuitry malfunctioning. This includes both the cases where the thyristor concerned is triggered too early and too late. As suggested in section 3.1, the last pulse in a half cycle can be the longest and gives more scope of investigation. Therefore the last pulse is selected to study the effect of its variation.

The output waveform using firing angles produced by the cosinusoidal control method and its spectrum is shown in Fig.5.2b. Then the firing angle of the last pulse in a half cycle was varied over the possible range of 0 to 120 degrees using a convenient increment step of 10 hexadecimal (approximately 7 degree). Any firing angles greater than 120 degrees would produce no output current as the load is resistive. Two other waveforms and their associated spectrum are shown in Fig. 5.9. It shows that all the firing angles are the same except for the last one in a half cycle.



0 300 500

Freq. (Hz)

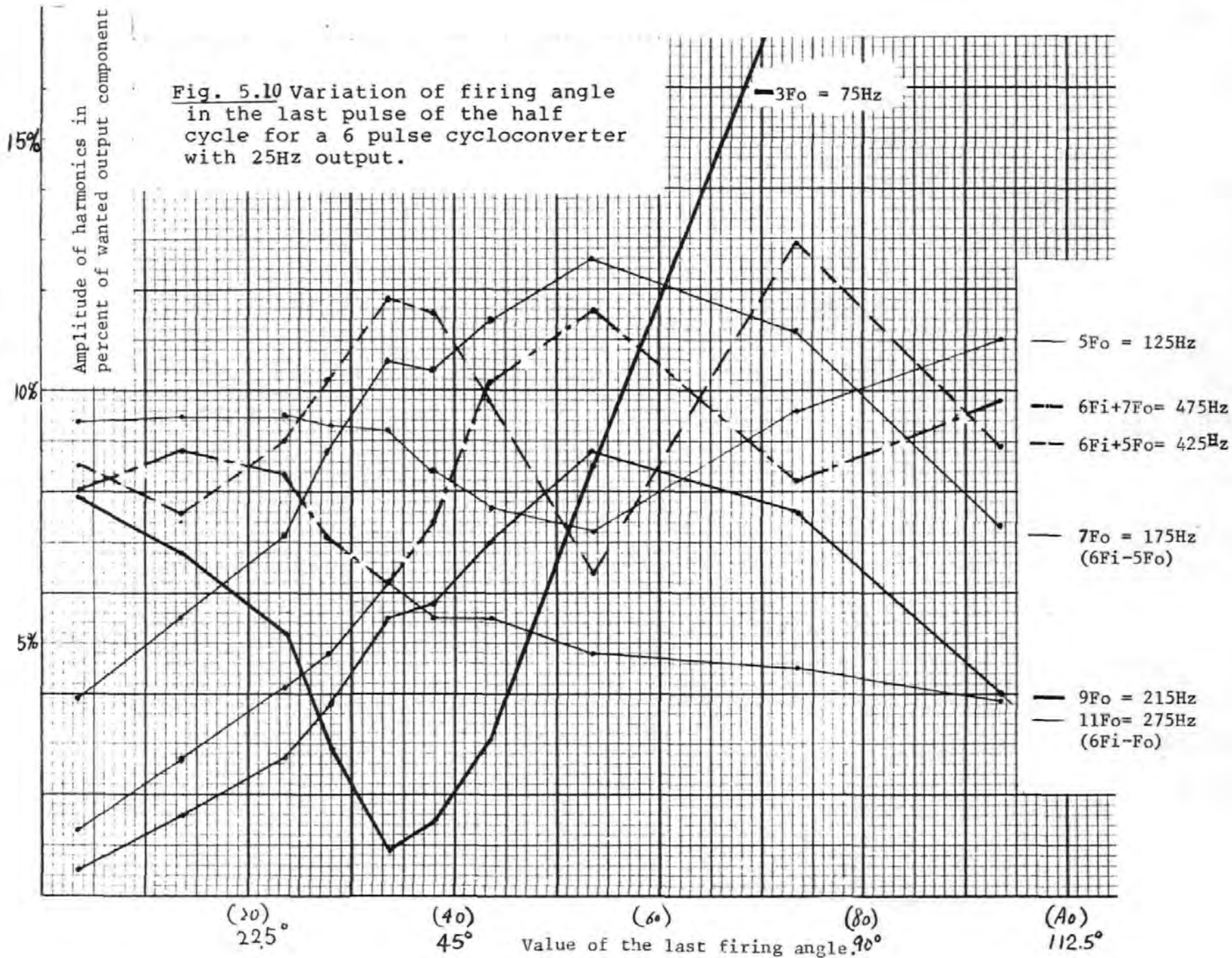
Fig. 5.9 25Hz output with different last pulse.

The frequencies of the fundamental and all harmonics of the output waveform remain the same. This is to be expected because the period of the waveform is unchanged. The amplitude variation of some of the harmonics over the range of firing angles is plotted in Fig.5.10 The value of firing angles is also shown in degrees. The third harmonic shows a clear minimum when the firing angle is approximately 35 degrees. This is close to the value obtained by the cosinusoidal control method. The higher order harmonics shows a different pattern of fluctuation. There is no clear minimum nor an overall trend.

At a large firing angle or when the firing pulse is missing i.e. firing angle larger than 120 degrees for the resistive load, there is a large increase of the third harmonic from less than two percent to more than fifteen percent.

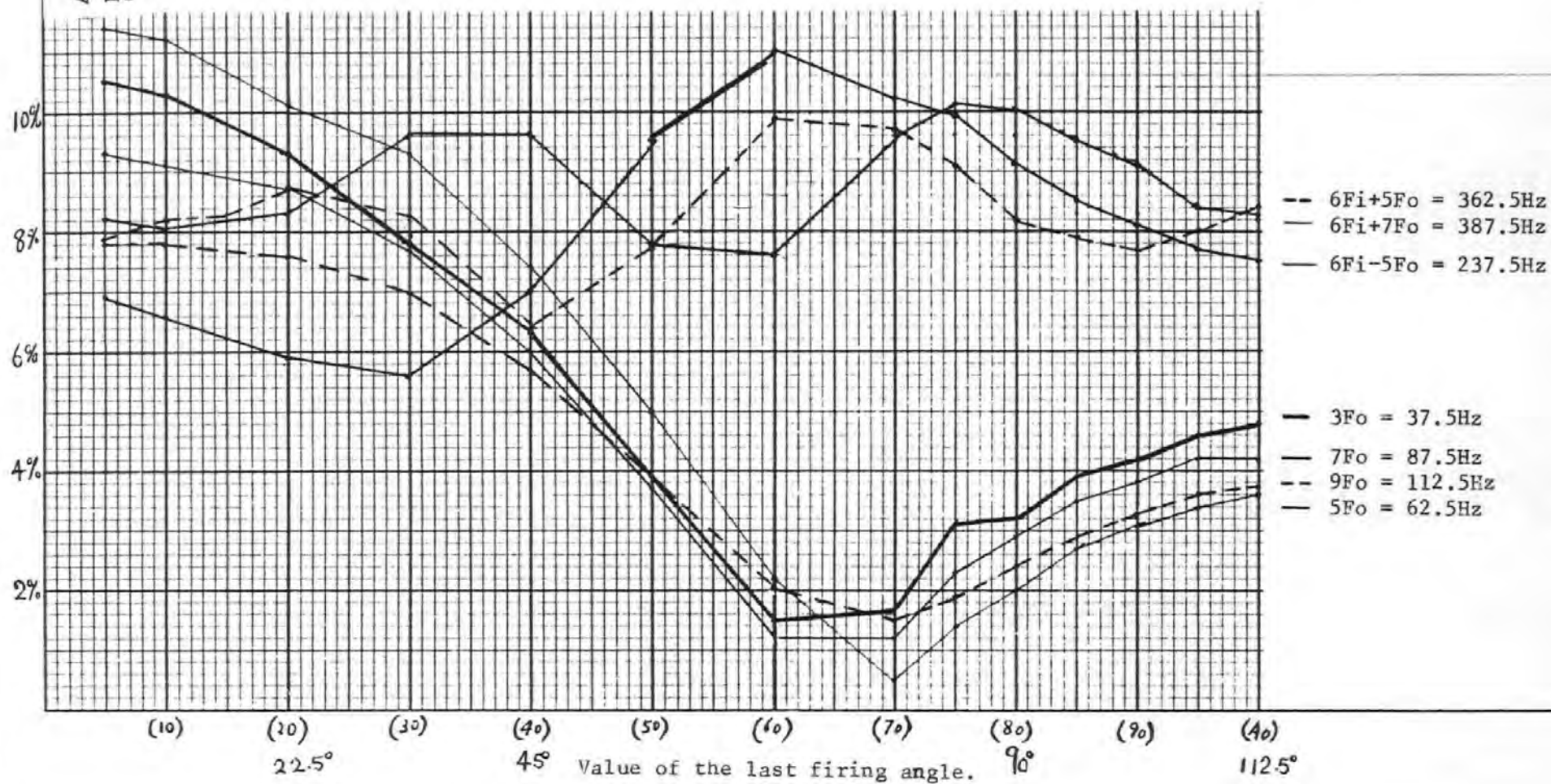
At a lower output frequency of 12.5Hz, there are more pulses in a half cycle. Therefore changing one of the firing angles is expected to have less effect upon the output waveform and harmonic content. Fig.5.11 is the plot of the last firing angle against amplitude of various harmonics at an output frequency of 12.5Hz. The lower order harmonics (3rd, 5th, 7th and 9th) shows a clear minimum around a firing angle of 80 degrees. This is also close to

Fig. 5.10 Variation of firing angle in the last pulse of the half cycle for a 6 pulse cycloconverter with 25Hz output.



Amplitude of harmonics in
percent of wanted output
component.

Fig. 5.11 Variation of firing angle
in the last pulse of the half
cycle for a 6 pulse cycloconverter
with 12.5Hz output.



the value obtained by the cosinusoidal control method. The high order harmonics also show no trend as with 25Hz output. Assuming the higher order harmonics are more easily filtered out and have less harmful effects, it is not likely to be of advantage to change firing angles produced by the cosinusoidal control method at low output frequency.

5.4 Operation with high output frequencies.

The theoretical maximum output frequency of a naturally commutated cycloconverter with balanced output i.e. with both positive and negative output current, was shown by McMurray (31) (P.62) to be $P \cdot F_i / 2$. It follows that for a six pulse cycloconverter, the maximum output frequency is 150Hz. Although the condition is similar to the famous Nyquist limit in sampled signals it has a different origin. Its limit is due to the turn off restriction of thyristors and the maximum number of current zeros within an input cycle period which is equal to the pulse number of the converter. At the point of group change-over, the pair of thyristors with the largest firing angle is chosen to give the shortest conduction period before a current zero can occur. Only one pair of thyristors is fired in each output half-cycle and the conducting thyristors are allowed to turn off at the earliest instant i.e. the first zero current point. Then a group change-over is performed again after each pulse of current. That is each output cycle contains two pulses of current, one positive and one negative. This is feasible with the microprocessor controlled experimental cycloconverter. An output waveform with a frequency of 150 Hz together with its spectrum is shown in Fig.5.12.

As there is only one pulse in every half cycle, the only control over the waveform is to increase the firing angle. The possible range of firing angle in this case is between 60 and 120 degrees. However increasing the firing angle would reduce the conduction period and hence the wanted output component. The amplitudes of the harmonics are high compared with the harmonics of lower output frequencies e.g. 28 per cent of third harmonic and the maximum wanted output component is much lower than the input. This makes such a waveform unsuitable for most practical applications.

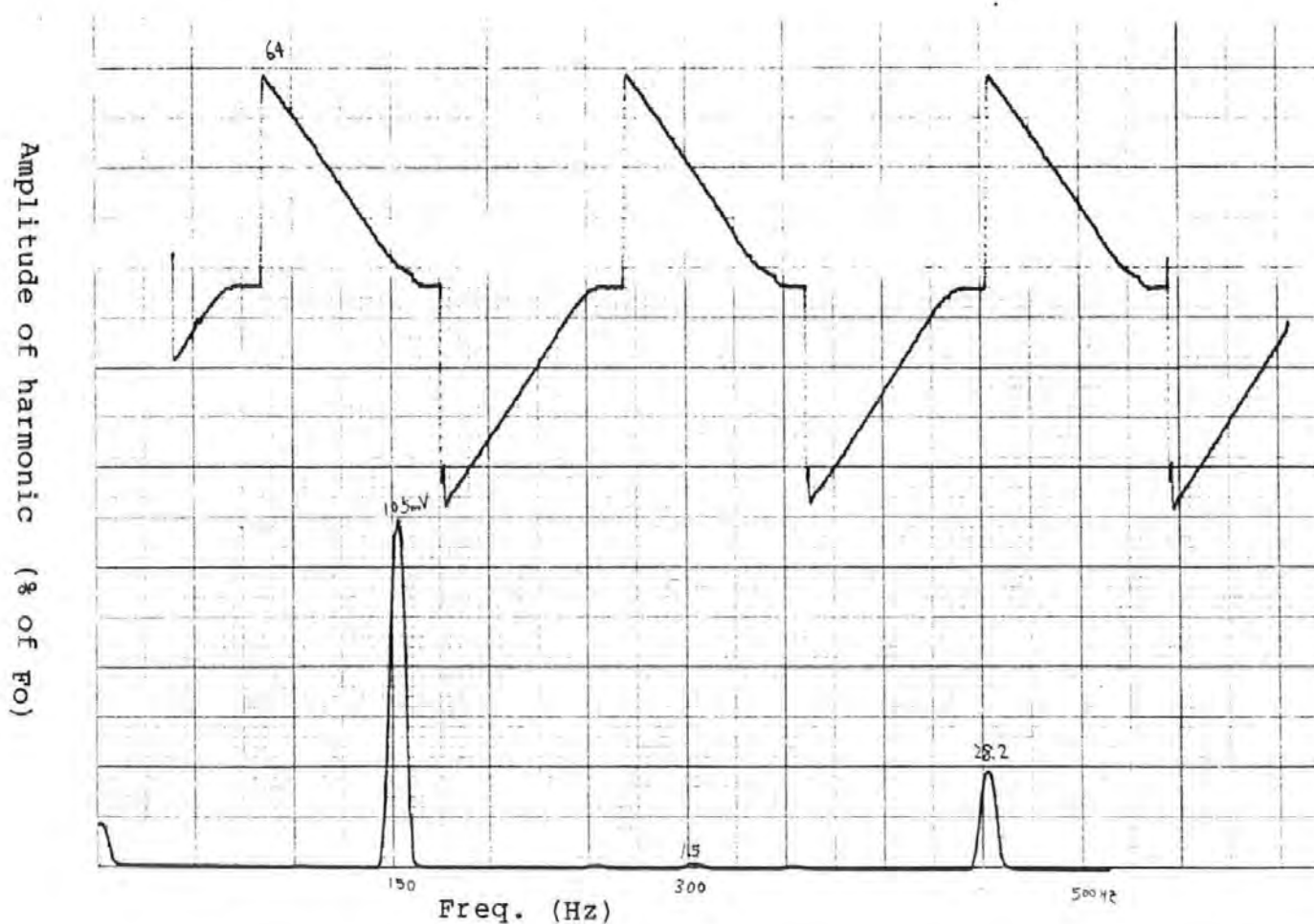


Fig. 5.12 150Hz output.

It was also noted by McMurray that to produce a balanced three phase output without a special phase-shifted supply to the three groups of converter, the maximum output frequency is further reduced by a factor of three. That is each output phase uses a different possible turn off point sequentially. At that output frequency of 50 Hz, there are three controllable switching instants. This should result in a more sinusoidal output. A number of combinations have been tried experimentally. The output waveform as shown in Fig. 5.13 uses firing angles produced by the cosinusoidal control method with zero phase between cosine timing and reference wave. It contains low order harmonics with small amplitudes.

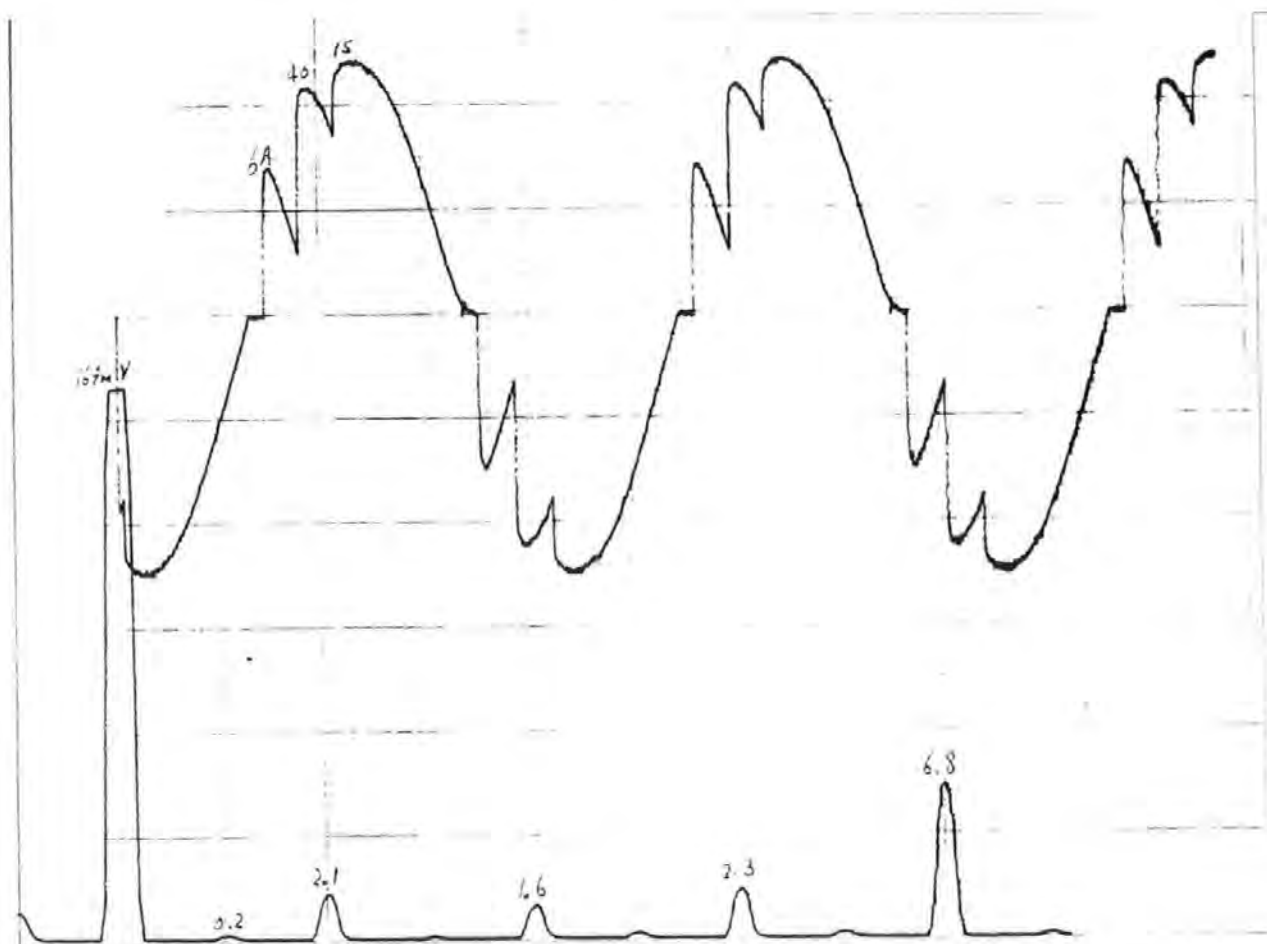


Fig. 5.13 50Hz output.

Only the ninth harmonic at a frequency of 450 Hz, within the range of zero to 500 Hz, has an amplitude greater than 5 per cent. This shows the possible scope of improvement at other output frequencies.

The case of 50 Hz output can be considered as a special case because the load can be connected directly to one input phase by a by-pass switch and with no current going through any thyristors. However, it can be envisaged that if the last conducting group has already stopped conducting, by firing the thyristor earlier, at the time when the incoming phase voltage is zero, the output becomes a pure sine wave. This is possible because before a group starts conducting there is no need to transfer current from one thyristor to the next or to commute. Any thyristor is free to start conducting if the thyristor is forward biased. Therefore, there is one chance to start conduction before the start of its associated cosine timing wave period in any half cycle of the output. This implies firing the thyristor before the normal cosine timing period which requires the control circuit to perform the extra function of selecting the thyristor before its cosine timing wave starts. For the six pulse case, altogether there are three combination of pairs of thyristors which would produce an output with the correct polarity after a group change-over. The possible output waveform in Fig.1.4 (P.14) should therefore alter to that shown in Fig.5.14. The firing angle

in such a case would be negative.

Starting a thyristor earlier could be a useful area of study to improve the output waveform at high output frequency and extend the operating range of cycloconverter. Using a microprocessor for the control, the output frequency can be controlled so that it changes continuously from say, 1 to 20 Hz and then in steps to higher frequencies. The step change of speed in the motor due to step change of frequency can be smoothed out by voltage control at these fixed frequencies.

But if the current is required to pass through the thyristors i.e. no by-pass switches, and conduction is restricted to the period of the corresponding cosine timing wave i.e. without extra function in the control circuit to start conduction early, then start of conduction is possible at the latter part of the cycle as shown in Fig.5.15. This is achieved experimentally by having the first two firing angles delayed so that all three thyristors fire together. The output is the same as that of voltage regulation with phase control and is not a pure sine wave but with part of the cycle remaining at zero. The output waveform in Fig.5.13 is much better. However an inductive load will improve the voltage waveform. This investigation also shows that the cosinusoidal control method is desirable.

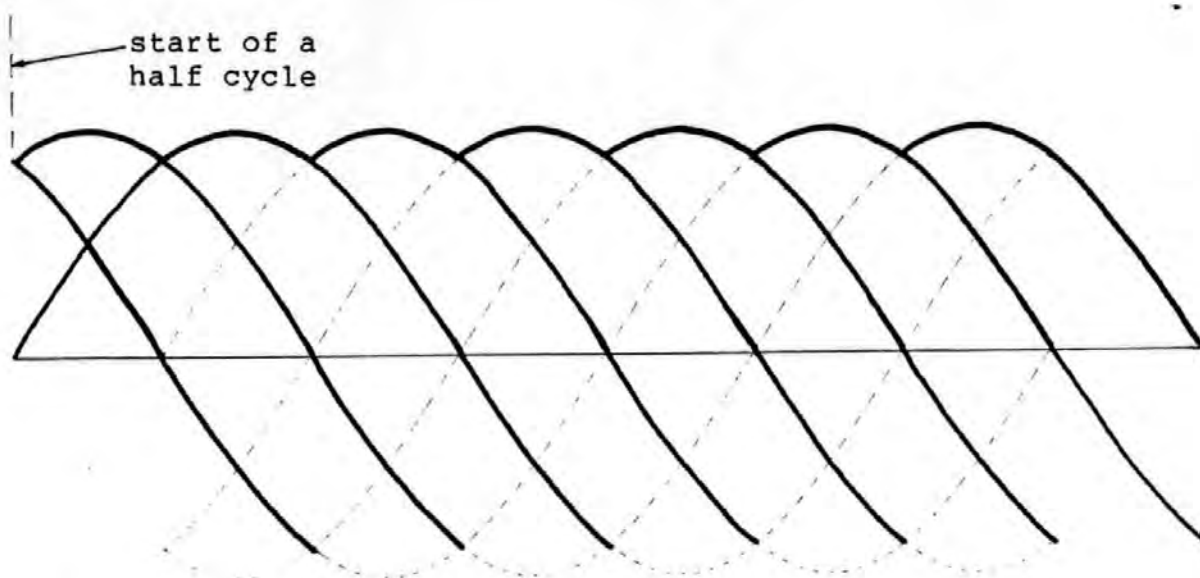


Fig. 5.14 Possible output voltage of a 6 pulse cycloconverter with early start.

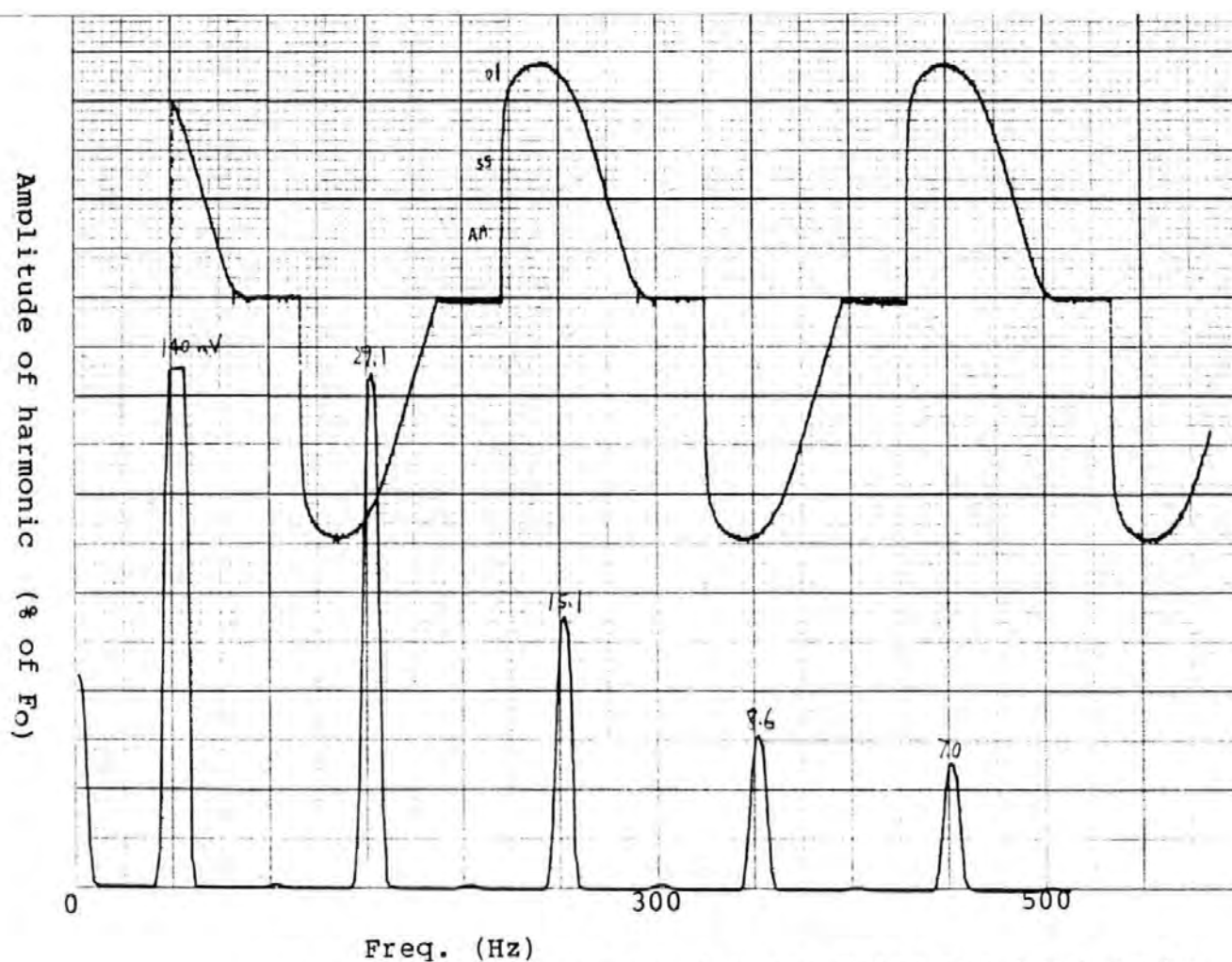


Fig. 5.15 50Hz output with one pulse per half cycle.

5.5 Operation at half modulation depth.

For motor drive applications, it is required to keep the flux of the motor constant. That is, when running at reduced speed, the output voltage of a cycloconverter must be reduced to avoid saturation in the magnitude component of the motor. If the resistance of the winding is neglected, then the voltage to frequency ratio should remain constant. For a standard 50Hz motor to run at half speed, a 25Hz output with modulation depth of 0.5 is required. Fig.5.16 shows such a waveform using firing angles produced by the cosinusoidal control method. It can be seen that the harmonic amplitudes normalised to the wanted output is much larger than that with unity modulation. However the absolute magnitude is about the same with unity modulation.

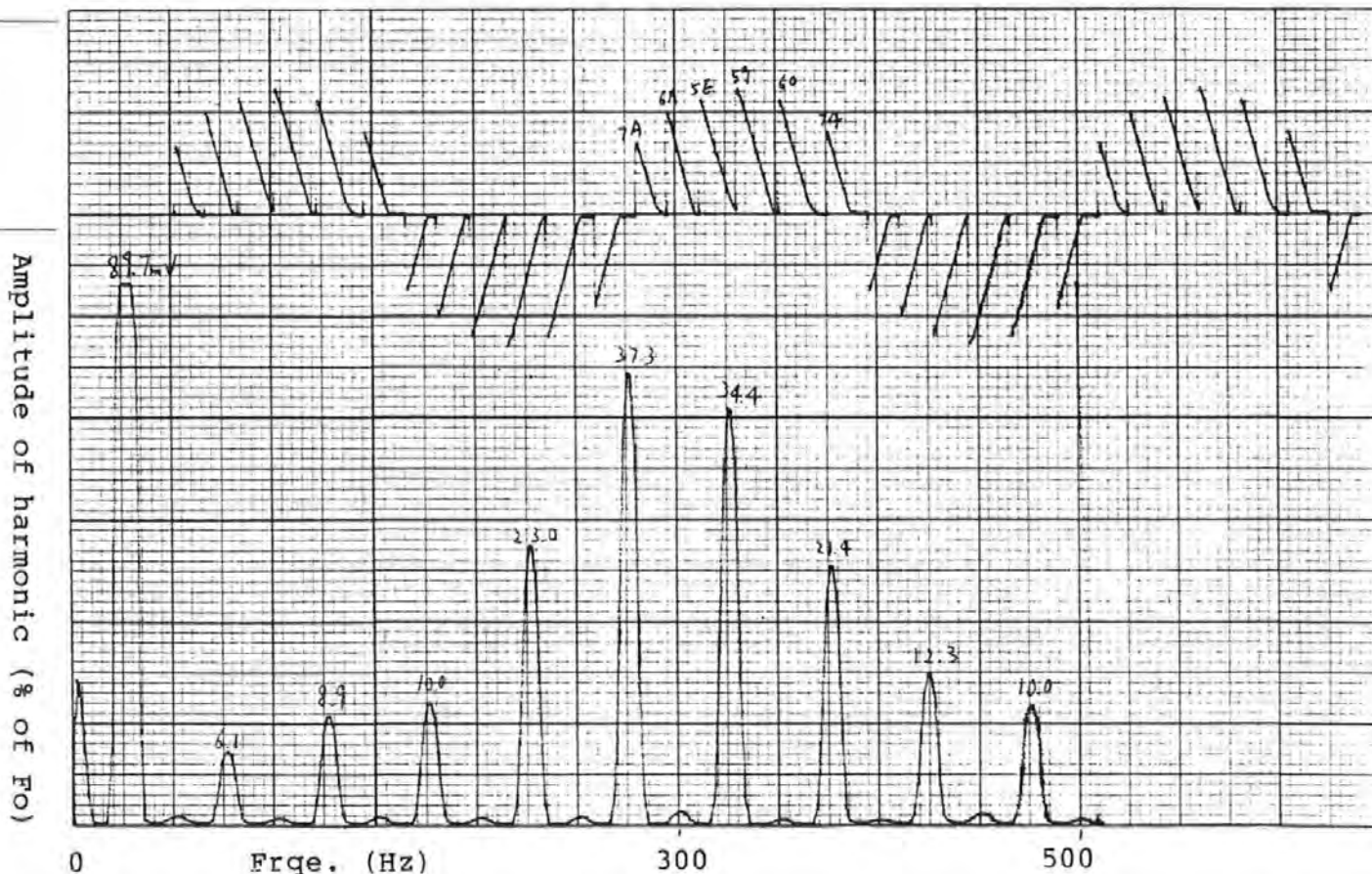


Fig. 5.16 25Hz output with half modulation depth.

CHAPTER SIX
CONCLUSIONS.

6 CONCLUSIONS.

The single stage frequency conversion process of cycloconversion by connecting the output to each input phase of a poly-phase supply in turn is an elegant concept. However the large number of thyristors required and the harmonics in the output waveform is unavoidable. The complex control circuit can be simplified and the control function enhanced by using a microprocessor for the control.

It has been established that the r.m.s. distortion is minimum with the cosinusoidal control method. Further, the T.H.D. of each family of harmonics is equivalent to the phase controlled converter with the same pulse number. However, for drive applications, it is the low order harmonics with frequency close to the wanted component that has the most adverse effect, the exact effect being dependent on the motor concerned. The amplitudes of the lower order harmonics are therefore of interest.

An experimental six pulse, three phase to single phase cycloconverter has been constructed. The effect of using a P.L.L. for the synchronisation and timing are as follows. There is no compensation of output voltage or frequency for a change of input voltage or frequency but the wanted output component and its harmonics remain at the same level relative to the input amplitude. This situation is more suitable for harmonic content investigation compared with other control circuits which produce fixed output voltage and frequency but varying harmonic amplitude.

It was demonstrated that in the cosinusoidal control method the phase relationship between the cosine timing and reference wave affect the output waveform and hence the harmonic amplitudes. The T.H.D. up to 500Hz remains fairly constant at the different output frequencies investigated, but each individual harmonic can change by a significant amount. There is no clear best value of phase and in practice the output may not be able to remain at a fixed phase or frequency. However, when comparing different control methods the phase must be taken into account.

The flexibility of the microprocessor control circuit enables the effect of changing one of the firing angles to be investigated. The last firing angle in a half cycle can have the largest range and was selected for investigation. The firing angles produced by the cosinusoidal control method produce the lowest low order harmonic amplitudes, thus confirming the desirability of the cosinusoidal control method.

Using the microprocessor control circuit, operation at maximum output frequency is possible. However, the wanted output component is small and there are large low order harmonics. The possibility of firing a thyristor before the correspondent cosine timing wave period at the beginning of a half cycle offers extra scope of investigation and may

lead to improvement of the output waveform.

Therefore the use of a microprocessor in the control circuit makes possible the implementation of more flexible control strategies together with a reduction of complexity and cost. With the development of more powerful microprocessors, it may eventually be possible to determine and minimise the loss in the load in real time, thus extending the operation frequency of the cycloconverter.

Since the completion of this work a paper by J. T. Boys entitled 'A loss minimised PWM inverter' was published in the IEE Proceedings Vol. 132 Pt.B No.5. It shows that for an inverter the relative phase of the carrier and reference waveforms made small difference to the harmonic content as in the case of cycloconverter shown earlier. However, when a third harmonic was added to the reference wave - in order to increase the output voltage, the phase alter the harmonic content significantly and can reduce harmonic losses at a certain phase angle.

This is a clear indication for further investigation of the relative phase between cosine timing and reference waves by adding a third harmonic to the reference wave.

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ACKNOWLEDGEMENTS

The author would like to thank everyone who helped during the course of this work. In particular,

Dr.Hanley, Dr.Mapps and Prof.Bird for their expert guidance.

All the technicians for helping with the hardware construction.

And staff in the Civil Engineering Department for the use of their spectrum analyser over such a long period of time.

APPENDICES

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APPENDIX I

Modification of the Fourier series.

The input voltage waveforms of a cycloconverter are sinusoidal :

$$\text{i.e.} \quad F(t) = \sin(\omega t + \phi_k)$$

$$\text{where} \quad \phi_k = k * 2\pi/p$$

$$k = 1, 2 \dots p$$

To evaluate A_n one of the term required to intergate is

$$2/T \int_0^T \sin(\omega t/r + \phi_k) * \cos(n\omega t) dt$$

This can be reduce to summation as follow :

$$= 1/T \int_0^T \sin(\omega_D t + \phi_k) + \sin(\omega_S t + \phi_k) dt$$

$$\text{where} \quad \omega_D = \omega(1/r - n)$$

$$\omega_S = \omega(1/r + n)$$

$$= 1/T \left[1/\omega_D * \cos(\omega_D t + \phi_k) + 1/\omega_S * \cos(\omega_S t + \phi_k) \right]_0^T$$

For the complete contribution of input phase :

$$1/T \left\{ \sum_{m=0}^M \left[1/w_D * \cos(w_D t + \phi_k) + 1/w_S * \cos(w_S t + \phi_k) \right]_{t_m}^{t_{m+1}} \right\}$$

Then for p input phases :

$$1/T \left\{ \sum_{k=1}^p \sum_{m=0}^M \left[1/w_D * \cos(w_D t + \phi_k) + 1/w_S * \cos(w_S t + \phi_k) \right]_{t_{(m)k}}^{t_{(m+1)k}} \right\}$$

The order of summation can be interchange so that the limits used are in accending order.

$$1/T \left\{ \sum_{m=0}^M \sum_{k=1}^p \left[1/w_D * \cos(w_D t + \phi_k) + 1/w_S * \cos(w_S t + \phi_k) \right]_{t_{(m)k}}^{t_{(m+1)k}} \right\}$$

Similarly for B_n

$$1/T \left\{ \sum_{m=0}^M \sum_{k=1}^p \left[1/w_D * \sin(w_D t + \phi_k) + 1/w_S * \sin(w_S t + \phi_k) \right]_{t_{(m)k}}^{t_{(m+1)k}} \right\}$$

APPENDIX II:

BASIC program for plotting sinusoidal waveforms.

```
100 REM 26-8-83 TIMING WAVEFORMS
110 PUT$7["INIT",5], "IN; "
120 REM SH=TIME SHIFT TO NEXT PHASE
130 REM SIZE=SIZE OF GRAPH
140 PI=ATN(1)*4/30 : SIZE=2400
150 REM FIRST THREE OUTPUT
160 FOR Z=0 TO 2
170 P$="SP"+STR$(Z+1)+"; " : PUT$7,P$
180 PRINT P$
190 SH=2*SIZE/3*Z:DEL=10 : GOSUB 420
200 NEXT Z
210 REM FIRST THREE TIMING WAVE
220 FOR Z=0 TO 2
230 P$="SP"+STR$(Z+4)+"; " : PUT$7,P$
240 PRINT P$
250 SH=2*SIZE/3*Z : DEL=15 : GOSUB 420
260 NEXT Z
270 REM SECOND THREE OUTPUT
280 FOR Z=0 TO 2
290 P$="SP"+STR$(Z+1)+"; " : PUT$7,P$
300 SH=2*SIZE/3*Z+2*SIZE:DEL=10:GOSUB420
302 NEXT Z
310 FOR Z=0 TO 2
320 P$="SP"+STR$(Z+4)+"; " : PUT$7,P$
330 SH=2*SIZE/3*Z+2*SIZE:DEL=15:GOSUB420
340 NEXT Z
360 PUT$7, "PU; PA 100,100; SP7; SS; LB CYCLOCONVERTER",CHR$(3)
370 PUT$7, "PU; PA 0,4800; PU; PA9600,4800; PA15000,4800; PU; "
375 PI=PI/2
377 FOR Z=0 TO 1
380 SH=2*SIZE/3+Z*SIZE*2 : DEL=Z*30 : GOSUB 420
385 NEXT Z
390 PUT$7, "PU; "
400 END
410 REM SUB. TO PLOT ONE SINE CURVE
420 Y0=SIN(DEL*PI)*SIZE
430 Y0=SIZE*2+Y0
440 P$="PU; PA"+STR$(SH)+"; "+STR$(Y0)+"; PD; "
450 PUT$7,P$ : PRINT P$
460 FOR A=1 TO 30
470 Y=SIN((A+DEL)*PI)*SIZE
480 X=A*SIZE/15+SH : Y=SIZE*2+Y
490 P$="PA"+STR$(X)+"; "+STR$(Y)+"; "
500 PUT$7,P$
510 NEXT A
520 RETURN
530 END
```


APPENDIX III

Listing of Z-80 assembly language program
for firing angle control.

	0001	*HEADING CYCLOCONVERTER
	0002	*FORMFEED OFF
	0003	;18-5-83
	0004	;HL ADDRESS OF FIRING ANGLE
	0005	;D POSITION IN TABLE
	0006	;E BIT 7 =1 CURRENT DETECTION IN OPERATION
	0007	; BIT 0 AND 1 +/- GROUP CONTROL
	0008	;
	0009	RADH ;HEX
2070 =	0010	TIME EQU 2070
000C =	0011	LEN EQU 0C
	0012	;
2000	0013	ORG 2000
2000 FB	0014	D1 ;DISABLE INT
2001 ED5E	0015	IM2 ;MODE 2
	0016	;SET UP JUMP TO 0180
	0017	;
	0018	;SET START ADDRESS AND GROUP
2003 1E81	0019	LD E,81
2005 160C	0020	LD D,LEN
2007 210021	0021	LD HL,2100
	0022	
	0023	;SET MODE WORD
200A 3E3F	0024	LD A,3F
200C D383	0025	OUT (83),A
	0026	;SET INTERRUPT CONTROL WORD
200E 3E87	0027	LD A,87
2010 D383	0028	OUT (83),A
	0029	;SET INTERRUPT VECTOR
2012 3E80	0030	LD A,80
2014 D383	0031	OUT (83),A
2016 3E20	0032	LD A,20 ; INTERRUPT VECTOR
2018 ED47	0033	LD D,A
	0034	;
201A 3EFF	0035	LD A,0FF ;MODE 3
201C D382	0036	OUT (82),A ;CONTROL
	0037	;ONE CURRENT ZERO INPUT AND
	0038	;TWO GROUP CONTROL OUTPUT
201E 3E80	0039	LD A,80
2020 D382	0040	OUT (82),A
2022 3E07	0041	LD A,07 ;DISABLE INTERRUPT
2024 D382	0042	OUT (82),A
	0043	;
2026 FB	0044	EI; ENABLE INTERRUPT
	0045	
	0046	;CARRY ON MAIN PROGRAM
2027 00	0047 MAIN	NOP
2028 CD7020	0048	CALL TIME
202B 18FA	0049	JR MAIN
	0050	
	0051	;TIME DELAY ROUTINE
2070	0052	ORG TIME
2070 3EFF	0053	LD A,0FF
2072 3D	0054 DEL:	DEC A
2073 20FD	0055	JR NZ,DEL

```

2075 09      0056      RET
                0057      ;
2080          0058      ORG 2080          ; INTERRUPT VECTOR
2080 8220     0059      DEFB 82,20
                0060      ; INTERRUPT SERVICE ROUTINE
2082          0061      ORG 2082
                0062      ;
2082 7E      0063      ANG: LD A,(HL)
2083 D381     0064      OUT (81),A
2085 23      0065      INC HL
2086 15      0066      DEC D
2087 200B     0067      JR NZ,PLU
2089 210021   0068      LD HL,2100      ; RESET LOOK-UP TABLE
208C 160C     0069      LD D,LEN      ; ADDRESS AND LENGTH
208E 7B      0070      LD A,E
208F EE83     0071      XOR A,83
2091 5F      0072      LD E,A
2092 D380     0073      OUT (80),A      ; DISABLE BOTH GROUP
                0074      ;
2094 FB      0075      PLU: EI
2095 ED40     0076      RETI
                0077      ;
                0078      ; DATA
                0079      ; NUMBER REPRESENT FIRING ANGLE
2100          0080      ORG 2100
                0081      RADH
2100 4D3C2B1A 0082      DEFB 4DH,3CH,2BH,1AH
2104 021A2B3C 0083      DEFB 02H,1AH,2BH,3CH
2108 4D5E805E 0084      DEFB 4DH,5EH,80H,5EH
210C 4D3C2B1A 0085      DEFB 4DH,3CH,2BH,1AH
0000          0086      END

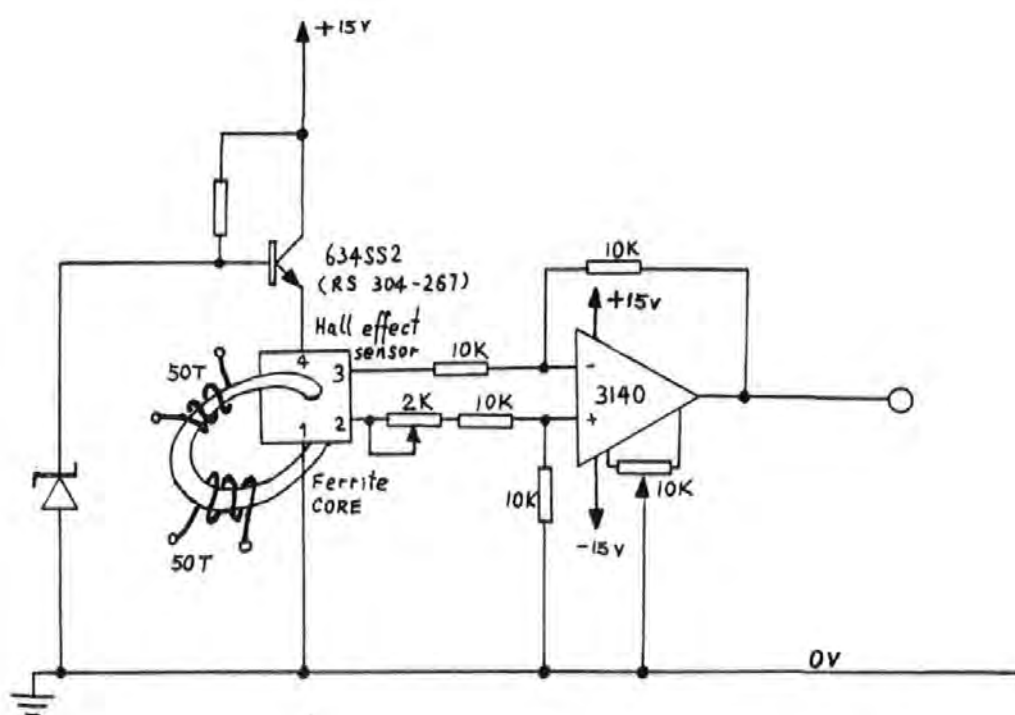
```

2082 ANG 2072 DEL 000C LEN 2027 MAIN 2094 PLU
2070 TIME

NO ERRORS

APPENDIX IV

Circuit diagram of the Hall effect current sensor.



APPENDIX V

Va Listing of Fortran 77 program for calculating harmonic amplitudes.

Vb Typical input data.

Vc Typical results.

APPENDIX Va

THYRISTOR CYCLOCONVERTER WITH SINGLE PHASE OUTPUT
CIRCULATING CURRENT-FREE MODE
R-L LOAD
GROUP CHANGEOVER AT COMPUTATION POINT NEAREST
INSTANT OF ZERO OUTPUT CURRENT AFTER ZERO VR

REQUIRED INPUT DATA
NP=PULSE NUMBER OF INPUT (=2, 3 OR 6 IN THIS PROGRAM)
FI=INPUT FREQUENCY (HZ)
FO=OUTPUT FREQUENCY (HZ)
FF=FREQUENCY OF FUNDAMENTAL (LOWEST FREQUENCY IN OUTPUT)
FO/FF MUST EQUAL AN INTEGER (=IFA)
NP*FI/FF MUST EQUAL AN INTEGER
RAT= NP*FI/FO
NH=HIGHEST ORDER OF HARMONIC IN THE OUTPUT
DM=MODULATION DEPTH

ADDITIONAL SYMBOLS USED IN THIS PROGRAM
ND=NUMBER OF SETS OF DATA IN INPUT FILE
PHASE= PHASE BETWEEN COSINE TIMING AND REFERENCE
TIMCO=TIME CONSTANT OF LOAD
N=ORDER OF HARMONIC, REFERRED TO FF
I=COUNTER FOR SAMPLING
IC=COUNTER FOR COMPUTATION FOR NATURAL SAMPLING FOLLOWED
BY REGULAR SAMPLING
KP=COMMUTATION POINT (=1 AT FIRST COMMUTATION AFTER T=0)
NKP=NO. OF KP IN CC MODE
KC=COMMUTATION POINT FOR CHANGEOVER IN INHIBITED CYCLOCONVERTER
KC=1 IS FOR CHANGEOVER FROM POSITIVE TO NEGATIVE GROUP
FOURIER ANALYSIS IS FROM KC=2 TO AVOID START ERRORS
KCL=LAST KC
KPP(KC)=POSITIVE GROUP KP NEXT TO KC
KPN(KC)=NEGATIVE GROUP KP NEXT TO KC
KPS='DUMMY'KP FOR START OF FOURIER ANALYSIS FOR ONE GROUP
KPF='DUMMY'KP FOR FINISH OF DITTO
TIME=INSTANT AT WHICH COMPUTATION IS TO BE CARRIED OUT
TIME=0.0 SECS AT START OF SINUSOIDAL REFERENCE VOLTAGE
TIMC=TIME OF START OF VC
DT=INCREMENTAL TIME INTERVAL FOR COMPUTATION
TP(KP)=TIME OF COMMUTATION OF POSITIVE GROUP
TN(KP)=DITTO OF NEGATIVE GROUP
TC(KC)=TIME OF START OF CONDUCTION AT
CHANGEOVER COMMUTATION POINT KC
VR=REFERENCE VOLTAGE
AL IS PROPORTIONAL TO THE LOAD CURRENT
ANGC(NO)=COMPUTED (OUTPUT) PHASE SHIFT OF AO FROM VR
VC=COSINUSOIDAL CONTROL VOLTAGE
L=2 FOR POSITIVE GROUP
=1 FOR NEGATIVE GROUP
A(N)=FOURIER COEFFICIENT FOR SINE TERMS
B(N)=FOURIER COEFFICIENT FOR COSINE TERMS
H(N)=MAGNITUDE (PEAK, P.U.) OF THE NTH HARMONIC

```

C      WD=DIFFERENCE IN ANGULAR FREQUENCY BETWEEN FI AND HARMONIC
C      WS=SUM OF DITTO
C      1P.U.=PEAK OF INPUT VOLTAGE
C
C      DIMENSION TP(500),TN(500),TC(120),KPP(120),KPN(120),T(500)
C      DIMENSION H(300),ANGC(300),ALOAD(5000),TLOAD(5000)
C
C      PI=4.0*ATAN(1.0)
C      OPEN (6,FILE='DATA20')
C      READ(5,199)ND
199  FORMAT(I5)
C      DO 90 JOBS=1,ND
C      READ(5,200)FI,RAT,IFA,NH,NP,PHASE,TIMCO
C      PRINT *, 'PHASE=', PHASE
C      DM=1.0
200  FORMAT(2F10.0,3I10,F10.0,F10.0)
C      FO=NP*FI/RAT
C      FF=FO/IFA
C
C      DO 91 IC=1,2
C      PRINT *
C      PRINT *, '***** START *****'
C      PRINT *
C      PRINT *, 'INPUT DATA'
C      PRINT *, 'NUMBER OF PULSES' = ', NP'
C      PRINT *, 'INPUT FREQUENCY' = ', FI, 'HZ'
C      PRINT *, 'RATIO' = ', RAT'
C      PRINT *, 'OUTPUT FREQUENCY' = ', FO, 'HZ'
C      PRINT *, 'FUNDAMENTAL FREQUENCY' = ', FF, 'HZ'
C      PRINT *, 'HIGHEST ORDER OF HARMONIC' = ', NH'
C      PRINT *, 'MODULATION DEPTH' = ', DM, 'P.U.'
C      PRINT *, 'TIME CONSTANT' = ', TIMCO, 'SECS'
C      PRINT *
C      PRINT *, '      *** COSINUSOIDAL TIMING ***'
C
C      IF(IC.EQ.1) THEN
C      PRINT *, 'OUTPUT DATA FOR INDIBITED CYCLO. NATUAL SAMPLING'
C      ELSE
C      PRINT *, 'OUTPUT DATA FOR INHIBITED CYCLO. REGULAR SAMPLING'
C      END IF
C
C      PRINT *, '  COMMUTATION POINTS FOR UNINHIBITED CYCLOCONVERTER'
C      PRINT *, '  GROUP          KP          TIME(SECS)          MODULATION(SECS)'
C
C      CONSTANTS
C      DT=1.0/FI/300.0
C      NKP=IFIX(NP*FI/FF+0.01)+IFIX(3*NP*FI/FO+0.01)
C      NKP1=NKP+1
C      KCL=2+2*IFA
C      WI=2.0*PI*FI
C      WO=2.0*PI*FO
C      FR=FO/FI
C      NIA=IFIX(KCL/2.0/FO/DT/4.0)

```

```

C      IF((KCL .GE. 110) .OR. (NKPL .GE. 500) .OR.
1      (NH .GE. 300) .OR. (NIA .GE. 5000)) THEN
      PRINT *, '      DIMENSION EXCEEDED ????'
      GO TO 90
      END IF
C      *****
C      COMMUTATION POINTS FOR UNINHIBITED CYCLOCONVERTER
      DO 5 L=1,2
      DO 10 KP=1,NKPL
      IF(NP-3) 77,78,79
77 TIMC=(KP-1)/2.0/FI
      GO TO 92
78 TIMC=(2*KP-4+L)/6.0/FI
      GO TO 92
79 TIMC=(KP-3+L)/6.0/FI
92 DO 20 I=1,151
      TIME=(I-1)*DT+TIMC
      VC=(2*L-3)*COS(2.0*PI*(I-1)/300.0)
      IF(IC-1) 93,93,94
C      NATURAL SAMPLING
93 VR=DM*SIN(2.0*PI*FO*TIME+PI*PHASE/48)
      GO TO 95
C      REGULAR SAMPLING
94 VR=DM*SIN(2.0*PI*FO*TIMC+PHASE/48*PI)
C      IS THIS FOR POSITIVE OR NEGATIVE GROUP
95 IF(L-1) 15,15,16
16 IF(VR-VC) 20,21,21
15 IF(VC-VR) 20,22,22
20 CONTINUE
      PRINT *, '??? NO COMMUTATION'
      GO TO 10
C
21 TP(KP)=TIME
      TIMOD=TIME-TIMC
      GO TO 10
22 TN(KP)=TIME
      TIMOD=TIME-TIMC
C
10 CONTINUE
5 CONTINUE
C      *****
C      TIMES FOR CHANGEOVER IN INHIBITED CYCLOCONVERTER
      PRINT *, '      KC      TIME OF VZ      TIME OF AZ'
C
      L=2
      IA=0
      TIMI=0.0
      KCL1=KCL+1
C
      DO 150 KC=1,KCL1
      AO=0.0
      TIMVZ=KC/2.0/FO
      IF(KC .EQ. 1) GO TO 141

```



```

      TIMI=TC(KC-1)
141 IF(L-1)154,154,151
C
C      FIND FIRST KP OF INCOMING(POSITIVE) GROUP
151 DO 153 KP=1,NKP1
      IF(TP(KP) .GE. TIMI) GO TO 152
153 CONTINUE
152 IF(KC .EQ. 1) GO TO 142
      KPP(KC-1)=KP
      GO TO 142
C
C      FIND FIRST KP OF INCOMING(NEGATIVE) GROUP
154 DO 156 KP=1,NKP1
      IF(TN(KP) .GE. TIMI) GO TO 140
156 CONTINUE
140 KPN(KC-1)=KP
C
142 KP1=KP
      DO 146 KP=KP1,1000
      IF(NP-3)158,159,160
158 R=(L-KP)*PI
      VPK=2.0
      GO TO 161
159 R=(4*(L-KP)+1)*PI/6.0
      VPK=1.0
      GOTO 161
160 R=(2*L-KP-1)*PI/3.0
      VPK=SQRT(3.0)
161 TIM=TIMI
162 IF(TIMCO .LT. 0.0001) GO TO 155
      AL=VPK/(1.0+(WI*TIMCO)**2)*(SIN(WI*TIM+R)-WI*TIMCO
1* $\cos(WI*TIM+R)$ +(EXP((TIMI-TIM)/TIMCO))*(WI*TIMCO* $\cos(R+WI*TIMI)$ -
2 $\sin(R+WI*TIMI)$ ))+AO*EXP((TIMI-TIM)/TIMCO)
      GO TO 157
155 AL=VPK*SIN(WI*TIM+R)
157 IF(L-1)168,168,169
168 IF(AL+0.00001)164,170,170
169 IF(AL .GT. 0.00001) GO TO 164
170 AL=(2*L-3)*0.00001
      IF(TIM .GE. TIMVZ) GO TO 163
164 TIM=TIM+DT*4.0
      IF(L-1)143,143,144
143 IF(TIM .LT. TN(KP)) GO TO 165
      TIMI=TN(KP)
      GO TO 145
144 IF(TIM .LT. TP(KP)) GO TO 165
      TIMI=TP(KP)
      GOTO 145
C
165 IA=IA+1
      ALOAD(IA)=AL
      TLOAD(IA)=TIM-DT*4.0
      GO TO 162
C

```

```

145 AO=AL
146 CONTINUE
C
C
C      CHANGEOVER OCCURS NOW
163 IA=IA+1
    ALOAD(IA)=AL
    TLOAD(IA)=TIM
    TIM=TIM+DT*2.0
    TC(KC)=TIM
C      RECORD LAST KP OF OUTGOING GROUP AND THEN CHANGE TO OTHER
GROUP
    IF(L-1) 171,171,166
171 KPN(KC)=KP-1
    L=2
    GO TO 167
166 KPP(KC)=KP-1
    L=1
C
167 PRINT *, '    CHANGEOVER ', KC, TIMVZ, TIM
150 CONTINUE
C
    PRINT *, '    FIRST AND LAST KPS AT CHANGEOVER'
    PRINT *, '                KC                KPP                KPN'
    DO 41 KC=1, KCL
    PRINT *, '    ', KC, KPP(KC), KPN(KC)
41 CONTINUE
C
    ILAST=IA
C      *****
C      FOURIER ANALYSIS
    DO 67 N=1, NH
    A=0.0
    B=0.0
    DCCOMP=0.0
    L=2
    IKP1=2
C
    DO 53 KC=3, KCL
C      WHICH GROUP IS THIS
    IF(L-1) 51, 51, 52
C      POSITIVE GROUP
52 KPS=KPP(KC-1)-1
    KPF=KPP(KC)+1
    GO TO 57
C      NEGATIVE GROUP
51 KPS=KPN(KC-1)-1
    KPF=KPN(KC)+1
C
57 TS=TC(KC-1)
    TF=TC(KC)
    KPS1=KPS+1
    DO 50 KP=KPS1, KPF
C      WHICH GROUP IS THIS

```

```

      IF(L-1) 59,59,58
C     POSITIVE GROUP
58    T(KP)=TP(KP)
      T(KP-1)=TP(KP-1)
      GO TO 64
C     NEGATIVE GROUP
59    T(KP)=TN(KP)
      T(KP-1)=TN(KP-1)
64    T(KPS)=TS
      T(KPF)=TF
C
C     CHECK FOR CURRENT DISCONTINUITIES
      ISTART=IKP1-1
      DO 60 IA=ISTART,ILAST
      IF(TLOAD(IA) .GE. T(KP-1)) GO TO 61
60    CONTINUE
      PRINT *, '    ERROR_'
61    IKP1=IA
      DO 62 IA=IKP1,ILAST
      IF(TLOAD(IA) .GT. T(KP)) GO TO 63
62    CONTINUE
      PRINT *, '    ERROR____'
63    IKP2=IA-1
      IKPX=IKP1+1
      IF(IKPX .GT. IKP2) GO TO 69
      IF(L .EQ. 2) GO TO 42
      IF(ALOAD(IKPX)+0.00002) 43,43,44
42    IF(ALOAD(IKPX)-0.00002) 44,43,43
44    DO 45 IA=IKPX,IKP2
      IF(L .EQ. 2) GO TO 46
      IF(ALOAD(IA)+0.00002) 49,49,45
46    IF(ALOAD(IA)-0.00002) 45,49,49
45    CONTINUE
      T(KP-1)=T(KP)
      GO TO 69
49    T(KP-1)=TLOAD(IA-1)
43    DO 65 IA=IKPX,IKP2
      IF(L .EQ. 2) GO TO 66
      IF(ALOAD(IA)+0.00002) 65,65,68
66    IF(ALOAD(IA)-0.00002) 68,65,65
65    CONTINUE
      GO TO 69
68    T(KP)=TLOAD(IA)
C
69    IF(NP-3) 76,73,74
76    R=(L-KP)*PI
      VPK=2.0
      GO TO 75
73    R=(4*(L-KP)+1)*PI/6.0
      VPK=1.0
      GO TO 75
74    R=(2*L-KP-1)*PI/3.0
      VPK=SQRT(3.0)/PI
75    WD=2.0*PI*(FI-N*FF)

```

```

      WS=2.0*PI*(FI+N*FF)
      DCCOMP=DCCOMP+VPK*FF/WI*(COS(WI*T(KP-1)+R)-COS(WI*T(KP)+R))
C     FOURIER COEFFICIENTS
C     IS N=FI/FF
      Y=N-FI/FF
      ABSY=ABS(Y)
      IF(ABSY-0.01) 48,48,47
47  A=A+VPK*(COS(WS*T(KP-1)+R)-COS(WS*T(KP)+R))*FF/WS
      1+(COS(WD*T(KP-1)+R)-COS(WD*T(KP)+R))*FF/WD*VPK
      B=B+VPK*(SIN(WD*T(KP)+R)-SIN(WD*T(KP-1)+R))*FF/WD
      2-(SIN(WS*T(KP)+R)-SIN(WS*T(KP-1)+R))*FF/WS*VPK
      GO TO 50
48  A=A+VPK*(T(KP)-T(KP-1))*(SIN(R))*FF-(COS(4.0*PI*FI*T(KP)+R)
      3-COS(4.0*PI*FI*T(KP-1)+R))*FF/4.0/PI/FI*VPK
      B=B+VPK*(T(KP)-T(KP-1))*(COS(R))*FF-(SIN(4.0*PI*FI*T(KP)+R)
      4-SIN(4.0*PI*FI*T(KP-1)+R))*FF/4.0/PI/FI*VPK
50  CONTINUE
C
      IF (L .EQ. 1) THEN
        L=2
      ELSE
        L=1
      END IF
53  CONTINUE
C
      HARMONICS DUE TO COMPLETE CYCLOCONVERTER
      H(N)=SQRT(A*A+B*B)
C
      PHASE DIFFERENCE BETWEEN H(N) AND B(N)
      ANG(N)=ATAN2(A,B)*180.0/PI
67  CONTINUE
C
      END FOURIER COEFFICIENTS
C
      OUTPUT VOLTAGE (AT OUTPUT FREQUENCY)
      VO=H(IFA)*100.0
      PRINT *, 'VOLTAGE AT OUTPUT FREQ (VO) = ', VO, ' 0/0 OF INPUT
VOLTAGE'
C
      PHASE ANGLE (AT OUTPUT FREQUENCY)
      PRINT *, 'PHASE BETWEEN VO ,VR= ', ANG(IFA), ' DEGS AT OUTPUT FREQ'
      PSFI=ANG(IFA)*FI/FO
      PRINT *, 'PHASE BETWEEN VO ,VR= ', PSFI, ' DEGS AT INPUT FREQ'
C
      DC COMPONENT
      DCCOMP=DCCOMP*10000.0/VO
      PRINT *, ' DC COMPONENT= ', DCCOMP, '% OF VO'
C
      PRINT *, '      HARMONICS IN OUTPUT'
      PRINT *, '      FREQ(HZ)      0/0 OF VO      PHASE ANGLES'
      PRINT *, '                                     B(N)      FI      FO'
C
      EXPRESS HARMONICS AS 0/0 OF OUTPUT VOLTAGE
      AND CALCULATE DISTORTION FACTOR
      SUSQ=0.0
      SUSQWL=0.0

```

```

DFW2=0.0
DO 70 N=1,NH,2
H(N)=H(N)*100.0/VO*100.0
FREQ=N*FF
ANGFI=ANGC(N)*FI/FREQ
ANGFO=ANGC(N)*FO/FREQ
PRINT *,      FREQ,H(N),ANGC(N),ANGFI,ANGFO
IF (IC.EQ.1) THEN
  WRITE(6,381) FREQ,H(N)
381  FORMAT(2F10.3)
END IF
SUSQ=SUSQ+H(N)**2
SUSQW1=SUSQW1+(H(N)*FO/FREQ)**2
DFW2=DFW2+H(N)*FO/FREQ
70 CONTINUE

C
DF=SUSQ-10000
DFW1=SQRT(SUSQW1)
DFW2=DFW2-100.0
PRINT *, '    UNWEIGHTED DISTORTION FACTOR      = ',DF
PRINT *, '    WEIGHTED(FO/FH) DISTORTION FACTOR   = ',DFW1

C
PRINT *, '    INDUCTIVE WEIGHTED DISTORTION FACTOR = ',DFW2
91 CONTINUE
90 CONTINUE
CLOSE (6)
END

```

APPENDIX Vb

Typical input data.

```

.top.
  18
50 12 1 20 6 -4.0000 0.00
50 12 1 20 6 -3.5000 0.00
50 12 1 20 6 -3.0000 0
50 12 1 20 6 -2.5000 0
50 12 1 20 6 -2.0000 0.0
50 12 1 20 6 -1.5000 0
50 12 1 20 6 -1.0000 0
50 12 1 20 6 -0.5000 0
50 12 1 20 6 0.0000 0
50 12 1 20 6 0.5000 0
50 12 1 20 6 1.0000 0
50 12 1 20 6 1.5000 0.0
50 12 1 20 6 2.0000 0
50 12 1 20 6 2.5000 0
50 12 1 20 6 3.0000 0
50 12 1 20 6 3.5000 0
50 12 1 20 6 4.0000 0
50 12 1 20 6 0.0000 0
123456789012345678901234567890123456789012345678901234567890

```

```

STOP
END
.bottom.

```

APPENDIX VC

Typical results.

PHASE= 0.000000E+00

***** START *****

INPUT DATA

NUMBER OF PULSES	=	6	
INPUT FREQUENCY	=	50.0000	HZ
RATIO	=	12.0000	
OUTPUT FREQUENCY	=	25.0000	HZ
FUNDAMENTAL FREQUENCY	=	25.0000	HZ
HIGHEST ORDER OF HARMONIC	=	20	
MODULATION DEPTH	=	1.00000	P.U.
TIME CONSTANT	=	0.000000E+00SECS	

*** COSINUSOIDAL TIMING ***

OUTPUT DATA FOR INDIBITED CYCLO. NATUAL SAMPLING

COMMUTATION POINTS FOR UNINHIBITED CYCLOCONVERTER

GROUP	KP	TIME(SECS)	MODULATION(SECS)
	KC	TIME OF VZ	TIME OF AZ
CHANGEOVER	1	2.000000E-02	2.026662E-02
CHANGEOVER	2	4.000000E-02	4.026662E-02
CHANGEOVER	3	5.999999E-02	6.026661E-02
CHANGEOVER	4	8.000000E-02	8.026651E-02
CHANGEOVER	5	0.100000	0.100267

FIRST AND LAST KPS AT CHANGEOVER

KC	KPP	KPN
1	5	7
2	12	12
3	17	19
4	24	24

VOLTAGE AT OUTPUT FREQ (VO)= 52.4535 0/0 OF INPUT VOLTAGE

PHASE BETWEEN VO ,VR= -1.64363 DEGS AT OUTPUT FREQ

PHASE BETWEEN VO ,VR= -3.28726 DEGS AT INPUT FREQ

DC COMPONENT= -1.893343E-03% OF VO

HARMONICS IN OUTPUT

FREQ(HZ)	0/0 OF VO	B(N)	FI	FD	
25.0000	100.000	-1.64363	-3.28726	-1.64363	
75.0000	4.17763	-91.4310	-60.9540	-30.4770	
125.000	9.71825	-33.0501	-13.2200	-6.61002	
175.000	12.5405	22.0974	6.31353	3.15677	
225.000	8.13397	76.6993	17.0443	8.52214	
275.000	2.47669	142.710	25.9473	12.9736	
325.000	4.31844	-133.543	-20.5451	-10.2726	
375.000	10.8491	-85.5900	-11.4120	-5.70600	
425.000	8.66870	-139.017	-16.3549	-8.17747	
475.000	8.00488	107.151	11.2791	5.63954	

UNWEIGHTED DISTORTION FACTOR = 617.021

WEIGHTED(FD/FH) DISTORTION FACTOR = 100.054

INDUCTIVE WEIGHTED DISTORTION FACTOR = 8.24326

APPENDIX VI

Tables of computed results when the phase between cosine timing and reference wave is changing.

<PSTAF2>P03408>REPRINT.COMO

[FTN77 VER 188]

NO ERRORS [C. MAIN. >FTN77-VER 188]

PROGRAM ENTERED

FREQUENCY OF HARMONIC

12.500/ 37.500 62.500 87.500 112.500 137.500 162.500 187.500 212.500 237.500

AMPLITUDE OF HARMONIC

	3	5	7	9	11	13	15	17	19
100.000	5.867	5.172	4.692	4.240	4.161	4.413	6.706	8.570	7.384
100.000	4.931	4.312	3.870	3.644	3.726	4.186	6.773	9.189	8.116
100.000	3.909	3.455	3.199	3.081	3.170	3.551	6.193	9.223	8.813
100.000	3.855	3.447	3.244	3.197	3.358	3.738	6.514	10.006	9.488
100.000	2.718	2.698	2.732	2.787	2.875	3.114	5.816	9.814	9.933
100.000	1.857	2.424	2.521	2.462	2.826	2.873	5.210	9.842	10.709
100.000	1.040	1.985	2.163	2.165	2.476	2.452	4.706	9.575	10.898
100.000	0.536	1.731	1.917	1.977	2.243	2.236	4.383	9.725	11.402
100.000	0.494	1.806	1.966	2.049	2.133	2.096	3.976	9.817	11.859
100.000	1.156	1.973	2.340	2.370	2.112	1.665	3.408	9.724	12.269
100.000	1.726	2.165	2.329	2.244	1.899	1.405	3.099	9.609	12.406
100.000	2.118	2.412	2.510	2.351	1.907	1.308	2.636	9.476	12.728
100.000	3.285	3.037	2.695	2.237	1.826	1.342	2.893	9.498	12.387
100.000	3.715	3.075	2.617	2.320	1.543	1.261	2.049	9.169	12.912
100.000	3.911	3.124	2.556	2.229	1.577	1.282	1.378	8.626	13.117
100.000	4.377	3.430	2.766	2.315	1.720	1.329	1.184	8.309	13.234
100.000	4.641	3.545	2.779	2.217	1.835	1.461	1.456	7.650	13.155

FREQUENCY OF HARMONIC

262. 500 287. 500 312. 500 337. 500 362. 500 387. 500 412. 500 437. 500 462. 500 487. 500

AMPLITUDE OF HARMONIC

21	22	23	24	25	26	27	28	29	30
3. 730	3. 865	3. 018	9. 158	10. 142	7. 099	5. 968	4. 164	3. 011	1. 801
3. 717	3. 019	2. 243	8. 803	10. 906	6. 528	4. 892	4. 714	3. 955	1. 943
4. 269	2. 374	1. 664	8. 341	11. 530	7. 234	3. 336	4. 593	4. 386	2. 437
4. 102	2. 258	1. 514	8. 188	11. 479	7. 455	2. 088	4. 830	4. 809	2. 393
4. 744	1. 794	1. 126	7. 839	11. 557	8. 285	0. 851	4. 049	4. 740	2. 853
5. 345	1. 379	0. 805	6. 996	11. 206	8. 579	2. 726	3. 372	4. 652	3. 243
5. 848	0. 935	0. 468	6. 682	10. 985	8. 961	4. 274	3. 278	4. 048	3. 007
6. 201	0. 710	0. 590	6. 506	10. 531	8. 947	5. 305	3. 681	3. 827	2. 965
6. 492	0. 830	0. 969	6. 642	10. 025	8. 503	5. 607	4. 275	4. 074	2. 748
6. 903	1. 376	1. 498	6. 360	9. 660	7. 681	5. 370	5. 107	4. 520	2. 453
7. 079	1. 469	1. 618	6. 079	9. 677	7. 911	4. 704	5. 767	5. 339	2. 245
7. 368	1. 886	2. 004	6. 119	9. 556	7. 783	3. 447	5. 709	5. 852	2. 516
7. 144	1. 279	1. 601	5. 898	9. 544	8. 453	2. 980	5. 278	6. 721	3. 201
7. 586	1. 572	1. 988	6. 045	9. 454	7. 888	2. 912	3. 961	6. 299	3. 629
8. 143	1. 885	2. 334	6. 222	9. 108	7. 736	3. 808	3. 049	5. 725	3. 662
8. 457	2. 068	2. 456	6. 337	9. 003	7. 759	4. 807	2. 692	5. 026	3. 756
8. 966	2. 366	2. 790	6. 428	8. 599	7. 493	5. 444	3. 252	4. 443	3. 438

**** STOP

[FTN77 VER 188]

[FTN77 VER 188]

NO ERRORS [(<.MAIN. >FTN77-VER 188]

PROGRAM ENTERED

FREQUENCY OF HARMONIC

~~15.000~~ 45.000 75.000 105.000 135.000 165.000 195.000 225.000 255.000

AMPLITUDE OF HARMONIC

100.000	6.325	5.726	5.453	5.399	7.679	9.225	7.144	2.927
100.000	5.097	5.146	4.941	4.708	7.224	9.593	8.150	3.654
100.000	3.914	4.001	4.110	4.179	6.182	9.134	9.285	4.133
100.000	3.262	3.091	3.269	3.719	6.192	9.564	10.014	4.354
100.000	2.075	2.195	2.525	2.871	5.505	9.740	9.679	5.287
100.000	1.456	2.162	2.720	2.625	5.137	10.056	10.274	5.635
100.000	0.925	2.019	2.564	2.372	4.989	10.293	11.026	5.680
100.000	0.859	1.794	2.107	2.512	4.785	9.737	11.730	6.135
100.000	0.805	1.771	1.967	2.349	4.402	9.752	12.055	6.581
100.000	1.715	2.209	2.231	1.788	3.424	9.894	11.952	6.766
100.000	1.884	2.194	2.139	1.754	2.792	10.024	12.841	6.945
100.000	3.453	2.939	2.447	1.935	2.838	9.744	12.565	6.775
100.000	3.543	3.175	2.454	2.191	2.990	9.805	12.460	7.244
100.000	3.719	3.428	2.823	2.053	1.967	9.382	12.852	7.778
100.000	4.359	3.854	3.064	1.970	1.391	8.828	13.394	7.684
100.000	4.394	3.709	2.734	2.188	1.057	7.724	13.793	8.880
100.000	4.645	3.811	2.848	2.141	1.725	7.331	13.588	9.028

FREQUENCY OF HARMONIC

285.000 315.000 345.000 375.000 405.000 435.000 465.000 495.000

AMPLITUDE OF HARMONIC

4.343	3.497	10.077	9.131	9.042	4.962	1.349	1.665
3.547	3.115	9.173	10.205	7.501	6.861	1.598	0.638
2.803	2.204	8.844	11.666	5.212	7.252	2.586	0.644
2.004	1.328	8.175	12.862	3.967	7.097	3.136	0.996
1.236	0.544	7.183	13.194	5.681	5.863	3.624	1.507
0.924	0.474	6.941	12.691	7.955	4.252	3.643	1.752
0.745	0.414	6.703	11.833	10.006	2.519	3.314	1.600
1.124	1.142	6.314	10.276	11.648	1.233	2.790	1.079
1.194	1.600	6.282	8.949	12.164	2.943	2.364	1.026
0.890	1.353	6.244	7.953	12.056	5.968	1.338	0.754
1.449	1.575	6.538	8.037	10.652	7.435	1.696	0.295
0.642	1.133	5.987	8.980	9.165	9.515	2.959	1.229
1.207	1.545	6.261	9.732	5.971	9.597	3.771	1.440
1.976	2.119	6.145	10.526	3.634	8.953	4.217	1.781
2.020	2.192	5.903	10.746	4.486	7.598	4.706	1.903
2.658	2.672	6.534	10.175	6.890	5.311	4.413	1.374
2.701	2.742	6.532	9.652	9.202	3.084	4.037	1.313

*** STOP

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NO ERRORS [C.MAIN. >FTN77-VER 188J

PROGRAM ENTERED

FREQUENCY OF HARMONIC

████████ 56.250 93.750 131.250 168.750 206.250 243.750

AMPLITUDE OF HARMONIC

100.000	6.085	6.190	7.937	9.231	7.686	3.128
100.000	4.893	4.913	7.024	10.005	8.401	3.730
100.000	3.994	4.165	6.444	10.084	9.081	4.155
100.000	2.626	3.071	6.056	9.938	9.604	4.619
100.000	2.000	2.568	5.626	10.041	10.060	5.001
100.000	1.463	2.286	5.245	9.993	10.684	5.891
100.000	1.145	2.201	5.160	10.107	11.264	6.074
100.000	0.622	1.917	4.730	10.277	11.884	6.603
100.000	1.575	2.119	4.337	9.877	12.000	6.367
100.000	1.895	2.142	3.882	9.877	12.617	6.680
100.000	1.950	2.223	3.469	9.959	13.292	7.064
100.000	3.422	2.716	3.441	10.167	12.674	7.006
100.000	3.682	2.917	3.058	10.151	13.152	7.261
100.000	3.972	3.318	2.461	10.043	13.330	7.423
100.000	4.068	3.400	1.464	9.402	13.555	7.883
100.000	4.648	3.860	0.805	8.489	13.966	8.463
100.000	4.813	4.029	0.812	7.611	13.808	8.996

FREQUENCY OF HARMONIC

281.250 318.750 356.250 393.750 431.250 468.750 506.250

AMPLITUDE OF HARMONIC

4.368	4.749	8.725	9.853	9.208	2.280	2.074
3.067	3.980	10.269	7.669	10.228	2.233	0.905
2.358	3.219	10.770	6.636	10.652	3.347	0.862
2.163	2.748	11.174	7.342	10.049	4.186	1.879
1.577	2.195	10.956	8.570	9.045	4.719	2.605
0.669	1.268	10.256	11.228	7.016	4.491	3.100
0.958	0.958	9.089	12.818	5.561	4.106	2.569
1.419	1.873	6.899	14.348	4.863	3.532	2.155
0.749	2.634	4.681	14.732	6.041	3.663	0.864
1.141	3.032	2.473	14.409	7.536	3.667	0.499
1.776	3.526	2.821	13.403	8.560	3.305	1.279
1.441	2.627	4.184	11.588	10.521	4.574	2.866
1.773	2.735	6.176	9.467	10.959	4.484	2.861
1.791	2.683	8.832	5.894	10.757	4.376	2.259
2.091	2.380	10.131	3.947	10.012	4.249	1.560
2.206	1.249	10.671	4.944	8.942	4.398	0.815
2.571	1.274	10.611	7.194	7.460	4.162	1.142

**** STOP

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[FTN77 VER 188]

NO ERRORS [C. MAIN. >FTN77-VER 188]

PROGRAM ENTERED

25.000/	75.000	125.000	175.000	225.000	275.000	325.000	375.000	425.000	475.000
100.000	7.923	10.821	8.861	1.910	5.845	3.764	6.678	10.547	8.643
100.000	6.495	11.244	9.573	2.855	6.497	4.608	5.127	12.098	6.299
100.000	4.978	10.362	9.851	4.200	5.938	4.717	4.322	13.131	5.905
100.000	4.024	10.112	10.649	4.848	5.669	5.327	4.777	12.608	7.009
100.000	3.387	9.849	11.340	5.606	5.068	5.737	6.359	11.162	8.851
100.000	3.170	9.653	11.852	6.454	4.122	5.793	8.268	9.052	10.277
100.000	3.531	9.238	11.717	7.181	3.594	5.353	9.398	7.698	11.322
100.000	3.859	9.420	12.142	7.683	2.808	4.982	10.534	7.193	10.233
100.000	4.178	9.718	12.541	8.134	2.477	4.318	10.849	8.669	8.005
100.000	4.339	10.648	12.706	8.323	3.884	3.187	9.913	11.607	4.633
100.000	4.373	10.876	13.246	8.357	4.722	3.426	8.512	12.759	4.589
100.000	5.282	11.559	13.134	7.298	4.770	4.734	6.694	12.973	7.393
100.000	4.895	11.575	13.826	7.191	5.411	5.224	5.404	11.725	9.630
100.000	4.401	11.439	14.512	7.207	6.177	6.065	4.657	9.596	11.090
100.000	3.761	10.923	14.943	7.160	6.600	6.840	4.735	7.188	11.217
100.000	2.841	9.852	15.100	7.419	6.797	7.880	5.126	6.130	9.785
100.000	2.488	8.905	15.164	7.590	6.446	8.401	5.384	7.640	7.633

**** STOP

LFTN77 VER 188J

LFTN77 VER 188J

NO ERRORS [C.MAIN. >FTN77-VER 188J

PROGRAM ENTERED

FREQUENCY OF HARMONIC

~~50.000~~ 90.000 150.000 210.000 270.000 330.000 390.000 450.000 510.000

AMPLITUDE OF HARMONIC

100.000	10.316	8.284	1.296	3.402	0.529	7.535	10.656	7.543
100.000	9.683	8.598	1.041	4.110	0.057	8.190	9.888	7.561
100.000	9.146	9.089	1.565	4.925	0.977	8.687	8.773	8.450
100.000	9.126	9.336	2.678	5.136	2.164	8.958	8.386	9.064
100.000	8.831	10.113	3.788	5.622	3.620	8.117	9.129	8.993
100.000	8.191	10.065	5.514	5.433	4.237	6.235	11.312	8.297
100.000	8.504	10.942	6.530	5.612	5.323	4.205	12.770	6.178
100.000	8.999	11.883	7.523	5.512	6.030	4.173	12.778	6.360
100.000	9.597	12.872	8.507	5.162	6.204	6.743	10.923	8.960
100.000	11.173	12.529	8.581	6.440	5.468	8.338	9.045	12.387
100.000	11.858	13.956	9.271	5.696	5.263	10.575	6.675	11.930
100.000	12.325	15.247	10.186	5.095	4.600	11.293	8.460	8.963
100.000	12.557	16.310	11.189	5.254	4.058	10.200	11.872	5.133
100.000	12.445	16.906	11.903	6.027	4.193	7.729	13.465	5.458
100.000	12.311	16.710	12.923	7.382	4.306	5.743	13.580	7.669
100.000	11.745	16.567	12.825	8.479	5.490	3.438	11.566	9.521
100.000	10.918	16.062	12.256	9.082	6.468	4.754	8.648	9.174

*** STOP

APPENDIX VII

UPEC' 85 paper.

THE OUTPUT HARMONICS OF A DIGITAL CYCLOCONVERTER

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ABSTRACT

A variable in the cosinusoidal method of controlling a thyristor cycloconverter is identified as one which has not received attention in previous publications. The results of an investigation into the effect of this variable on the harmonic spectra of the output voltage waveform are presented, indicating that control strategies for cycloconverters should take this variable into account.

INTRODUCTION

The principles of operation of a thyristor cycloconverter is to modulate the firing angles of two phase controlled converters to produce the positive and negative half cycles current of a near sinusoidal output waveform. The mean output voltage of a phase controlled converter is proportional to the cosine of the delay angles (1). The ideal output voltage of a cycloconverter is sinusoidal. This leads to the cosinusoidal control method.

THE COSINUSOIDAL CONTROL METHOD

The traditional cosinusoidal control method generates two analog signals: the reference wave and the cosine timing waves. The cosine timing waves are synchronized to the supply with their spacing fixed according to the converter configuration - i.e. the period of the supply voltage waveform divided by the pulse number of the circuit, as shown in Fig. 1. The instantaneous value of the cosine timing waves is proportional to the mean output voltage if the next thyristor is fired at that instant of time. One of the appropriate cosine timing waves is compared with the reference wave, and when they are equal, the appropriate thyristor is gated and hence an output wave proportional to the reference wave is generated.

THE PHASE ANGLE BETWEEN COSINE TIMING AND REFERENCE WAVES

With analog control circuitry, the reference wave is produced by a free running oscillator. As it is difficult to maintain an exactly fixed frequency ratio between the input supply voltage and the output of the oscillator, the phase angle between the two waves is meaningless.

With digital control circuitry, the natural choice is to have the cosine timing and reference waves synchronized by using the same clock as reference. This clock signal is derived from, and synchronized to, the input supply using P.L.L. circuitry. Therefore the timing and reference wave are in effect phase-locked to the supply. As pointed out by Tso (2), the timing and reference waves need not be generated physically; for a particular output waveform the delay angles can simply cycle through a sequence stored as a look-up table in memory.

For output frequencies which have output to input frequency ratios that can be analysed by Fourier series, i.e. when the frequency ratio is a rational number, there is a fixed phase relationship between the timing and reference waves. The waveform in Fig. 2a shows that for an input to output frequency ratio of two, the length of one half cycle of the output equals one full cycle of the input. Therefore all half cycles of the output can be of equal length and the values of delay angle in

corresponding pulses are identical. That is, the phase relationship between the timing and reference waves at the start of the period identified for Fourier analysis is the same as at the end or the start of the next half cycle. The phase between them can, of course, be varied over a range of values. Fig. 2a, b and c show three different phase angles. In Fig. 2b, the reference waves start at the point when the cosine timing wave is at its maximum. This phase angle is usually chosen to illustrate the cosinusoidal control method and is taken as the reference i.e. $\delta=0$.

The first and subsequent cross-over points of the waves, or switching instants, varies with the phase between the two waves. Therefore output waveforms derived by the same control method, same circuit configuration and the same output to input frequency ratio, can be different. The harmonic spectra also differ.

The phase shift described is not the same as the phase between reference waves in order to produce poly-phase output. In fact poly-phase reference waves can be shifted together with respect to the timing wave.

Apart from the pulse number of the circuit, the range of phase angle or length of time the reference wave can be shifted before the same output waveform is produced, depends on the ratio of the fundamental repetition frequency to the wanted output frequency (or the quantity F_o/F_f). With six-pulse operation and wanted output frequency equal to fundamental repetition frequency, the range is 60 degrees in the input frequency - 16 or 3.4 ms for a 50 Hz input.

If the fundamental repetition frequency is $1/n$ times the output frequency, the range is reduced by $1/n$ times. For example with fundamental repetition frequency equal to half the wanted output frequency, the range is reduced by half to 30 degrees or 1.66 ms.

An undefined situation occurs if a cross-over point corresponds to when the instantaneous values of both waves equal zero. The cross-over can be included or left out in that half cycle. The two results are shown in Figs. 2a and 2c. The net result is the same as for the output waveform being shifted one pulse from the start to the end of the output half cycle. The problem is similar to that of the possibility of mis-fire when the converter is required to produce maximum output voltage, but the end-stop control approach is not suitable as this would produce an unbalanced output i.e. with a mean d.c. level.

When the repetition frequency becomes much lower than the wanted output frequency, the range of phase become very small and has little effect upon firing angle on successive cycles. If the ratio of output to repetition frequency is an irrational number and the waveform never repeats itself, the phase angle between the cosine timing and reference waves becomes meaningless. However, for waveforms that can be analysed by the Fourier series method, i.e. a waveform with finite repetition period, this phase angle has a finite value. This phase relationship between the cosine timing and reference waves must be taken into consideration when comparing the amplitude of harmonics between different control strategies. There is no known publication that considers this variable, and this therefore presents

a useful area of study. For some drive applications, with output frequency fixed over a long period of time, it could well be profitable to phase shift the reference wave so that the harmonic losses produced in the load are a minimum at that frequency and load displacement angle.

EXPERIMENTAL RESULTS

An experimental six-pulse cycloconverter has been constructed. The firing circuit is an adaptation of the circuit published by S. Ashoka (3) with a single board Z-80 microcomputer for the control. The output voltage waveform is analysed by a HP 1582A FFT spectrum analyser. A Fortran 77 program was also written to verify the experimental results.

Fig. 3 shows the results for an output frequency of 25 Hz and unity modulation depth (i.e. maximum output) with resistive load. This corresponds to the waveforms in Fig. 2 and shows that there is good agreement between the computed and experimental results. It also confirms that for a specific output voltage and frequency, the cosinusoidal control does not uniquely define an output waveform.

The amplitude of harmonics for output frequencies of 12.5 and 15 are also plotted in Fig. 4. This diagram shows the minimum and maximum amplitude of the harmonics for the whole range of phase angles between the cosine timing and reference waves. The dot on the line shows the value that would be obtained if the reference wave is zero at the point when the cosine timing wave is at maximum. It is clear for those two output frequencies that for harmonics with frequencies up to 150 Hz, the minimum possible value occurs with zero phase angle between the two waves.

CONCLUSION

It has been shown that the phase angle between the cosine timing and reference wave affects the output waveform and hence the harmonic content. For low output frequencies, the low order harmonic can be minimized if the instantaneous value of the reference wave is equal to zero when the cosine timing wave is at its peak value.

REFERENCES

1. C. Lander, Power electronics (Book), McGraw Hill 1981. p.78.
2. S. Tso. Microprocessor control of triac cycloconverter, IEE Proc. Vol. 130 Pt.B No.3 May 1983. p.193.
3. S. Bhat. An equidistant trigger scheme for 3- ϕ thyristor converters. Trans. IEEE Vo.69 No.9 Sept. 1981 p.1159.

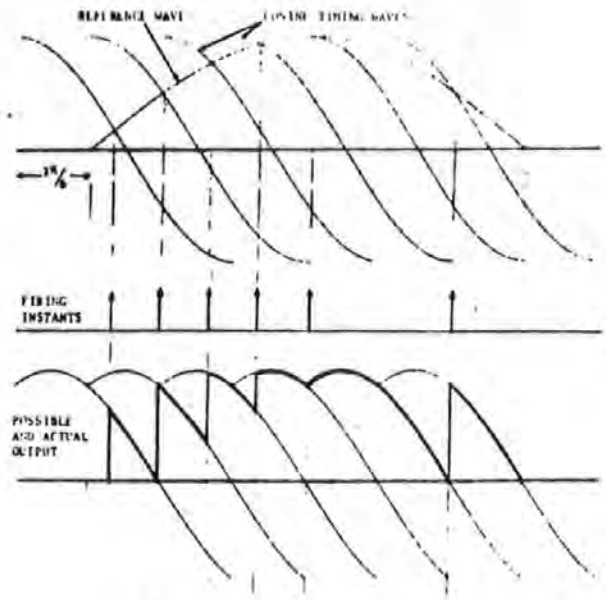


Fig. 1 The cosinusoidal control method.



Fig. 2a



Fig. 2b

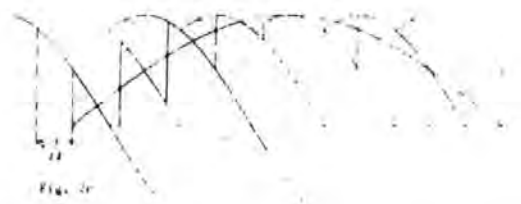


Fig. 2c

Fig. 2 Cosinusoidal control method with different phase angle between cosine timing and reference waves.

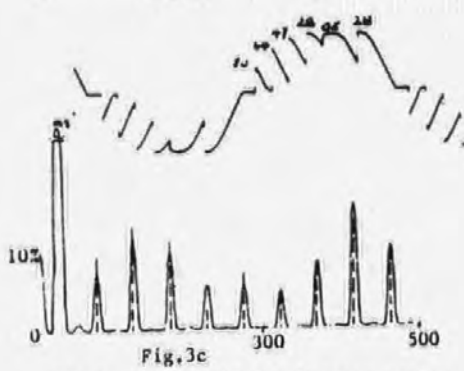
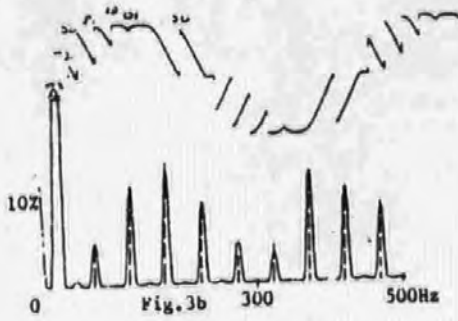
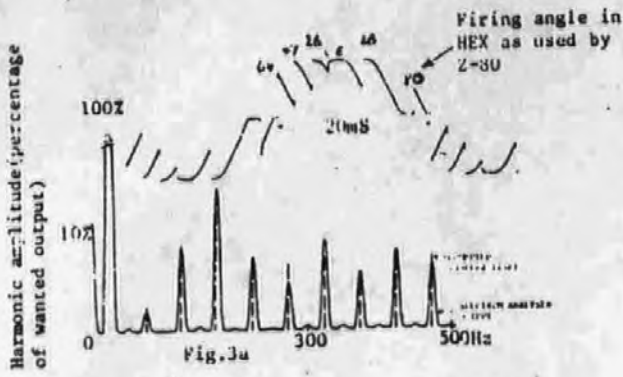


Fig. 3 Phase shifting between cosine timing and reference wave.

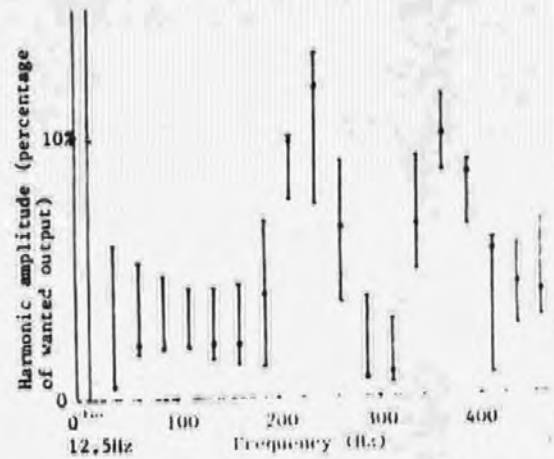
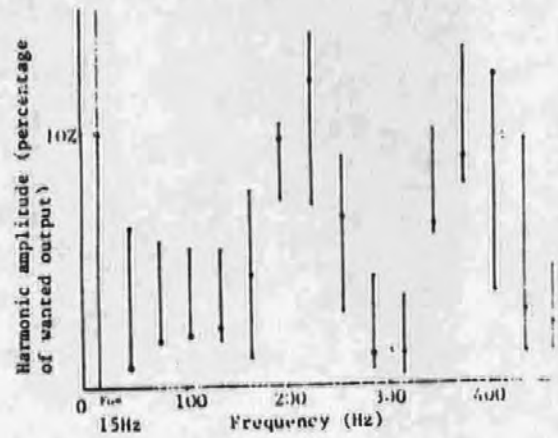


Fig. 4 Max. and min. harmonic amplitude of harmonics over the whole range of phase