APPLICATION OF SAMPLING TECHNIQUES TO THE
PHASE-CONTROLLED THYRISTOR CYCLOCONVERTER

by

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ABSTRACT

Low frequency distortion components of the output voltage of a cycloconverter are largely responsible for the restriction on its practical range of frequencies, and the object of this thesis is to show that these components can be attenuated by the application of sampling techniques to the control system.

After a general description of the operation and control of the cycloconverter, the distortion of the output waveform due to low frequency components is discussed. Under these circumstances, the fundamental repetition frequency of the waveform is less than the wanted output frequency, and two methods of determining it for given input and output frequencies are developed.

The characteristics and properties of the low frequency distortion components, and the requirements for attenuating them are analysed. The particular effects on the magnitudes of these components due to operation of the cycloconverter in the inhibited mode, rather than the circulating-current mode, are examined and the requirements for attenuating them are identified.

It is shown that the communications engineering processes of pulse width modulation and of natural sampling can be identified in the control of the cycloconverter. Regular sampling is more widely used in communications engineering, and its effect on the low
frequency distortion components in the cycloconverter output is compared with natural sampling. A modified control method for the inhibited cycloconverter is then developed to attenuate these components.

Digital computer programs were written to test the effect of introducing modifications to the control of the cycloconverter, and the more significant results are given in graphical and tabulated form. An experimental cycloconverter, with an inhibition control circuit designed for this project, was constructed to check the validity of the computer programs. The design details are described, and the experimental results are discussed.
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LIST OF SYMBOLS

A Amperes
mA Milliamperes
f Frequency

\( f_f \) Fundamental repetition frequency
\( f_h \) Frequency of a harmonic component
\( f_i \) Frequency of the cycloconverter input
\( f_o \) Frequency of the cycloconverter output

F Farads
pF Picofarads
\( \mu F \) Microfarads
Hz Hertz
kHz Kilohertz

p Pulse number of the cycloconverter circuit

r Modulation factor
s Seconds
ms Milliseconds
\( \mu s \) Microseconds
t Time
T Period time
V Volts

w Angular frequency

\( w_f \) Fundamental repetition angular frequency

\( w_i \) Angular frequency of the cycloconverter input
\( w_o \) Angular frequency of the cycloconverter output
\(\alpha\)  Firing angle (or trigger, or delay angle)

\(\circ\)  Degrees (angular)

\(\Omega\)  Ohms

\(k\Omega\)  Kilo-ohms

\(M\Omega\)  Megohms
INTRODUCTION

(1.1) General

The thyristor is a solid-state silicon controlled device which was developed in the mid 1950's to replace the grid-controlled mercury-arc rectifier, and its main advantages over the mercury-arc rectifier are that it is smaller in size, lighter in weight, more robust in terms of mechanical effects, and that it has no vacuum or ignition requirements.

With continuing improvements in the characteristics of the thyristor, reductions in its cost, and reductions in its size/rating ratio, the application of thyristors for the control of electrical power is steadily increasing. Moreover, the availability of low-cost, low-power, integrated circuits enables considerable versatility to be built into the design of the thyristor control systems.

By appropriate control of the thyristors forming part of an electrical circuit, electrical power may be converted or controlled according to the relevant requirements of the application in question. In the case of the cycloconverter, the thyristors are controlled in such a manner that an alternating current supply at one frequency is converted to an alternating current supply at another frequency. The cycloconverter is
therefore a frequency changer. The principle of operation of the cycloconverter is based on the commutation of alternating voltages of differing phase in a similar way to the operation of a rectifier containing more than one rectifying device. The essential difference between a rectifier and a cycloconverter is that a rectifier produces direct voltage at its output whilst the cycloconverter produces alternating voltage.

The basic principles of cycloconversion are analysed by Rissik(1) who records that the first cycloconverters to be commissioned were for railway traction systems requiring fixed frequency ratio. Some of the earlier forms of cycloconverter called 'envelope' cycloconverters, depended on non-controlled mercury-arc rectifiers for their operation. The use of controlled rectifiers provides greater versatility by enabling the input/output frequency and voltage ratios to be varied. The 'phase-controlled' cycloconverter has therefore emerged as the most practical form of cycloconverter, and the availability of thyristors has created a renewed interest in its potential applications.

Cycloconverters may be operated in the circulating-current mode, or in the inhibited mode, or possibly in a combination of both modes. Operation of the cycloconverter in the circulating-current mode results in the flow of circulating current, the magnitude of which
must be limited by circulating-current reactors. For operation in the inhibited mode, reactors are not required because the control system does not allow circulating current to flow. The inhibited cycloconverter therefore has as advantages over the circulating-current cycloconverter, higher power factor and efficiency, and lower weight, size and cost. Additional control circuits are, however, required for operation in the inhibited mode but with the availability of low power integrated circuits, they do not represent a major disadvantage.

(1.2) Cycloconverter Applications

The principle application areas of the cycloconverter are:

a) variable speed control of large a.c. motors for industrial drives such as steelworks, cement mills, and pumps described by Wilson(2), Langer(3) and Weiss(4) respectively;

b) variable speed, constant frequency (VSCF) generating systems for aircraft power systems such as those described by Chirgwin(5), and Piper and Wilcock(6) to give a constant frequency output supply irrespective of the generator frequency which is dependent on the engine speed;

c) variable and fixed frequency supplies for traction drives such as the installation for the Swedish State Railway System described by Klerfors(7).
Other applications include the experimental tracked hovercraft described by F. Fallside et al., (8) and suggested applications extend to a ship's main propulsion and colliery haulage work.

Variable speed drives are more usually provided by d.c. motors fed from the rectified a.c. mains supply, or by variable-speed a.c. motors fed from 'd.c. link' forced commutated inverters. In some applications, however, the use of a d.c. motor must be avoided if the commutator presents problems due to sparking in an explosive environment, excessive wear in a dusty environment, inaccessibility for maintenance, or design and operational difficulties. The other alternative, the d.c.-link inverter, requires forced commutation which introduces technical and economic disadvantages for high-power installations. In particular applications in which these disadvantages can cause serious problems, an alternative method is presented by the cycloconverter which has the advantage of being a direct a.c./a.c. converter employing natural commutation.

(1.3) Limitation of the Frequency Ratio

The main disadvantage of the cycloconverter for variable-speed control applications is that its maximum working output frequency is less than the frequency of the input supply. This restriction is due to the presence of unwanted distortion components in the output voltage waveform. The actual maximum limit
on the output frequency depends on the acceptable level of distortion in the output voltage waveform, since in general, the magnitude of the distortion components increases with increase in output frequency.

Opinions on the maximum limit of the output frequency vary widely because it is not easy to quantify the 'acceptable level of distortion' to cover all applications of the cycloconverter. McMurray(9) attempts to do so by specifying a criterion for the 'Nth degree of good waveform'. Another factor which affects the maximum limit of the output frequency is the number of phases of the input supply voltage. An increase in the number of phases (or an appropriate alternative cycloconverter circuit which increases the apparent number of phases), eliminates some of the distortion components resulting in an increase in the maximum limit of the output frequency. Very approximately, the maximum limit derived from McMurray's analysis is in the range 35-60% of the input frequency, whilst Pelly(10) quotes a range of 33-50%. Pelly emphasises that these figures are based upon purely arbitrarily defined criteria. Similar figures are quoted by Langer(3), and Weiss(4) and Takahashi and Miyairi(11).

It is shown by Pelly(10) and Takahashi and Miyairi(11) that some of the distortion components may be at frequencies less than the wanted output frequency. These 'sub-harmonic' frequency components
are not easily filtered out, and they give rise to high sub-harmonic fluxes in motors supplied from the cycloconverter. This may result in pulsations of the developed torque and in the motor locking onto a 'sub-synchronous' speed. Furthermore, the associated losses reduce the efficiency of the motor. The higher frequency components tend to be suppressed by the inductance of the windings of the motor, and are more easily filtered from the output voltage. The super-harmonics, that is the harmonic components having frequencies greater than the wanted output frequency, therefore have less effect on the torque and efficiency of the motor than the sub-harmonic components. Thus, in order to increase the working frequency range of the cycloconverter, it is necessary to reduce the magnitude of the sub-harmonic components in the output voltage.

(1.4) Aim of this Thesis

It is shown above that the sub-harmonic components of the output voltage of the cycloconverter must be suppressed if the working frequency range is to be increased. To this end, the aim of this thesis is to investigate the characteristics of these components and to show that they can be attenuated by appropriate modification of the control technique.

It has been shown (Ford12) that frequency changing in a cycloconverter can be related to frequency changing in communications engineering in terms of modulation, and that a process of natural sampling can
be identified in the control of the cycloconverter. Replacing natural by regular sampling is shown to result in significant attenuation of the sub-harmonic components when the cycloconverter is operated in the circulating-current mode.

However, as discussed in section 1.1, operation of the cycloconverter in the inhibited mode has practical advantages over operation in the circulating-current mode, and it is to the application of sampling techniques to the control of the inhibited cycloconverter that the majority of this thesis relates. It is proposed to analyse the sub-harmonic components of the output voltage of the inhibited cycloconverter, and to investigate a new approach to the control technique, using sampling, to attenuate these components.

The digital computer was used for most of the analysis since this permits a quick assessment to be made of the effect of modifications to the control method. Analysis by computer also enables the effect of different control methods to be studied without the essential details being obscured by practical issues, such as imperfections in the pulse timing circuits which control the gating of the thyristors. The basic computer results are checked against practical results obtained from an experimental cycloconverter designed and constructed for this purpose.
CHAPTER 2

OPERATION OF THE PHASE-CONTROLLED THYRISTOR CYCLOCONVERTER

(2.1) Introduction

The phase-controlled thyristor cycloconverter can be considered as an array of thyristor switches which connect each of the input supply phases to the load in turn by natural commutation. Typical thyristor configurations making up a 3-phase cycloconverter circuit are shown in figure 2.1. Figure 2.1a shows the simplest (3-pulse) configuration, whilst figure 2.1b shows a 6-pulse circuit which gives an output waveform similar to that obtained from a 6-phase cycloconverter.

The thyristors are switched in an appropriate sequence by means of firing, or switching, signals applied to their gates. By this means, the output voltage waveform is synthesised from segments of the phase voltage waveforms of the input supply, and in this respect, cycloconverter operation is similar to rectifier operation.

In the thyristor rectifier, the switching instants are defined by the firing angles (α) which can be controlled from 0° to about 180° for each thyristor, and which are normally held at a constant value as shown in figure 2.2a to give an output voltage waveform having a predominant d.c. component. The firing angle represents the delay of the switching instant from the earliest possible
FIG 2.1 THYRISTOR CONFIGURATIONS IN 3-PHASE CYCLOCONVERTERS

a) 3-PULSE CIRCUIT

b) 6-PULSE CIRCUIT
WANTED D.C.
OUTPUT COMPONENT

a) RECTIFIER OPERATION

WANTED A.C.
OUTPUT COMPONENT

b) CYCLOCONVERTER OPERATION

FIG 2.2 COMPARISON OF OUTPUT VOLTAGE
WAVEFORMS FOR RECTIFIER AND
CYCLOCONVERTER OPERATION
instant at which the thyristor can turn on.

In the thyristor cycloconverter, the firing angles are not held at constant values as in the rectifier, but vary for successive commutations as shown in figure 2.2b. Thus, the constant firing angles in the rectifier are modulated for cycloconverter operation to give an output voltage waveform which has a predominant a.c. component having the required frequency.

The usual method of controlling the cycloconverter to give the required output frequency is known as cosinusoidal control. This is considered by McMurray(9), Pelly(10), Bland(13) and Datta(14) as the most appropriate control method since it gives an output voltage having less total distortion from the required sinusoid than occurs with other control methods.

To facilitate the analysis of the cycloconverter, it is convenient to divide the array of thyristors which constitute each output phase of the cycloconverter into two groups: the positive group and the negative group. The positive group supplies the positive half cycle of load current, and the negative group supplies the negative half cycle. This arrangement is shown schematically in figure 2.3. The cycloconverter may be operated in the circulating-current mode or in the inhibited mode. In the circulating-current mode, both groups are in
FIG 2.3 SCHEMATIC DIAGRAMS OF THE CYCLOCONVERTER
continuous conduction and a circulating-current reactor is required to limit the magnitude of the circulating current; in the inhibited mode, each group conducts alternately and therefore circulating current cannot flow.

For a 3-phase load, three similar cycloconverters are connected in parallel on the input side, whilst the outputs are connected as required by the phase connections of the load. Such an arrangement for a 3-pulse inhibited cycloconverter is shown in figure 2.4.

(2.2) **Cosinusoidal Control**

Cosinusoidal control employs a series of cosine timing waves at input frequency and a sinusoidal reference wave at the wanted output frequency. The cosine timing waves correspond to the periods during which current can be naturally commutated from one thyristor to the next. It is convenient to derive the timing waves from the input supply voltages, but alternatively they can be derived independently, provided they are then synchronised with the supply voltage. The reference voltage is derived from a variable frequency signal generator.

The instantaneous magnitudes of the timing and reference waves are compared in the control system, and the switching instants (commutation points) of the
P = POSITIVE GROUP
N = NEGATIVE GROUP

FIG. 2.4  3-PULSE INHIBITED CYCLOCONVERTER
WITH 3-PHASE OUTPUT
thyristors coincide with intersections of the two waves. This is shown in figure 2.5, and the modulated sequence of firing angles results in the alternating output voltage.

The ratio of the magnitudes of the reference and timing waves is defined as the 'modulation factor'. For unity modulation factor, the firing angles are modulated successively between the limits of $0^\circ$ and $180^\circ$ resulting in maximum magnitude of the wanted component of output voltage. For zero modulation factor, the reference voltage is zero and the firing angles remain fixed at $90^\circ$ giving zero mean output voltage.

For a cycloconverter feeding a 3-phase load, three reference waves are required with a phase displacement between each of $2\pi/3$ radians. Each of the reference waves is fed to each of three similar cycloconverters which make up the 3-phase cycloconverter.

(2.3) The Circulating-current Cycloconverter

In the circulating-current cycloconverter, firing pulses are applied continuously to both the positive and negative groups. The control system is designed so that the output voltage waveforms of the two groups have the same wanted component; typical waveforms are shown in figures 2.6a and 2.6b. Load current flows from whichever group is forward biased by the combined effect of the input voltage and the back-
3-PHASE SUPPLY VOLTAGES

COSINE TIMING WAVES AND SINUSOIDAL REFERENCE WAVE

COMMUTATION POINTS

OUTPUT VOLTAGE WAVEFORM

FIG 2.5 COSinusoidal Control OF A Three-Pulse Cycloconverter
FIG 2.6 OUTPUT VOLTAGE WAVEFORMS FOR A 3-PULSE CIRCULATING-CURRENT CYCLOCONVERTER.
The instantaneous differences between the two voltage waveforms shown in figures 2.6a and 2.6b cause instantaneous forward biasing of the group which is not conducting load current, resulting in the flow of circulating current between the two groups. To limit the magnitude of the circulating current, a circulating-current reactor is required in each output phase of the cycloconverter. The output terminals of the cycloconverter are located at the centre taps of the reactors, so that the output voltage waveform of the cycloconverter is composed of the averages of the instantaneous voltages of the separate groups. Figure 2.6c shows the output voltage waveform of the cycloconverter corresponding to the two group waveforms shown in figures 2.6a and 2.6b. Since the wanted components for the positive and negative groups are identical, then at wanted output frequency there is no volt drop across the reactor and the wanted component of the cycloconverter output is identical with those for the separate groups.

(2.4) The Inhibited Cycloconverter

In the inhibited cycloconverter, firing pulses are applied alternately to the two thyristor groups for durations equivalent to the respective half cycles of load current. During the conducting periods of
each group, the commutation points are determined by the same method of control as for the circulating-current cycloconverter. Thus, for cosinusoidal control of the inhibited cycloconverter, the commutation points are determined as in figure 2.5.

During the non-conducting periods of each group, the firing pulses are blocked and the group is said to be 'inhibited' so that circulating current cannot flow. Circulating-current reactors are not therefore required.

Although the inhibited cycloconverter does not require circulating-current reactors, it does require additional electronic circuits for inhibition control. Figure 2.7 shows a schematic diagram of a cycloconverter with a method of inhibition control using zero current detection.

The requirements of the inhibition control system are to determine the correct instant for changeover of load current from one group to the other, and then to inhibit the outgoing group and to release the firing pulses of the incoming group. A time-delay is required between inhibition and the release of firing pulses to ensure that the thyristors in the outgoing group have regained their fully blocked (non-conducting) state. Failure to provide sufficient time-delay can result in
FIG. 2-7 SCHEMATIC DIAGRAM OF POWER AND CONTROL CIRCUITS OF AN INHIBITED CYCLOCONVERTER
a short-circuit across the input phases.

The output voltage waveform of the inhibited cycloconverter is effectively synthesised from portions of the separate positive and negative group voltage waveforms of the circulating-current cycloconverter (figures 2.6a and 2.6b). The instants of transition between the positive and negative group waveforms depend on the instants of the load-current zeros. The output voltage waveform of the inhibited cycloconverter is therefore load-dependent. Figure 2.8 shows examples of output voltage waveforms for a 3-pulse inhibited cycloconverter for two different loads, one having unity displacement factor and the other having a lagging displacement factor. The term 'displacement factor' is defined as the cosine of the phase difference between the wanted components of the non-sinusoidal voltage and current waveforms. The waveforms shown in figure 2.8 are based on the assumptions that the back-e.m.f. of the load is at all times able to sustain continuous current, and that changeover of load current from one group to the other occurs without time-delay.

(2.5) Distortion in the Output Voltage Waveform

A visual comparison of figures 2.6c and 2.8 leads to a qualitative conclusion that the output voltage waveform of the inhibited cycloconverter is more
FIG 2-8 OUTPUT VOLTAGE WAVEFORMS FOR A 3-PULSE INHIBITED CYCLOCONVERTER
irregular and therefore more distorted than that of the circulating-current cycloconverter. It is not possible, however, to draw quantitative conclusions about the relative magnitudes and frequencies of the distortion components by a visual comparison; this can only be done by mathematical analysis or by experiment, as discussed later in this thesis.

The extent of the distortion of the output voltage waveform from a pure sinusoid is closely related to the range of working output frequencies which can be obtained from the cycloconverter. In particular, as discussed in chapter 1, the upper limit of the output frequency is imposed mainly by the effect on the load of low frequency components (especially sub-harmonic components). A close examination of the conditions which result in the presence of sub-harmonics is therefore of particular relevance.

(2.6) Sub-Harmonics and the Determination of the Fundamental Frequency

(2.6.1) General

The waveforms of consecutive cycles of the output voltage of the cycloconverter are not necessarily identical, and in order to investigate the sub-harmonic content of the output voltage, it is necessary to identify the period over which the waveform is repeated.
As an example, figures 2.9a and 2.9b are waveforms for a 50 Hz, 3-pulse inhibited cycloconverter having output frequencies of 25 Hz and 20 Hz respectively.

In figure 2.9a, all 25 Hz cycles are identical with each other; that is, the output voltage waveform is repeated every cycle of the wanted output frequency. The repetition frequency is therefore equal to 25 Hz in this case.

In figure 2.9b, the waveform is repeated after two cycles of the wanted output frequency. Therefore in this case, the repetition frequency is 10 Hz.

The examples given in figure 2.9 enable the repetition of the output voltage waveform to be seen clearly. Other output frequencies would be associated with waveforms which may not be repeated until after a large number of cycles at output frequency. In such cases, the repetition frequencies would be considerably less than the respective output frequencies.

(2.6.2) Fourier Analysis of the Output Voltage Waveform

The magnitudes of the wanted and the distortion components in the output voltage of the cycloconverter are calculated by Fourier Analysis. The analysis must be carried out over a period which completely defines the waveform, and the inverse of this period gives the
FIG 2.9 OUTPUT VOLTAGE WAVEFORMS OF 50Hz 3- PULSE INHIBITED CYCLOCONVERTER
fundamental frequency of the Fourier Series.

The output voltage waveform of the cycloconverter is completely defined by the repetition period, and therefore the fundamental frequency of the Fourier Series is given by the repetition frequency. When the repetition frequency is less than the wanted output frequency, the wanted component of the output voltage is one of the harmonic components defined by the Fourier Series. The fundamental component, together with any other components having a frequency less than that of the wanted component, are termed the 'sub-harmonic' components. As discussed in chapter 1, it is essential that the sub-harmonic components should have low magnitudes since they tend, more than the super-harmonic components, to be responsible for limiting the range of acceptable output frequencies obtainable from the cycloconverter.

Before the harmonic components of the output voltage can be computed for a given set of operating conditions, the appropriate values of the fundamental frequency must be determined. Two methods were developed and they are discussed in the following sections. The analyses of the output voltage waveforms were then carried out using computer programs which are described in chapters 3 and 4.
(2.6.3) **Determination of the Fundamental Frequency: First Method**

In the cosinusoidal method of controlling the cycloconverter, the number of cosine timing waves per second is equal to the product of the pulse number \( p \) of the cycloconverter circuit and the input frequency \( f_i \). Denoting the fundamental (repetition) frequency by \( f_f \), the number of cosine timing waves during one repetition period \( 1/f_f \) is therefore equal to

\[
\frac{pf_i}{f_f}
\]

The number of cosine timing waves in one repetition period must be an integer, and therefore the value of \( f_f \) must be selected to give an integer value to the above expression.

\[
\therefore \frac{pf_i}{f_f} = \text{integer} \tag{2.2}
\]

From the discussion in sections 2.6.1 and 2.6.2 on the repetition and fundamental frequencies of the output voltage waveform, it follows that there must be an integer number of cycles of the wanted component (at output frequency, \( f_0 \)) in each cycle of the fundamental component. Therefore another necessary condition for determining the fundamental frequency is that

\[
\frac{f_0}{f_f} = \text{integer} \tag{2.3}
\]

For a given set of values of \( p, f_i \) and \( f_0 \), the highest value of \( f_f \) which satisfies the two conditions
given by expressions (2.2) and (2.3) is the required value of fundamental frequency to be used in the Fourier Analysis. Any other, lower, value of \( f_f \) which also satisfies the two conditions could also be used, but the computer storage requirements would be unnecessarily greater, and the run times would be longer.

The above procedure was followed for the earlier computer analysis, but an alternative method of determining the fundamental frequency was developed later. This method is described in the next section.

(2.6.4) Determination of the Fundamental Frequency: Second Method

Dividing expression (2.2) by expression (2.3) gives:

\[
\frac{pf_i}{ff} \div \frac{f_o}{f_f} = \frac{pf_i}{f_o}
\]  

(2.4)

As discussed in section 2.6.3, \( \frac{pf_i}{f_f} \) and \( \frac{f_o}{f_f} \) must take integer values. However, \( \frac{pf_i}{f_o} \) depends on any combination of values of \( p, f_i \) and \( f_o \) and can therefore be any integer or non-integer value. Manipulation of this value of \( \frac{pf_i}{f_o} \) (hereafter referred to as the \( \frac{pf_i}{f_o} \) Ratio) forms the basis of the second method of determining the fundamental frequency. It should be noted that, although a particular combination of values of \( p, f_i \) and \( f_o \) gives a particular value of the \( \frac{pf_i}{f_o} \) Ratio, the same Ratio value will also be representative of other combinations of \( p, f_i \) and \( f_o \).
The first step in this method is to convert the numerical value of the $p_{f_1}/f_0$ Ratio to a fractional number of the form:

$$M \frac{N}{D} \quad (2.5)$$

For example, if $p = 3$, $f_1 = 50$ Hz and $f_0 = 33\frac{1}{3}$ Hz, then the Ratio is 4.5. As an alternative, the $p_{f_1}/f_0$ Ratio is also 4.5 for $p = 6$, $f_1 = 60$ Hz and $f_0 = 80$ Hz. Thus in either case, the $p_{f_1}/f_0$ Ratio in fractional form is $4\frac{1}{2}$ which from expression (2.5) above, gives $M = 4$, $N = 1$, $D = 2$. To ensure that the selected value ($F_f$) of the fundamental frequency is the highest value, the fractional element $\frac{N}{D}$ must be in minimal form; that is, $4\frac{1}{2}$ must be reduced to $4\frac{1}{2}$.

The numerical value of expression (2.5) is then re-written as the quotient of two components in the form of the left-hand side of expression (2.4):

$$\frac{MD + N}{F_f} = \frac{D}{F_f} \quad (2.6)$$

The two components of this expression must again take integer values, and the only value of $F_f$ which will inevitably satisfy this condition is 1 Hz. Expression (2.6) then becomes:

$$\frac{MD + N}{I} = \frac{D}{I} \quad (2.7)$$

Comparing expressions (2.4) and (2.7) gives:

$$f_0 = D \text{ Hz} \quad \text{(2.8a)}$$

$$p_{f_1} = MD + N \text{ Hz} \quad \text{(2.8b)}$$

$$f_f = 1 \text{ Hz} \quad \text{(2.8c)}$$
The significance of these expressions is that when the \( pf_i/f_o \) Ratio of the cycloconverter has the numerical value \( N \frac{M}{D} \), the fundamental frequency is 1 Hz for input and output frequencies as given by expressions (2.8a) and (2.8b). For any other input and output frequencies, the fundamental frequency is obtained from expressions (2.8a) and (2.8c) to give:

\[
f_f = \frac{f_o}{D}
\]  

(2.9)

Thus in order to determine the fundamental frequency, it is only necessary to consider the element \( \frac{N}{D} \) of expression (2.5). Hence, for all \( pf_i/f_o \) Ratio values which give \( D = 2 \) in expression (2.5), that is \( 1\frac{1}{2}, 2\frac{1}{2}, 3\frac{1}{2}, 4\frac{1}{2}, 5\frac{1}{2}, \) etc., the fundamental frequency is always \( \frac{1}{2} \) of the output frequency. Similarly, for \( pf_i/f_o \) Ratio values of \( 1\frac{1}{3}, 1\frac{2}{3}, 2\frac{1}{3}, 2\frac{2}{3}, 3\frac{1}{3}, 3\frac{2}{3}, \) etc., for which \( D = 3 \), the fundamental frequency is always \( \frac{1}{3} \) of the output frequency.

For \( D = 2 \), the wanted component is the second harmonic of the fundamental component, whilst for \( D = 3 \), it is the third harmonic.

Similar conclusions can be drawn for any other value of \( D \). In the case of \( D = 1 \), for which the \( pf_i/f_o \) Ratio values are integers, the fundamental frequency is equal to the output frequency; in this case there can be no sub-harmonic components in the output voltage waveform.
In the example quoted earlier in this section for $p = 3$, $f_1 = 50$ Hz and $f_0 = 33\frac{1}{2}$ Hz, giving a ratio $\frac{f_1}{f_0}$ = $\frac{33\frac{1}{2}}{2} = 16\frac{1}{2}$ Hz. In the example where $p = 6$, $f_1 = 60$ Hz and $f_0 = 80$ Hz which also gives a ratio of $4\frac{1}{2}$, the fundamental frequency is $\frac{80}{2} = 40$ Hz.

The above procedure for determining the fundamental frequency is readily written into the computer program, and it replaced the first method described in the previous section. It also enables the output data of the computer analysis to be of a more generalised form than with the first method. This is discussed in greater detail in chapter 4.
CHAPTER 3

MODULATION AND SAMPLING IN THE CYCLOCONVERTER

(3.1) Introduction

In the cosinusoidal method of controlling the cycloconverter described in chapter 2, the thyristor switching instants depend not upon the continuous value of the reference wave, but upon its values at discrete points in time. These values can therefore be considered as samples of the reference wave. This concept was recognised by Griffith and Ulmer (15), and Ford (12) investigated the sampling process in the circulating-current cycloconverter in the context of the application of communication techniques. Ford's investigation was based on the principle that whether frequency-changing occurs in communication practice or in power practice, it can only be achieved by modulation. A corollary to this principle is that a cycloconverter is a modulator, and this opens up the possibility of applying techniques used in communications engineering to the cycloconverter.

In this chapter, further consideration is given to the identification and application of sampling and modulation in the cycloconverter. The effect of resultant modifications to the control system on the magnitudes of the principle sub-harmonic components of the output voltages is investigated by digital computer.
(3.2) **Amplitude Modulation in the Cycloconverter**

Consider the operation of one group of the cycloconverter with a constant d.c. reference voltage. The output voltage waveform (figure 3.1a) is that of a rectifier operated with a fixed firing angle $\alpha$, and can be considered as a modulated wave. Each input phase voltage is modulated (multiplied) by a square-wave switching function corresponding to the switching instants of the thyristors and to the conducting periods of the associated input phase. The switching function has unity magnitude and its mark/space ratio is the inverse of the number of input phases. The fundamental frequency of the switching function is equal to the input frequency ($w_i$), and its phase displacement from the associated input phase voltage is given by $\alpha$; in terms used in communications engineering, the input phase voltage is the carrier wave, and the switching function is the modulating wave. The modulated waves associated with the separate input phases are multiplexed to give the resultant output waveform.

For cycloconverter action (figure 3.1b), the firing angle is continuously varied under sinusoidal control, as described in chapter 2. For a required output frequency $w_o$, the input reference voltage is $r \sin w_o t$, where $r$ is the modulation factor which is
FIG 3.1 SWITCHING FUNCTIONS AND MODIFIED SWITCHING FUNCTIONS
the magnitude of the reference wave relative to the magnitude of the cosine timing waves. The switching instants are given by the values of t which satisfy the equation

\[(r \sin w_0 t) = \cos (w_1 t - \theta_k)\]  

(3.1)

where \(\theta_k\) is the phase displacement of the \(k^{th}\) cosine timing wave from an arbitrary reference point.

The thyristor switching functions, \(f(w_1 t, w_0 t)\), which result from these switching instants have variable mark/space ratio as shown in figure 3.1b and are functions of the input frequency \(w_1\) and the output frequency \(w_0\). In all other respects the identification of modulation for cycloconverter action is the same as for rectifier action as discussed above. The output voltage of a cycloconverter can therefore be considered as the result of amplitude modulating the carrier input supply voltages with the unity magnitude thyristor switching functions. However, it is to be emphasised that these switching functions are not externally applied voltages but are merely modulation components of the output voltage waveform.

It has been shown above, that the output voltage waveform of the cycloconverter is dependent on the nature of the thyristor switching function which, in turn, is dependent on the switching instants. However, there is not just one unique switching function
for a given output frequency; control methods other than cosinusoidal control will result in different switching functions. The differing effect of these switching functions is in the magnitude and frequency of the harmonic components of the output voltage. Since the main objective of this investigation is to reduce the magnitude of the sub-harmonics, the effect of small but highly significant variations in the switching instants is of particular interest.

(3.3) Identification of Pulse Width Modulation and Natural Sampling in the Cycloconverter

Cattermole (16) describes the well-established telecommunications process of obtaining pulse time modulation (PTM) of a signal by natural sampling. The basis of this process is to compare the signal with a reference waveform which may be either a saw-tooth or some other waveform containing a periodically repeated ramp. The sampling epochs are defined by the intersections of the signal with the reference wave. A sequence of impulses at the epochs so defined constitutes naturally-sampled PTM, and a sequence of rectangular pulses whose leading edges occur at uniformly spaced instants and whose trailing edges are defined by the sampling epochs, constitute naturally-sampled pulse width modulation (PWM) of the original signal.
The above description of PTM and PWM can be directly related to the cycloconverter by rephrasing the terms 'signal' and 'reference wave' used in the telecommunications context, to 'reference wave' and 'cosine timing waves' respectively used in the cycloconverter context. The cosine timing waves are therefore used in place of the saw-tooth waveform referred to by Cattermole, and the starts of the cosine timing waves define the reference points for the PTM impulses and also the leading edges of the rectangular PWM pulses. The sampling epochs defining the naturally sampled PTM impulses and the naturally sampled PWM trailing edges are coincident with the thyristor switching instants.

It therefore follows that conventional cosinusoidal control of the cycloconverter can be identified with the process of natural sampling of the reference wave. The continuously varying value of the reference voltage between the sampling instants has no significance, and the reference wave is represented by time deviations from the starts of the cosine timing waves. The resulting trains of PTM impulses are shown in figure 3.2 and the corresponding trains of PWM pulses are shown in figure 3.3.

Cattermole also describes 'double-edged' PWM in which the reference wave (in the telecommunications context) has two slopes as shown in figure 3.4a giving
FIG 3·2 COSINUSOIDAL CONTROL WITH NATURAL SAMPLING OF THE REFERENCE WAVE: PULSE TIME MODULATED IMPULSES
FIG 3.3 COSINUSOIDAL CONTROL WITH NATURAL SAMPLING OF THE REFERENCE WAVE: PULSE WIDTH MODULATED PULSES
a) LINEAR MODULATION

b) NON-LINEAR MODULATION (ONE COSINE WAVE)

c) NON-LINEAR MODULATION (TWO COSINE WAVES)

FIG 3.4 DOUBLE-EDGED PWM
linear modulation. The equivalent process in the cycloconverter would be to utilise the continuous cosine timing wave over its range from 0 to $2\pi$ etc., as shown in figure 3.4b. However, without recourse to forced commutation techniques, it is not possible to employ a thyristor firing angle ($\alpha$) in the range $180^0 \leq \alpha \leq 360^0$ corresponding to the positive slope of the cosine timing wave (or the negative slope in the negative group of thyristors). Any attempt to do so in the naturally commutated cycloconverter would lead to commutation failure. Double-edged PWM of the form shown in figure 3.4b is therefore not possible. However a form of double-edged PWM can be obtained by using two timing waves and omitting the sampling epochs on their positive slopes. This is a method referred to by Panter(17) and is shown in figure 3.4c. It is seen that the train of PWM pulses corresponds exactly to the thyristor switching functions which were discussed in the previous section (section 3.2).

Thus it is shown that the operation of the cycloconverter can be identified with naturally-sampled PWM with either trailing-edge modulation or a form of double-edged modulation.
Application of Regular Sampling to the Control of the Circulating-current Cycloconverter

(3.4.1) General

It has been shown in section 3.3 that conventional cosinusoidal control of a cycloconverter can be identified with natural sampling of the reference wave. Natural sampling was seen to be a process in which the sampling instants coincide with the PTM train of impulses, and with the modulated edges of the PWM train of rectangular pulses, which in turn, define the thyristor switching instants. The sampling intervals are not equal but depend on the amplitudes of the samples of the modulating signal (that is, the cycloconverter reference wave).

This interpretation of natural sampling is also given by Pantor(17) who then proceeds to a definition of 'uniform' sampling as a process of sampling in which the widths of the PWM pulses are dependent on the amplitude of the modulating signal at uniformly spaced sampling instants. The term 'regularly sampled' is often used in communications engineering synonymously with 'uniformly sampled', and the term 'regular' sampling will be used in this thesis.

The resulting difference between natural and regular sampling when applied, for example, to trailing-
edge PWM is that with natural sampling, the sampling instant is at the end of the pulse, whilst with regular sampling, the sampling instant is at the beginning of the pulse. In the cycloconverter, therefore, the sampling instants for regular sampling are at the starts of the cosine timing waves.

Sampling at regular intervals is used in the majority of applications in communications engineering (18). The reason for this is based on the principle that with irregular sampling some samples will be close to each other giving much the same information, whilst others will be separated by a wider interval resulting in the possibility of error. Regular sampling, on the other hand, is credited with greater accuracy and less distortion than natural sampling. Later considerations in this thesis will be concerned with assessing whether these advantages of regular sampling also apply to the cycloconverter.

Sample-and-hold is another process used in communications engineering practice when a sample, which is evaluated instantaneously, is required to be retained for a certain period of time. The two processes of regular sampling and sample-and-hold can be combined and applied to cosinusoidal control of the cycloconverter as an alternative to natural sampling. It is convenient to
sample at the starts of the cosine timing waves which occur at regular intervals, and the sampled values are then held constant and compared with the instantaneous values of the associated cosine timing waves as shown in figure 3.5. The PTM impulses, or the trailing edges of the PWM pulses, then mark the instants when the held samples and the associated timing waves are equal, corresponding also to the thyristor switching instants.

As with natural sampling, regularly sampled double-edged PWM gives a train of rectangular pulses which are identical to the thyristor switching functions. The switching functions for regular sampling will be different from those for natural sampling resulting in slight differences in the output voltage waveforms. However, since each switching function is derived from samples of the same reference wave having the required output frequency of the cycloconverter, the output voltage in each case has a predominant component at this frequency.

(3.4.2) **Significance of the Sampling Theorem**

It is seen from the foregoing that the application of sampling techniques to the cycloconverter involves sampling the reference wave, converting these amplitude-modulated samples to time-modulated pulses and then using these pulses to synthesize the output waveform.
FIG 3.5 COSINUSOIDAL CONTROL WITH REGULAR SAMPLING OF THE REFERENCE WAVE
from sections of the input voltage waveforms. The objective is to fabricate a waveform which, as closely as possible, has the form of the sampled wave, that is a sinusoidal waveform of wanted output frequency. The identification of PWM and the switching functions are introduced as intermediate steps in the process of synthesizing the output voltage waveform.

A basic theorem in communications signal processing is the sampling theorem, which is applicable to regular sampling. This states that, if the samples of a given analogue signal are to carry sufficient information to enable the signal to be reconstituted, the rate of sampling must be equal to at least twice the highest frequency of the significant components present in the original signal. A rate of sampling exactly equal to twice the highest frequency is called the Nyquist Rate of Sampling.

This theorem can be applied to the cycloconverter in which the sampled signal is the sinusoidal reference wave having a frequency equal to the wanted output frequency, $f_0$.

$$\therefore \text{Nyquist sampling rate} = 2f_0 \quad (3.2)$$

Each sample of the reference wave is associated with one cosine timing wave, and therefore the actual sampling rate is equal to the number of cosine timing waves per second. The number of cosine timing waves
per second is the product of the pulse number \((p)\) of the cycloconverter circuit and the input frequency \((f_1)\).

\[
\therefore \text{Actual sampling rate} = pf_1
\]

(3.3)

Since the Nyquist sampling rate is the minimum permissible sampling rate, then from expressions (3.2) and (3.3),

\[
\begin{align*}
\text{pf}_1 & \geq 2f_0 \\
\text{or } f_0 & \leq \frac{pf_1}{2}
\end{align*}
\]

(3.4)

This gives the theoretical maximum output frequency of the cycloconverter.

Taking, as an example, a 3-pulse cycloconverter with a 50Hz input frequency, the theoretical maximum output frequency is therefore given by

\[
\frac{3 \times 50}{2} = 75 \text{ Hz}
\]

In practice, the maximum output frequency is considerably less than 75 Hz. The reason for this difference is that the Sampling Theorem assumes that the analogue signal is to be reconstituted from the samples using an ideal filter or demodulator. The cycloconverter clearly does not act as an ideal filter or demodulator since the output voltage waveform is not an exact replica of the input reference voltage which is sinusoidal.

This then leads to the conclusion that the reference wave must be sampled at a rate higher than the
Nyquist Rate in order that it can be reconstituted with acceptable distortion. It is not possible, however, to make a quantitative conclusion for the minimum practical sampling rate since this is dependent upon the load and on the acceptable levels of distortion at the various distortion frequencies. The acceptable levels of distortion are generally based on subjective or empirical considerations rather than theoretical ones.

From expression (3.4) it is seen that the theoretical minimum value of the $p_{f_1}/f_0$ Ratio of the cycloconverter is 2.0. It was shown in section 2.6.4 that the magnitude of the $p_{f_1}/f_0$ Ratio is significant in terms of the fundamental repetition frequency of the output voltage waveform. For example, when the value of the $p_{f_1}/f_0$ Ratio is an integer, the fundamental frequency is equal to the wanted output frequency and there are then no sub-harmonic components in the output waveform. In terms of sampling, this corresponds to the situation where the trains of amplitude-modulated samples (and of the corresponding PWM pulses) of consecutive cycles of the reference wave are identical. This is only possible when there are an integer number of cosine timing waves in one cycle of the reference wave.

When there is a non-integer number of cosine timing waves in one cycle of the reference wave, the trains of amplitude-modulated samples of consecutive cycles of the reference wave are not identical. As
discussed in section 2.6.4, the fundamental repetition frequency is then less than the wanted output frequency. Although this indicates that a sub-harmonic component may be present in the output voltage waveform, no indication is given of its magnitude relative to the wanted output component. The magnitudes of the sub-harmonic components, and the relative effects of natural and regular sampling, were investigated using a digital computer as described in the next section.

(3.4.3) Spectral Analysis by Computer Investigation

A computer program simulating the circulating-current cycloconverter was written in Fortran IV language and run initially on an IBM 1130 digital computer and subsequently on an ICL 4-75 computer. The primary purpose of the program, a copy of which is shown in Appendix 1, was to compute the magnitudes of the sub-harmonic components of the output voltage, together with a limited range of the low-order super-harmonic components.

The first part of the program computes the instants at which the cosine timing waves and the samples of the reference wave are equal. These are the switching instants of the thyristor switching functions (as discussed in section 3.2), and are the starts and finishes of segments of the input supply waveforms from which the output voltage waveform is fabricated. The computed
switching instants are then used in the second part of
the program for computation of the magnitudes of the
harmonic components of the output voltage waveform by
Fourier Analysis. Appendix 2 shows the adaptation of the
basic Fourier Expressions for the computer program.
The fundamental repetition frequency was determined by
the method described in section 2.6.3.

From the Fourier coefficients, the magnitude
of the wanted and the distortion components are computed
for the positive and negative groups. The mean value
at each frequency gives the magnitude of each component
for the complete cycloconverter.

The program was arranged to run with natural and
regular sampling for a range of output frequencies as
specified in the input data.

Table 3.1 shows the computed magnitudes of the
components of the output voltage waveform of a 50 Hz
3-pulse circulating-current cycloconverter
for natural sampling. The modulation factor is 0.9.
The magnitudes are expressed as percentages of the wanted
component, and those components which are less than 1%
of the wanted component are ignored.

As discussed in chapter 1, high frequency
components are of less importance than those which are near,
and particularly below, the wanted output frequency.
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**TABLE 3.1** 3-PULSE 50Hz CIRCULATING-CURRENT CYCLOCONVERTER: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT, NATURAL SAMPLING
The tabulated results are therefore limited to a maximum component frequency of 70 Hz. The results for natural sampling, that is normal cosinusoidal control, are very similar to those given by Pelly\textsuperscript{(10)}, and Ford\textsuperscript{(12)}, the differences being less than 1.5\% of the magnitude of the wanted component.

It was shown in the previous section and in section 2.6.4 that if the value of the $\text{pf}_1/f_0$ Ratio of the cycloconverter is an integer, then no sub-harmonic can exist in the output voltage waveform. In table 3.1, this is confirmed by the results for output frequencies of 5, 10, 15, 25, 30 and 50 Hz all of which correspond to integer values of the $\text{pf}_1/f_0$ Ratio. Output frequencies of 35, 40 and 45 Hz correspond to non-integer values of the $\text{pf}_1/f_0$ Ratio giving theoretically-present sub-harmonic components, and indeed are shown in the tabulated results to give significant sub-harmonic components of magnitude up to almost 10\% of the wanted component.

The tabulated results also show that a significant low-frequency super-harmonic is present when the output frequency is 40 Hz or 45 Hz. In both cases, its magnitude is 31\% of the magnitude of the wanted component. Although, as stated in chapter 1, the super-harmonics have less effect on the torque and efficiency of an a.c. motor than the sub-harmonics, these 31\% super-harmonics are of sufficiently high magnitude as also tending to be
responsible for imposing a limit on the practical output frequency of a cycloconverter.

Table 3.2 shows the results for regular sampling corresponding to table 3.1. A comparison of these results shows that the 10 Hz sub-harmonic components at wanted output frequencies of 35 Hz and 40 Hz, and the 30 Hz sub-harmonic component at 45 Hz wanted output frequency are attenuated from about 9% of the wanted component for natural sampling to less than 1% for regular sampling. The 20 Hz and 15 Hz sub-harmonic components occurring at wanted output frequencies of 40 Hz and 45 Hz respectively, are also attenuated to less than 1% of the wanted component by employing regular sampling. However, similar significant improvements are not obtained with the high magnitude 70 Hz and 60 Hz super-harmonic components which occur at wanted output frequencies of 40 Hz and 45 Hz respectively.

Figure 3.6a compares the magnitude of the wanted components for natural and regular sampling. The theoretical maximum output voltage is \( \frac{2\sqrt{2}}{2\pi} (= 0.825) \) of the input phase voltage. The modulation factor for the computed results is 0.9 and therefore the theoretical output voltage is \( 0.9 \times 82.5 = 74.3\% \) of the input voltage. For natural sampling, the computed output voltage is 73.2% of the input voltage for all output frequencies except 30 Hz. At this frequency, there is a rise in voltage which, as shown in figure 3.6a, is then negated by
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**Table 3.2** 3-Pulse 50Hz Circulating-Current Cycloconverter: Magnitudes of the Low-Frequency Components as Percentages of the Wanted Component. Regular Sampling
FIG 3-6 MAGNITUDE AND PHASE ANGLE OF THE WANTED COMPONENT (CIRCULATING-CURRENT CYCLOCONVERTER) NATURAL & REGULAR SAMPLING
regular sampling. Figure 3.6a also shows that regular sampling causes voltage attenuation which increases with wanted output frequency. However, in relation to the voltage, this attenuation is insignificant.

The phase angles between the wanted components of the output voltage and the reference voltage for natural and regular sampling are shown in figure 3.6b as functions of wanted output frequency. For natural sampling, the wanted component is in phase with the reference voltage, except at 30 Hz when there is a phase-shift of $+7^\circ$. With regular sampling, however, a phase-shift is introduced for all wanted output frequencies, increasing linearly from $0^\circ$ at 0 Hz to $-90^\circ$ at 50 Hz.

(3.4.4) Effect of Regular Sampling on the Thyristor Switching Instants

The computed switching instants for the positive group of a 50 Hz 3-pulse circulating-current cycloconverter are shown as firing angles in figures 3.7a and 3.7b for 35 Hz output frequency at 0.9 modulation factor for natural sampling and regular sampling respectively. These figures show the oscillations of the firing angles about the 'quiescent' angle, $90^\circ$, and a comparison between them does not indicate any significant differences between the effects of employing natural or regular sampling. However, when the differences between successive firing angles are plotted as shown in figures 3.8a and 3.8b, it is seen that
FIG 3-7 FIRING ANGLES FOR ONE REPETITION CYCLE OF WANTED OUTPUT COMPONENT OF THE CIRCULATING-CURRENT CYCLOCONVERTER (POSITIVE GROUP)
FIG 3.8 DIFFERENCES BETWEEN SUCCESSIVE FIRING ANGLES IN FIGURES 3.7a & 3.7b
natural sampling results in very uneven differences between successive firing angles, whereas when regular sampling is employed, the differences are not so great, and are dispersed more evenly in positive (firing angle increasing) and negative (firing angle decreasing) directions.

The output frequency for figures 3.7 and 3.8 is 35 Hz which corresponds to a non-integer $\frac{p_{f_1}}{f_0}$ Ratio value. In section 3.4.2 it was shown that if the $\frac{p_{f_1}}{f_0}$ Ratio has a non-integer value (that is, if there is a non-integer number of cosine timing waves in one cycle of the reference wave), a sub-harmonic component may be present in the output voltage waveform. However, the results in tables 3.1 and 3.2 show that when the output frequency is 35 Hz (corresponding to a non-integer $\frac{p_{f_1}}{f_0}$ Ratio value), there is a sub-harmonic component of significant magnitude when natural sampling is employed, but when natural sampling is replaced by regular sampling, the sub-harmonic component is eliminated.

Thus, although as discussed in section 3.4.2, a sub-harmonic component cannot exist if the $\frac{p_{f_1}}{f_0}$ Ratio has an integer value, the results show that this is not the exclusive criterion for ensuring that no sub-harmonic components will be present. It is shown in figures 3.7 and 3.8 that an alternative criterion can involve an appropriate dispersion of the thyristor switching
instants for successive cycles of the reference wave. This is now given further consideration.

(3.5) Effect of Sub-harmonic Components on the Output Voltage Waveform

(3.5.1) General

If the cycloconverter is operated with constant firing angles, it acts as a rectifier; with 90° firing angles the output voltage is zero, and with constant firing angles less than or greater than 90°, the output voltage would be positive or negative respectively. With cycloconverter action, the firing angles oscillate on either side of 90° (within the range 0 - 180°) at a frequency equal to the output frequency. It is possible for both the rectifier and the cycloconverter actions to occur simultaneously: if the firing angles under cycloconverter action tend to be biased towards one side of 90°, as shown in figure 3.9a, then a d.c. component will be present in the output voltage. If this bias itself oscillates (at a frequency less than the wanted output frequency) as shown in figure 3.9b, then the polarity of the d.c. component will alternate; that is, a sub-harmonic component will be present in the output voltage.

It was seen in the preceding section that appropriate control of the thyristor switching instants can eliminate the sub-harmonic component even if this component is theoretically present. This conclusion is now considered with respect to the shape of the output voltage waveform.
FIG 3.9 EFFECT OF BIASED FIRING ANGLES ON THE OUTPUT VOLTAGE WAVEFORM
The Equal-area Criterion

The mean (d.c.) value of any waveform $f(t)$ between times $t_1$ and $t_2$ is equal to

$$\frac{1}{t_2 - t_1} \int_{t_1}^{t_2} f(t) dt \quad (3.5)$$

If the function is periodic and if $(t_2 - t_1)$ is the period $T_f$ of the repetition frequency, then this expression is identical to the expression for the d.c. component when Fourier Analysis is applied to a periodic function.

Furthermore, if $f(t)$ is the output voltage of a cycloconverter, expression (3.5) gives the d.c. component of the output voltage. Substituting $T_f$ for $(t_2 - t_1)$, and setting the time-axis to give $t_1 = 0$, expression (3.5) then becomes

$$\text{d.c. component} = \frac{1}{T_f} \int_{0}^{T_f} f(t) dt \quad (3.6)$$

If, however, $(t_2 - t_1)$ is the period $T_0$ of the wanted output frequency of the cycloconverter, then expression (3.5) becomes

$$\frac{1}{T_0} \int_{0}^{T_0} f(t) dt \quad (3.7)$$

Expression (3.7) gives the d.c. component of the cycloconverter output voltage only if the wanted output frequency of the cycloconverter is
equal to the repetition frequency of the voltage waveform. This is the condition corresponding to figure 3.9a when the wanted output component is the fundamental component of the Fourier series which defines \( f(t) \). If the fundamental repetition frequency is less than the wanted output frequency, then expression (3.7) only gives the mean value of the output voltage during the period \( T_0 \). Since successive cycles of the output voltage waveform, at wanted output frequency, are not identical for this condition, the mean value for each cycle will not be a constant, as was shown in figure 3.9b.

It therefore follows that, if there is a variation in the mean value of successive cycles of the output voltage waveform at wanted output frequency, a sub-harmonic component is present. The fundamental repetition frequency is then less than the wanted output frequency.

When comparing the relative magnitudes of the mean values of successive cycles of the output voltage, it is only necessary to consider the quantity \( \int_{0}^{T_0} f(t) \, dt \) since the component \( \frac{1}{T_0} \) of expression (3.7) is common. This integral gives the nett area enclosed by the output voltage waveform during the period \( T_0 \) seconds.

Although the above discussion refers to the output voltage waveform of the cycloconverter, it applies equally to the output current waveform.
It is concluded that in order to suppress a sub-harmonic component, the 'equal-area criterion' must be applied stipulating that the nett areas enclosed by successive cycles of the output waveform at wanted output frequency must be arranged to be equal, by appropriate control of the thyristor switching instants.

(3.5.3) Correlation between the Equal-area-Criterion and the Computer Results

Figure 3.10 shows the output voltage waveform (derived from the computed switching instants) of the positive group of a 50 Hz 3-pulse circulating-current cycloconverter with 35 Hz output frequency and 0.9 modulation factor. The area enclosed by the waveform is shown in black, and the wanted output component is also shown in the figure. By comparing the nett areas for each successive cycle of the wanted component, a sub-harmonic component is identifiable. The computer results in table 3.1 showed that the magnitude of this component is 9% of the magnitude of the wanted component.

Figure 3.11 shows the output voltage waveform for regular sampling corresponding to figure 3.10. In this case, no variation in the nett areas for successive cycles of the wanted component is discernible, showing that no significant sub-harmonic component is present. This was confirmed by the computer results in table 3.2. Thus, the modifications to the switching instants caused
FIG 3:10 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF CIRCULATING-CURRENT CYCLOCONVERTER (POSITIVE GROUP).

NATURAL SAMPLING
FIG 3-11 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF CIRCULATING-CURRENT CYCLOCONVERTER (POSITIVE GROUP).

REGULAR SAMPLING
by replacing natural sampling with regular sampling result in small, but highly significant, modifications to the output voltage waveform such that a balance of the enclosed areas of successive cycles of the wanted output component is obtained. It is believed that this is the basic requirement for attenuating the sub-harmonic component.

(3.6) **Inhibition Control**

The discussion so far in this chapter has been primarily concerned with the operation of the cycloconverter in the circulating-current mode. Since each thyristor group is in continuous conduction, the output voltage waveform of each group is continuous. This provides a convenient basis for the investigation into the attenuation of the sub-harmonic components of the output voltage waveform of the cycloconverter when operated in the inhibited mode.

During the conduction periods of each thyristor group of the inhibited cycloconverter, the thyristor switching instants are determined by cosinusoidal control as in the circulating-current cycloconverter. During the non-conducting periods when inhibition control is applied, the output voltage of the non-conducting thyristor group is zero. Thus, the two thyristor groups are effectively switched alternately at the output frequency \( w_0 \) of the cycloconverter. This leads to the concept of inhibition switching functions of unity.
magnitude as shown in figure 3.12. If the inhibition switching function for one group is \( f(\omega_0 t) \), then the inhibition switching function for the other group is \( f(\omega_0 t - \pi) \). The modulation of these functions with the respective group output voltages (for circulating-current mode), results in two modulated waves which, when combined, give the output voltage waveform of the inhibited cycloconverter.

As discussed in the preceding section, the presence of a sub-harmonic component in the output voltage waveform has the effect of varying the nett areas enclosed by successive cycles of this waveform at the wanted output frequency. In the inhibited cycloconverter, however, the variations of the nett areas are under the influence of not only modifications to the thyristor switching functions, but also modifications to the inhibition switching functions.

Transfer of control from one group to the other in the inhibited cycloconverter can only take place when the load current is zero. Thus each transition of the inhibition switching function is associated with a zero magnitude sample of load current, the continuously varying load current between the samples having no significance. Since the inhibition switching instants are dependent on the load current zeros, there is no possibility of exercising direct control over the inhibition switching instants. However, there still
Fig 3:12 Inhibition switching functions of the output voltage waveform of an inhibited cycloconverter

1. Positive group output voltage waveform
2. Positive group inhibition switching function
3. Negative group output voltage waveform
4. Negative group inhibition switching function
5. Output voltage waveform of inhibited cycloconverter

\[ \frac{1}{1} \times \frac{2}{2} + \frac{3}{3} \times \frac{4}{4} \]
remains the possibility of exercising indirect control over them by modifying the thyristor switching instants by, for example, replacing natural by regular sampling of the reference voltage.

(3.7) Application of Regular Sampling to the Control of the Inhibited Cycloconverter

(3.7.1) General

The computer results for the circulating-current cycloconverter, discussed in section 3.4.3, showed very significant attenuation of the sub-harmonic components of the output voltage waveform when natural sampling was replaced by regular sampling. In the inhibited cycloconverter, the output voltage waveform of the circulating-current cycloconverter is modulated with the inhibition switching functions, and it is necessary to reinvestigate the effect of regular sampling on the harmonic spectra.

For the initial investigation, it is assumed that the load current waveform is sinusoidal so that the load current zeros, which define the inhibition switching instants, occur at regular intervals. It is further assumed at this stage that the load has unity displacement factor so that the inhibition switching instants are also defined by the zeros of the wanted component of the output voltage. However, since regular sampling in the circulating-current
cycloconverter was shown to introduce a phase shift between the wanted output component and the reference voltage, then a similar effect is anticipated in the inhibited cycloconverter. In this case, there is a phase difference between the inhibition switching functions and the reference wave.

(3.7.2) Computer Program for Spectral Analysis of the Inhibited Cycloconverter

A computer program was written to compare the effects of natural and regular sampling on the harmonic spectrum of the output voltage waveform of the inhibited cycloconverter, and also to compare the spectrum of the inhibited cycloconverter with that of the circulating-current cycloconverter.

The first part of the program computes all the switching instants of the thyristor switching functions which would apply to the cycloconverter if operated in the circulating-current mode.

The second part of the program identifies the instants at which the load current is zero, thereby defining the inhibition switching functions. The phase difference between the reference wave and the inhibition switching function cannot be determined until the phase difference between the wanted component of the output voltage and the reference wave is known. This is to be determined by Fourier Analysis of the output voltage.
waveform - but this requires the inhibition switching instants to be known already.

Thus, in order that the computation may proceed, the phase angle between the wanted component and the reference wave must be estimated and included in the input data. The program then uses this angle to compute the inhibition switching instants. This enables those thyristor switching instants which are relevant to the inhibited cycloconverter to be identified, and Fourier Analysis (Appendix 2) is applied to the resulting output voltage waveform to determine the phase angle of the wanted component with respect to the reference wave. If the difference between this computed angle and the estimated angle is greater than 0.4°, then an iterative procedure amends the estimated value to the mean of the estimated and computed values. When the difference is less than 0.4°, the full Fourier Analysis procedure is applied to the output voltage waveform to compute the magnitudes of all the wanted and distortion components.

(3.7.3) Effect of Regular Sampling on the Performance of the Inhibited Cycloconverter

The computed magnitudes of the components of the output voltage for natural and regular sampling are shown in tables 3.3 and 3.4. A comparison of tables 3.3 and 3.1 shows that for natural sampling, the inhibited cycloconverter
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**TABLE 3.3** 3-PULSE 50HZ INHIBITED CYCLOCONVERTER:
MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS
AS PERCENTAGES OF THE WANTED COMPONENT,
NATURAL SAMPLING
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**TABLE 3.4.** 3-PULSE 50HZ INHIBITED CYCLOCONVERTER:
MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS
AS PERCENTAGES OF THE WANTED COMPONENT.
REGULAR SAMPLING
generates higher magnitudes of both sub- and super-harmonic components than those in the circulating-current cycloconverter. In addition, the number of significant (greater than the arbitrary figure of 1%) components is greater. Improvement of the performance of the cycloconverter therefore presents a more onerous problem when it is operated in the inhibited mode than in the circulating-current mode.

A comparison of tables 3.1 - 3.4 also shows that replacement of natural by regular sampling in the inhibited cycloconverter does not result in the significant improvements that were obtained in the circulating-current cycloconverter. For example, the 10 Hz sub-harmonic components at 35 Hz and 40 Hz wanted output frequencies are attenuated from about 16% to only about 12\%\%\%. A more significant reduction is, however, obtained from 10% to 4% in the case of the 20 Hz sub-harmonic at 40 Hz wanted output frequency.

The variation of the magnitude of the wanted output component with output frequency for natural and regular sampling is shown in figure 3.13a. As in the circulating-current cycloconverter, the output voltage for natural sampling is virtually independent of the output frequency. The rise in voltage again occurs at a frequency of 30 Hz, but the rise is not so significant as in the circulating-current cycloconverter. Figure 3.13a also shows that, when regular sampling is
NATURAL & REGULAR SAMPLING
employed in the inhibited cycloconverter, the output voltage becomes attenuated. This attenuation increases with frequency, reaching about 10% when the output frequency is equal to the input frequency.

Figure 3.13b shows that the phase angle of the wanted component with respect to the reference voltage is virtually zero at all output frequencies when natural sampling is employed in the cycloconverter, but increases (in the lagging sense) with output frequency for regular sampling.

The output voltage waveforms for 35 Hz output frequency are shown in figures 3.14 and 3.15 for natural and regular sampling respectively. In both cases, the nett area enclosed by the waveforms varies between successive cycles of wanted component, due to the presence of a sub-harmonic component. Tables 3.3 and 3.4 show that the predominant sub-harmonic component has a frequency of 10 Hz and its magnitude is 15% of the wanted component for natural sampling, and 12% for regular sampling.

The enclosed areas of the waveforms during the positive half cycles of the wanted component are exactly the same as the corresponding periods in figures 3.10 and 3.11 for the positive group of the circulating-current cycloconverter. The enclosed areas during the negative half cycles are also exactly the same as the corresponding periods for the negative group of the circulating-current
FIG 3.14 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF INHIBITED CYCLOCONVERTER. UDF LOAD, SINUSOIDAL CURRENT; NATURAL SAMPLING
FIG 3-15 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF INHIBITED CYCLOCONVERTER.
UDF LOAD, SINUSOIDAL CURRENT; REGULAR SAMPLING
cycloconverter. But although regular sampling virtually eliminates the variation in nett areas during successive cycles of the wanted component for both the positive and negative groups of the circulating-current cycloconverter, when segments of these waveforms are put together in the inhibited cycloconverter, the variation again becomes evident.

It is concluded that the introduction of the inhibition switching functions in the operation of the cycloconverter have a negating effect on the improvements obtained with regular sampling when operated in the circulating-current mode.

(3.7.4) Application of Pulse Position Modulation

An extension of the process of regular sampling in communications engineering is described by Taub and Schilling (18). This is pulse position modulation, and involves adding ramp functions of constant slope to the held samples and comparing the resultant functions with a d.c. reference level. At instants of intersections, short duration pulses are generated. By this means, the original analogue signal is converted to a train of digital time-modulated samples (referred to as pulse position modulation).

The process of sampling in the cycloconverter is to convert the sinusoidal reference wave to switching instants. Pulse position modulation is therefore
applicable to the control of the cycloconverter, with appropriate modifications, as shown in figure 3.16. Cosine timing waves are used in place of the d.c. reference level, and the ramp functions have durations equal to the timing waves. The ramps are added to the held samples of the reference wave in the positive group of the cycloconverter (as in figure 3.16), but are inverted and subtracted in the negative group. The intersections of the cosine timing waves with the ramp functions give the switching instants of the thyristor switching functions.

In order to investigate the effect of applying pulse position modulation to the control of the inhibited cycloconverter, the first part of the computer program was modified to give a different set of values of the thyristor switching instants. The computed magnitudes of the output voltage components (as percentages of the wanted component) are shown in table 3.5. By comparing these results with the results shown in table 3.3, it is seen that the sub-harmonic components are significantly attenuated when pulse position modulation is employed in place of natural sampling. For example, the 10 Hz sub-harmonic components at output frequencies of 35 and 40 Hz are attenuated from about 16% to 6% of the wanted component. This is compared with an attenuation down to only about 12 1/2% with regular sampling (table 3.4). The 20 Hz sub-harmonic component for 40 Hz output frequency, and the 15 Hz sub-harmonic
FIG 3.16 PULSE POSITION MODULATION FOR THE POSITIVE GROUP OF THE CYCLO-CONVERTER
### Table 3.5: 3-Pulse 50Hz Inhibited Cycloconverter

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**Table 3.5** 3-Pulse 50Hz Inhibited Cycloconverter: Magnitudes of the low-frequency components as percentages of the wanted component. Pulse position modulation.
component for 45 Hz output frequency, are attenuated to about 2% with pulse position modulation, compared with 3% when regular sampling is applied.

The voltage and phase angle of the wanted output component are shown in figure 3.17 as functions of wanted output frequency. The voltage is maintained between 76% and 77% of the input voltage which is an improvement over the 73% for natural sampling (figure 3.13). The phase angle with respect to the reference voltage increases in a lagging direction with increasing output frequency reaching 60° at 50 Hz compared with 82° at 50 Hz for regular sampling (also shown in figure 3.13).

Figure 3.18 shows the areas enclosed by the output voltage waveform when pulse position modulation is applied. In spite of the attenuation of the sub-harmonic components by employing pulse position modulation, the variation in nett areas during successive cycles of the wanted component is still evident, showing that even a 6% sub-harmonic has a noticeable effect on the output waveform.

PULSE POSITION MODULATION
FIG 3:18 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF INHIBITED CYCLOCONVERTER. UDF LOAD, SINUSOIDAL CURRENT; PULSE POSITION MODULATION
CHAPTER 4

IDENTIFICATION OF THE SUB-HARMONIC COMPONENTS IN THE INHIBITED CYCLOCONVERTER

(4.1) Introduction

The last part of the previous chapter is concerned with the sub-harmonic components of the output voltage of an inhibited cycloconverter on the assumption that the cycloconverter supplies sinusoidal load current. This is an assumption invariably made in the literature, and there is no evidence to show that the effect of a non-sinusoidal current has been considered in detail. Furthermore, there is no evidence to show that attenuation of the sub-harmonic components in the inhibited cycloconverter has been investigated.

In this chapter, the characteristics of the sub-harmonic components of the output voltage of the inhibited cycloconverter are investigated in greater detail than in the last chapter. The effect of a non-sinusoidal load current at a non-unity displacement factor is also discussed. Finally, a method of attenuating the sub-harmonic components under these operating conditions is proposed and investigated.

(4.2) Identification of the Sub-harmonic Components in terms of the General Harmonics

The waveform analysis of a multi-phase cycloconverter is stated by Takahashi and Miyairi (11) to

87
be extremely difficult and exact analysis to be almost impossible. To facilitate this problem, their analysis involves the replacement of cycloconverter operation by an equivalent mechanical commutator and identifying the constituent harmonics in the resulting frequency-modulation expression. Similar results are obtained by Pelly\textsuperscript{(10)}, whose analysis is based on the switching function concept discussed in chapter 3, giving the same constituent harmonics in the output waveform as are identified by Takahashi and Miyairi.

The frequencies of the harmonic components present in the output voltage of a 3-pulse inhibited cycloconverter are shown by Pelly\textsuperscript{(10)} to be given by the following general relationships:

\[
\begin{align*}
\mathbf{f}_h &= \left| 3(2m-1)f_i \pm 2nf_0 \right| \quad (4.1) \\
\text{and} \quad \mathbf{f}_h &= \left| 6mf_i \pm (2n+1)f_0 \right| \quad (4.2)
\end{align*}
\]

where \( m = \) any integer from 1 to infinity  
\( n = \) any integer from 0 to infinity  
\( f_i = \) input frequency  
\( f_0 = \) output frequency

The frequencies of the components for \( m = 1 \) and \( n = 0, 1, 2, 3, 4 \), using expressions (4.1), are:

\[
\begin{align*}
\mathbf{n} = 0 & \quad \mathbf{f}_h = 3f_i \\
\mathbf{n} = 1 & \quad \mathbf{f}_h = 3f_i \pm 2f_0 \\
\mathbf{n} = 2 & \quad \mathbf{f}_h = 3f_i \pm 4f_0 \\
\mathbf{n} = 3 & \quad \mathbf{f}_h = 3f_i \pm 6f_0 \\
\mathbf{n} = 4 & \quad \mathbf{f}_h = 3f_i \pm 8f_0
\end{align*}
\]
\[
\begin{align*}
n = 0 & \quad f_h = 6f_1 \pm f_o \\
n = 1 & \quad f_h = 6f_1 \pm 3f_o \\
n = 2 & \quad f_h = 6f_1 \pm 5f_o \\
n = 3 & \quad f_h = 6f_1 \pm 7f_o \\
n = 4 & \quad f_h = 6f_1 \pm 9f_o
\end{align*}
\tag{4.4}
\]

and for \( m = 2 \) in expression (4.1):
\[
\begin{align*}
n = 0 & \quad f_h = 9f_1 \\
n = 1 & \quad f_h = 9f_1 \pm 2f_o \\
n = 2 & \quad f_h = 9f_1 \pm 4f_o \\
n = 3 & \quad f_h = 9f_1 \pm 6f_o \\
n = 4 & \quad f_h = 9f_1 \pm 8f_o
\end{align*}
\tag{4.5}
\]

The above components represent only a small number of the infinite number of components in the frequency spectrum of the 3-pulse inhibited cycloconverter. The presence of these components is also confirmed by Takahashi and Miyairi (11).

For 6-pulse operation, the components shown in expressions (4.3) and (4.5) would not be present, the remaining components then being on either side of \( 6f_1 \), \( 12f_1 \) etc.

Thus, for a \( p \)-pulse inhibited cycloconverter, the frequencies of the harmonic components are given by
\[
f_h = |mp f_1 \pm (2n + k)f_o|
\tag{4.6}
\]
where \( k = 0 \) for odd values of \( (mp) \)
\[
= 1 \quad \text{for even values of \( (mp) \)}
\]

If the alternative sign in this expression is positive, then \( f_h > f_o \) and the component cannot be a sub-harmonic. Therefore all sub-harmonic components must have the general form:
expression (4.7) can be re-arranged as follows:

\[
\frac{f_h}{f_0} = \left| \frac{m}{p} \right| - \left( \frac{2n + k}{p} \right) \frac{f_0}{f_i}
\]  

Expression (4.7) can be re-arranged as follows:

\[
f_h = \left| \frac{m}{p_i} \right| - \left( \frac{2n + k}{f_0} \right) f_0
\]  

The factor \( \frac{p_i}{f_0} \) in this expression was referred to in chapter 2 as the 'Ratio' of the cycloconverter, and its numerical value expressed as a fraction was shown to be significant in determining the frequency of the fundamental component of the series of harmonic components which constitute the output voltage of the cycloconverter.

If the quantity \( \left| m \left( \frac{p_i}{f_0} \right) - (2n + k) \right| < 1.0 \), then the harmonic is a sub-harmonic component for the particular values of \( m, p, f_i, f_0, n \) and \( k \).

In order to proceed further towards the identification of the sub-harmonic components, it is convenient to consider the operation of the cycloconverter separately for integer and non-integer values of the \( \frac{p_i}{f_0} \) Ratio.

(4.2.1) Operation of the Cycloconverter at Integer \( \frac{p_i}{f_0} \) Ratio Values

For a component to be a sub-harmonic, its frequency is less than the output frequency. Thus, in expression (4.8), the coefficient of \( f_0 \) must be less than 1.0. \( m, n \) and \( k \) are integers, and since the

\[
f_h = \left| m\frac{p_i}{f_0} - (2n + k) \right| f_0
\]  

This does not mean, however, that all components represented by expression (4.7) are sub-harmonic components.
pf/f Ratio is also an integer for the present discussion, then the coefficient of f can only be zero or an integer.

It is therefore concluded that when the cycloconverter is operated at an integer value of the pf/f Ratio, there will be no alternating sub-harmonic components. There may however be a d.c. component if the coefficient of f in expression (4.8) is zero.

For a 3-pulse cycloconverter and m = 1, expression (4.8) becomes

$$f_h = \left| \frac{pf}{fo} - 2n \right| f_o$$ (4.9)

Therefore f is a d.c. component if the pf/f Ratio is an even-valued integer.

For a 3-pulse cycloconverter and m = 2, expression (4.8) becomes

$$f_h = \left| 2 \left( \frac{pf}{fo} \right) - (2n+1) \right| f_o$$ (4.10)

For any integer value of pf/f, the coefficient of f is never zero, so that in this case f can never be a d.c. component.

For a 3-pulse cycloconverter and m = 3, expression (4.8) becomes

$$f_h = \left| 3 \left( \frac{pf}{fo} \right) - 2n \right| f_o$$ (4.11)

In this case, f is a d.c. component if pf/f is an even integer.
It is concluded that a d.c. component will theoretically be present in the output voltage waveform of a 3-pulse cycloconverter when operated at an even-integer value of the pf$_1$/f$_0$ Ratio. Whether it is present in practice depends on its magnitude. When the 3-pulse cycloconverter is operated at an odd integer value of the pf$_1$/f$_0$ Ratio, a d.c. component cannot be present.

Similar arguments applied to the 6-pulse and 12-pulse cycloconverters show that in these cases, a d.c. component cannot be present for any integer value of pf$_1$/f$_0$.

In terms of the output voltage waveform, operation of a cycloconverter at any integer pf$_1$/f$_0$ Ratio corresponds to the situation in which the waveform is identical for successive cycles of the reference voltage. As discussed in chapter 2, the fundamental repetition frequency is then equal to the wanted output frequency, even if a d.c. component is present.

(4.2.2) **Operation of the Cycloconverter at Non-integer pf$_1$/f$_0$ Ratio Values**

It has been shown in chapter 2 that the fundamental repetition frequency can be determined from the fractional form of the pf$_1$/f$_0$ Ratio value; for any non-integer Ratio value, the fundamental repetition
frequency is less than the output frequency. This means that the output voltage waveform is not identical for successive cycles of the reference voltage.

It is also evident from expressions (4.9), (4.10) and (4.11), that for any non-integer value of \( pf_1/f_0 \), the coefficient of \( f_0 \) in these expressions is always a non-integer, and that for certain combinations of values of \( n \) and \( pf_1/f_0 \), the coefficient will be less than 1.0. In such cases, the component is a sub-harmonic.

(4.3) Computer Analysis

(4.3.1) Irregularity of the Inhibition Switching Functions at Low \( pf_1/f_0 \) Ratio Values

For decreasing values of \( pf_1/f_0 \) (or for increasing output frequency for a p-pulse cycloconverter operated at constant input frequency), the periods between the switching instants of the thyristor switching functions become increasingly larger proportions of the periods between switching instants of the inhibition switching functions. The 'pulsing' of the cycloconverter becomes more evident and it can no longer be assumed that the load current waveform is sinusoidal. The intervals between successive load current zeros, which define the inhibition switching instants, will not then necessarily be \( \frac{1}{2f_0} \) seconds, but will vary above
and below this value. The inhibition switching instants are also affected by modifications to the thyristor switching functions. This is shown in figure 4.1 in which it is assumed that the load has a lagging displacement factor and that the current does not fall to zero at the instant the voltage reaches zero. In figure 4.1a, the current falls to zero at time $t_1$ which therefore defines the inhibition switching instant. In figure 4.1b, a thyristor switching instant occurs just before $t_1$. The current does not then fall to zero at time $t_1$, but rises under the effect of the next phase voltage in sequence, and falls to zero at time $t_2$. The time interval $(t_2-t_1)$ may be a significant proportion of a half period of output frequency.

It is therefore clear that the transitions of the inhibition switching functions will become increasingly irregular as $pf_1/f_o$ decreases. It is then necessary to compute each instant at which the load current becomes zero in order to identify the nature of the inhibition switching functions. In the general case of a non-unity displacement factor load, the load current waveform between successive thyristor switching instants is the resultant of the forced and natural responses of the load to the corresponding phase voltage. During each of these periods, the output voltage waveform is a segment.
FIG. 4-1 MODIFICATION OF THE INHIBITION SWITCHING FUNCTION
of a sinusoid, and at each switching instant, there is a step change in voltage. The response of the load to the output voltage can therefore be analysed by considering each of these periods separately. Since the current is not necessarily zero at the switching instant, non-zero initial conditions must be taken into account.

The derivation of the expression for the current response of a load having lagging displacement factor is given in Appendix 3, and this was used as the basis for modifying the computer program to identify the inhibition switching instants. The initial condition for the first thyristor conduction period after an inhibition switching instant is zero; the initial condition for each of the other thyristor conduction periods is always given by the final condition of the previous conduction period. A step-by-step check on the instantaneous value of load current then enables the instants of the current zeros, and hence the inhibition switching instants, to be identified.

To avoid unstable switching between the two thyristor groups, load current zeros occurring before the associated reference voltage zero are ignored. Each inhibition switching instant therefore corresponds to the first current zero after the associated reference voltage zero.
(4.3.2) Computed Results

The computer program written for the inhibited cycloconverter when supplying non-sinusoidal load current is shown in Appendix 4. A lagging displacement factor load having a time constant of 0.005s was assumed; this is equivalent, for sinusoidal voltage and current, to a power factor of 0.67. The modulation factor was 0.9 and the program was run, initially, for natural sampling and for a range of values of the $pf_1/f_o$ Ratio from 9.0 down to 3.0. The second method of determining the fundamental frequency (described in chapter 2) was developed for use in this program.

The range of Ratio values from 9.0 to 3.0 correspond, for a 3-pulse cycloconverter, to a frequency range from 33$\frac{1}{3}$% to 100% of the input frequency. 33$\frac{1}{3}$% was selected as the lower limit since cycloconverters are commonly operated at frequency ratios up to this value, as discussed in chapter 1. The selection of 100% as the upper limit is arbitrary and there is no reason in theory why operation of the cycloconverter should not be possible at output frequencies greater than the input frequency.

The computed magnitudes of the components of the output voltage for the above input conditions are shown in Table 4.1. For convenience, the table is shown in 6 parts (Tables 4.1a - 4.1f). For each $pf_1/f_o$ Ratio value, the magnitudes are expressed as percentages of the wanted component, and the frequency of each component ($f_h$)
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TABLE 4.1a 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT. $p_f/f_i$ RATIO FROM 3.0 TO 4.0
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**TABLE 4.1b** 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT.

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**TABLE 4.1c** 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT. $f_d/f_e$ RATIOS 5.0 TO 6.0
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<td>1</td>
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<td>3</td>
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</tbody>
</table>

**TABLE 4.1a** 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT.  

**pf/f.< sub>c</sub> RATIOS 6.0 TO 7.0**
<table>
<thead>
<tr>
<th>Ratio</th>
<th>$f_0/f_1$</th>
<th>(\frac{(Frequency \ of \ component)}{f_0})</th>
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<tr>
<td></td>
<td>%</td>
<td>0.1</td>
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<td>7.9</td>
<td>38.0</td>
<td>5</td>
</tr>
<tr>
<td>8.0</td>
<td>37.4</td>
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**TABLE 4.1e** 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT.  
$P_0/f_0$ RATIOS 7.0 TO 8.0
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<th>0.3</th>
<th>0.4</th>
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<th>0.6</th>
<th>0.7</th>
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<th>1.2</th>
<th>1.3</th>
<th>1.4</th>
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**TABLE 4.1F** 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT. $p_f/f_i$ RATIO $8.0$ TO $9.0$
is expressed as a proportion of the wanted output frequency \((f_0)\). This enables the general harmonics (as defined by equation 4.8) to be conveniently identified, and the most significant ones are shown in the tabulated results by dotted lines. This clearly shows the variation in the frequency value of the general harmonic with variation in \(pf_1/f_0\) Ratio.

It is seen that each general harmonic maintains a fairly consistent magnitude over its range of specific frequencies. Variations in the magnitude are attributed to the coincidence with these general harmonics of less significant ones which are not identified in the tabulated results. The general harmonics also coincide with the wanted components at certain \(pf_1/f_0\) Ratio values which are all integer values. For example, when the \(pf_1/f_0\) Ratio is 5.0 there are no sub-harmonic components; but when the Ratio is reduced to 4.9 (representing an increase of output frequency of a 50 Hz 3-pulse cycloconverter from 30.0 Hz to only 30.6 Hz), a large number of sub-harmonic components appear with a very significant one at 0.9\(f_0\) or 27.54 Hz.

The fundamental frequencies in tables 4.1a - 4.1f are indicated by thick lines around the appropriate values, and no a.c. component can exist at frequencies below the fundamental frequency. However, it is seen that the general harmonics converge to a zero frequency, or d.c., component at a particular Ratio which can be referred to as the 'Centre Ratio' for that harmonic. For a range of Ratios on either side of the Centre Ratio, the harmonic is a sub-harmonic component with a frequency less than the output frequency. As the Ratios diverge further away from the Centre Ratio, so
the harmonic becomes a super-harmonic component.

It is seen from tables 4.1a to 4.1f, that for odd integer Ratio Values, there are no sub-harmonic or d.c. components in the output voltage waveform. For even integer Ratio Values, there is a d.c. component but no sub-harmonic component in the output. This confirms the conclusions drawn in section 4.2.

The presence of significant sub-harmonics can be predicted at Ratio Values other than those specifically quoted by interpolating between the tabulated results. For example, at a Ratio of 3.85, there would be no significant d.c. component, but there would be two significant sub-harmonic components, one at a frequency near 0.15 of the output frequency, and the other near 0.7 of the output frequency.

It is noted that the $3f_1$ harmonic family gives sub-harmonic components for a 2.0 range of Ratio Values, whilst the $6f_1$ harmonic family gives sub-harmonic components over a range of only 1.0. Thus the $3f_1$ harmonic family causes sub-harmonic distortion over a wider range of output frequencies than the $6f_1$ family.

The above conclusions drawn from tables 4.1a - 4.1f regarding the frequencies of the significant general harmonics and the sub-harmonic and d.c. components are summarized in figure 4.2. This figure shows the variations of the specific frequencies of the general harmonics (expressed as per unit of the output frequency)
FIG 4.2 VARIATION OF THE FREQUENCIES OF THE SIGNIFICANT GENERAL HARMONICS WITH THE $\frac{pf_i}{f_o}$ RATIO
with the $p_{f_1}/f_0$ Ratio. The characteristics shown with pecked lines are not extracted from the tabulated results, but are extrapolated from these results. The lower limit of the $p_{f_1}/f_0$ Ratio is set at 2.0 in figure 4.2 since this is the theoretical limit for operation of the 3-pulse cycloconverter determined by the Sampling Theorem (section 3.4.2). Although figure 4.2 does not give information on the magnitudes of the harmonics, the results in tables 4.1a - 4.1f show that they decrease for increasing coefficients of $f_0$.

Further data from the computer analysis showed that the $(3f_1-6f_0)$ and the $(6f_1-9f_0)$ harmonics in a 3-phase output each form co-phased sets, whilst the other harmonics form 3-phase sets. The $(3f_1-6f_0)$ and $(6f_1-9f_0)$ current harmonics can therefore be eliminated by using a 3-phase load with no neutral connection.

A comparison between tables 4.1a - 4.1f and table 3.3 shows that the magnitudes of the sub-harmonic components are increased when the unity displacement factor load taking sinusoidal current is replaced by an inductive load (0.005 s time constant) taking a non-sinusoidal current. For example, the magnitude of the $(3f_1-4f_0)$ harmonic is increased in the sub-harmonic region from about 15% to about 20% of the wanted component.

From the discussion in section 3.5 regarding
the equal-area criterion, this leads to the conclusion that the inhibition switching functions are responsible for increasing the variation in nett areas of successive cycles of the output voltage waveform at wanted output frequency. For the conditions under discussion in which the load current is non-sinusoidal, the mark/space ratios of the inhibition switching functions are generally irregular. The shape of the output waveforms were therefore investigated further using the digital computer, as described in the next section.

(4.3.3) **Computer Graph Plots**

A sub-routine program was written to obtain computer graph plots of the output voltage and current waveforms for selected values of the \( \frac{f_i}{f_o} \) Ratio. Examples are shown in figures 4.3 - 4.6 and they relate to the results in section 4.3.2; that is, they are for an inhibited cycloconverter operating with natural sampling at 0.9 modulation factor and supplying an inductive load of time constant 0.005 s. A 3-pulse cycloconverter is assumed, and in all cases, the waveforms represent one cycle of fundamental (repetition) frequency plus \( \frac{1}{2} \) cycle of wanted output frequency. The first and last half cycles (at output frequency) are therefore identical in each case. In order to limit the number of cycles of the wanted
FIG 4.3 COMPUTER GRAPH PLOT FOR $\beta_f$ RATIO = 4.0

OUTPUT VOLTAGE WAVEFORM, NATURAL SAMPLING

LOAD CURRENT WAVEFORM, NATURAL SAMPLING
FIG 4.4 COMPUTER GRAPH PLOT FOR $\frac{f_1}{f_2}$ RATIO = 4.5
FIG 4.5 COMPUTER GRAPH PLOT FOR $\frac{f_c}{f_o}$ RATIO = 5.0

OUTPUT VOLTAGE WAVEFORM. NATURAL SAMPLING

LOAD CURRENT WAVEFORM. NATURAL SAMPLING
FIG 4.6 COMPUTER GRAPH PLOT FOR \( \frac{f_s}{f_n} \) RATIO = 8.5
output component for each graph plot, appropriate \( \frac{f_1}{f_0} \) Ratios were selected to give a fundamental frequency of not less than 0.5 of the output frequency.

It is to be noted that an inductive load acts as a low-pass filter which tends to suppress high frequency components. The sub-harmonic components in the current waveform are therefore a greater proportion, and the super-harmonic components a smaller proportion, of the wanted component than in the voltage waveform. The effect of sub-harmonics can therefore more easily be seen in the current waveform.

Figure 4.3 is for a Ratio value of 4.0, and the nett area enclosed by the current waveform (and the voltage waveform) is the same for successive complete cycles of the wanted output component. Hence there is no a.c. sub-harmonic component. This is as expected since there can be no a.c. sub-harmonic component when the fundamental frequency is equal to the wanted output frequency. The nett area enclosed by the current waveform is not zero, however, and therefore as explained in section 3.5, a d.c. component must exist in the output. This is shown in tables 4.1a and 4.1b to have a magnitude of 15% of the wanted component.
Figure 4.4 shows the waveforms for a \( \frac{pf_1}{f_0} \) Ratio of 4.5 for which the fundamental frequency is 0.5 of the wanted output frequency. In this case, the nett area enclosed by the current waveform during the first complete cycle of the wanted component is less than that during the second cycle. A sub-harmonic component must be present, and table 4.1b shows that this is the 23% sub-harmonic at 0.5 of the wanted component (i.e. at the fundamental frequency in this case).

Figure 4.5 shows the waveforms for a \( \frac{pf_1}{f_0} \) Ratio of 5.0. In this case, it is clear that the nett areas enclosed by the current waveform during successive cycles of the wanted component are zero. There can be no d.c. component or sub-harmonic components as confirmed by the results shown in tables 4.1b and 4.1c.

Figure 4.6 shows examples of waveforms for a higher Ratio value than in figures 4.3 - 4.5. This is for 8.5 which is shown in table 4.1f to give a sub-harmonic component of magnitude 5% of the wanted component at 0.5 of the frequency of the wanted component. In figure 4.6, a variation in the nett areas enclosed by the current waveform during successive cycles of wanted component can be discerned, even for this 5% sub-harmonic component.
It is to be noted, however, that a variation in the nett areas enclosed by the voltage waveform is not so readily discernible.

Figures 4.3 - 4.6 identify the problem in diagrammatic form of the sub-harmonic components in the output of an inhibited cycloconverter with an inductive load taking non-sinusoidal current. The solution to the problem lies in controlling the switching instants in such a way that the nett areas enclosed by the current waveform during successive cycles of the wanted component are reduced.

(4.4) Application of Modified Control Techniques to the Inhibited Cycloconverter when supplying Non-sinusoidal Load Current

(4.4.1) General

It has been shown in this chapter that for any specified pf$_1$/f$_0$ Ratio, the frequency of a sub-harmonic component is only a particular value of a general harmonic. For any other pf$_1$/f$_0$ Ratio, the general harmonic takes one of an infinite number of other values of which only a limited band will be less than the wanted output component.

It is therefore anticipated that attenuation of a sub-harmonic component in the cycloconverter output will result in attenuation of the general harmonic over its whole frequency range, including

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the super-harmonic region. It is inherent in the operation of the cycloconverter that the output voltage must be non-sinusoidal and must contain a large number of harmonic components. It would therefore be impossible to attenuate all harmonics. The objective is therefore to attenuate those general harmonics which give rise to significant sub-harmonic components for the range of $\frac{pf_1}{f_0}$ Ratios over which the cycloconverter is to be normally operated. The discussion in this section is therefore concerned with the computer investigation into the application of modified control techniques, using the results given in the earlier part of this chapter for natural sampling as the basis for reference.

(4.4.2) Regular Sampling

In chapter 3, regular sampling was shown to give very significant attenuation of the sub-harmonic components in the output voltage of the circulating-current cycloconverter, and to give some attenuation in the case of the inhibited cycloconverter supplying sinusoidal load current.

Replacement of natural sampling by regular sampling was therefore investigated for the inhibited cycloconverter supplying non-sinusoidal load current by appropriate modification of the computer program. The program was run for the same conditions as
for natural sampling (section 4.3.2) and the results for $pf_i/f_o$ Ratios of 3.8, 4.8, 5.8, 6.8, 7.8 and 8.8 are shown in table 4.2. Those general harmonics, which are identified in tables 4.1a - 4.1f as being responsible for the significant sub-harmonics, are all represented at specific frequency values in table 4.2. This table therefore contains sufficient data for a preliminary comparison to be made between the relative effects of natural and regular sampling.

A comparison between the predominant sub-harmonic components in tables 4.2 and 4.1a - 4.1f shows that, at a $pf_i/f_o$ Ratio of 3.8, the replacement of natural by regular sampling results in 25% and 15% attenuation of the two sub-harmonic components of frequencies $0.2f_o$ and $0.6f_o$ respectively. There is also a 27% attenuation of the $0.8f_o$ sub-harmonic component for a $pf_i/f_o$ Ratio of 4.8. Further comparisons between the results show that replacement of natural by regular sampling has insignificant effect on the lower magnitude sub-harmonic components.

An example of a computer graph plot of the output voltage and current waveforms for operation with regular sampling is shown in figure 4.7. This is for a $pf_i/f_o$ Ratio of 4.5, for which as shown in table 4.1b there is one sub-harmonic component at $\frac{1}{2}$ of the wanted
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<th>0.3</th>
<th>0.4</th>
<th>0.5</th>
<th>0.6</th>
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<th>0.8</th>
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**TABLE 4.2** 3-PULSE INHIBITED CYCLOCONVERTER: MAGNITUDES OF THE LOW FREQUENCY COMPONENTS FOR REGULAR SAMPLING
OUTPUT VOLTAGE WAVEFORM, REGULAR SAMPLING

LOAD CURRENT WAVEFORM, REGULAR SAMPLING

FIG 4.7 COMPUTER GRAPH PLOT FOR $\frac{f_2}{f_1}$ RATIO = 4.5
output frequency. A comparison with the corresponding waveforms for natural sampling (figure 4.4) shows that regular sampling alters the waveshapes of the output voltage and current, and that by applying the equal-area criterion, the presence of a sub-harmonic component can again be detected in the load current waveform.

(4.4.3) Pulse Position Modulation

The application of pulse position modulation to the control of the inhibited cycloconverter is an extension of regular sampling and is described in section 3.7.4. The results for the inhibited cycloconverter when supplying sinusoidal load current showed significant attenuation of the sub-harmonic components in the output voltage compared with the case for natural sampling. This modified control technique was therefore investigated for the inhibited cycloconverter when supplying non-sinusoidal load current.

The results for $pf_1/f_0$ Ratios of 3.8, 4.8, 5.8, 6.8, 7.8 and 8.8 are shown in table 4.3, corresponding to the results shown in table 4.2 for regular sampling. A comparison between these results and the results in tables 4.1a - 4.1f for natural sampling shows that, contrary to the case of the inhibited cycloconverter supplying sinusoidal load current, pulse position modulation negates the improvements obtained with regular sampling. Furthermore, where regular sampling did not lead to improvements, pulse position modulation creates a deterioration.
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**TABLE 4.3** 3-PULSE INHIBITED CYCLOCONVERTER: MAGNITUDES OF THE LOW FREQUENCY COMPONENTS.
PULSE POSITION MODULATION
Thus, although pulse position modulation results in significant improvements when the inhibited cycloconverter is supplying sinusoidal load current, when the load current is non-sinusoidal, pulse position modulation has quite the opposite effect giving consistent deterioration in respect of all sub-harmonic components.

The reason for these differences can be related to the nature of the inhibition switching functions which are defined by the load current zeros. In the case of sinusoidal load current, the inhibition switching instants occur at regular intervals and are not affected by modifications, such as pulse position modulation, to the method of control. However, when the load current is non-sinusoidal at low values of the $\text{pf}_1/f_0$ Ratio, the conduction periods of the thyristor groups which determine the intervals between the inhibition switching instants, may vary as discussed in section 4.3.1, and may be affected by modifications to the control technique. The equal-area criterion for attenuating the sub-harmonic components is therefore compounded by the variable conduction periods in the case of the inhibited cycloconverter supplying non-sinusoidal current.

In chapter 3, figure 3.16 shows that pulse position modulation has the effect of advancing the thyristor switching instants in the positive group of the cycloconverter from their values for regular sampling. Advancing the thyristor switching instants therefore affects the inhibition switching functions (when the load-current
waveform is not sinusoidal) in such a way as to increase the imbalance of the nett areas enclosed by the waveform for successive cycles of the wanted output component, and thus to amplify the sub-harmonic components. This underlines the relevance of the irregularity of the inhibition switching instants to the magnitude of the sub-harmonic components and indicates the need for a closer study of this aspect of the control of the cycloconverter.

In section 4.3.1, it was seen that for low \( pf_1/f_0 \) Ratios, if a commutation occurs just before the load current reaches zero, the sudden rise in the driving voltage applied to the load through the thyristors of the cycloconverter causes a further pulse of current. This may delay the current zero by an appreciable proportion of a cycle of wanted output component, thereby delaying the inhibition switching instants at which the conduction period of the incoming thyristor group starts. This foreshortens the next conduction period, and the overall effect is to accentuate the imbalance of the nett areas enclosed by the output waveform and therefore to oppose the requirements of the equal area criterion for elimination of the sub-harmonic components. Thus, in order to avoid long delays of the current zero, the magnitude of the driving voltage after a commutation must be reduced if this commutation occurs near the anticipated current zero.

This objective can be achieved by retarding the firing angles - that is, the thyristor switching instants-
at the end of each conduction period. The end of the conduction period in the positive group, for example, occurs at or soon after the voltage zero on the negative slope of the reference wave. Therefore in terms of pulse position modulation, this can be arranged by making the slope of the ramp functions dependent on the slope of the reference wave. At the beginning of the conduction period of the positive group, the slope of the reference wave is positive, and therefore the slope of the ramp function is positive, as shown in figure 3.16 in chapter 3, thus advancing the thyristor switching instants. At the peak of the reference wave, the ramp is arranged to have zero slope. On the negative slope of the reference wave, the ramp functions have negative slope so as to retard the thyristor switching instants.

This modified pulse position modulation method of control was tested by making the appropriate modification to the computer program, and the results for $\frac{pf_1}{f_0}$ Ratios of 3.8, 4.8, 5.8, 6.8, 7.8 and 8.8 are shown in table 4.4. Compared with the results in table 4.3 for the unmodified pulse position modulation, they show substantial attenuation of most of the main sub-harmonic components. They also show significant attenuation compared with the results for regular sampling (table 4.2) and natural sampling (tables 4.1a - 4.1f). The modification to the pulse position modulation method of control is thus shown to have the desired effect on the
| Ratio | \( f_0/f_1 \) % | 0.1  | 0.2  | 0.3  | 0.4  | 0.5  | 0.6  | 0.7  | 0.8  | 0.9  | 1.0  | 1.1  | 1.2  | 1.3  | 1.4  | 1.5  |
|-------|----------------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|
| 3.8   | 79.1           | 10   | 10   | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  |
| 4.8   | 62.4           | 1    | 3    | 3    | 6    | 9    | 100  | 100  | 11    | 11    | 11    | 11    | 11    | 11    | 11    |
| 5.8   | 51.7           | 11   | 5    | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  |
| 6.8   | 44.2           | 2    | 2    | 3    | 8    | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  |
| 7.8   | 38.4           | 4    | 3    | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  |
| 8.8   | 34.0           | 1    | 2    | 2    | 5    | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  | 100  |

**Table 4.4** 3-Pulse Inhibited Cycloconverter: Magnitudes of the Low Frequency Components. Modified Pulse Position Modulation
output waveforms. Control of the inhibition switching instants by appropriate control of the thyristor switching instants was therefore given further investigation as discussed in the next section.

(4.4.4) **Distorted Reference Wave**

It is instructive to review the application of regular sampling in terms of natural sampling of a modified reference wave, and figure 4.8a shows a sinusoidal reference voltage which is regularly sampled at the starts of the cosine timing waves. Crosses are marked where the held samples intersect with the associated cosine timing waves. These crosses give the thyristor switching instants for cosinusoidal control with regular sampling of a sinusoidal reference voltage, as discussed earlier in chapter 3. The crosses would also mark intersections of the cosine timing waves and the natural samples of a non-sinusoidal reference wave. Figures 4.8b and 4.8c show two possible alternatives of such a non-sinusoidal reference wave. Figure 4.8c is characteristic of a sinusoid distorted by its second harmonic and as such, consists of the first two components of the mathematical series which defines the sawtooth waveform shown in figure 4.8b. It is therefore shown that regular sampling of a sinusoidal reference wave is equivalent to natural sampling of a reference wave containing second harmonic distortion.
FIG 4.8 EQUIVALENCE OF REGULAR SAMPLING OF A SINUSOIDAL REFERENCE WAVE AND NATURAL SAMPLING OF A NON-SINUSOIDAL REFERENCE WAVE
Referring now to figure 4.9, it is seen that towards the anticipated current zero at the end of the conduction period, second harmonic distortion of the reference wave retards the thyristor switching instants. This is the requirement discussed in section 4.4.3 for reducing delays of the current zeros in order that the differences in areas enclosed by successive cycles of the output waveform at wanted output frequency should be reduced, thereby attenuating the sub-harmonic components. The thyristor switching instants at the beginning of the conduction period are advanced thus compensating for any attenuation of the wanted output component.

A computer study was carried out to investigate control of the inhibited cycloconverter using the distorted reference wave described above. Greater attenuation of the sub-harmonic components of the output voltage waveform was obtained by employing regular sampling, rather than natural sampling, of the distorted reference wave. The results for pf1/f0 Ratios of 3.8, 4.8, 5.8, 6.8, 7.8 and 8.8 are shown in table 4.5. These results represent significant overall improvements on the results given in table 4.4 for the modified pulse position modulation method, and even more so when compared with the results for natural sampling given in tables 4.1a - 4.1f. The main exceptions are the 1.2f0 component at a pf1/f0 Ratio of 4.8, and the 0.2f0 component at a pf1/f0 Ratio of 5.8. These components are both specific values of the (3f1 - 6f0) general harmonic.
FIG 4.9 EFFECT OF DISTORTED REFERENCE WAVE ON THE SWITCHING INSTANTS
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<th>$\frac{\varnothing}{(D.C.)}$</th>
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<th>0.3</th>
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</table>

**TABLE 4.5** 3-PULSE INHIBITED CYCLOCONVERTER: MAGNITUDES OF THE LOW FREQUENCY COMPONENTS. DISTORTED REFERENCE WAVE
As an additional means of reducing delays of the current zeros at the end of conduction periods, inhibition can be applied to the conducting group of thyristors at, say, the reference voltage zero. No further commutations will occur after the voltage zero, and firing pulses to the other group of thyristors are released when the current reaches zero. This modification, which will be referred to as 'pre-inhibition', was incorporated in the computer program, and the associated results for the inhibited cycloconverter controlled by a reference wave with second harmonic distortion are shown in table 4.6. It is seen that the pre-inhibition now gives significant attenuation of the $1.2f_o$ component at a $pf_i/f_o$ Ratio of 4.8 and the $0.2f_o$ component at a $pf_i/f_o$ Ratio of 5.8.

(4.4.5) **Comparison of Results**

It is shown in the foregoing discussion that very significant improvements to the operation of the inhibited cycloconverter, when supplying non-sinusoidal current, are obtained by employing regular sampling of a distorted reference wave.

In order to make a more detailed comparison between this control technique and natural and regular sampling, further data was obtained from the computer studies. In section 4.3, it was shown that the magnitudes of the main general harmonics can be extracted from the tabulated computer results for natural sampling.
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**TABLE 4.6** 3-PULSE INHIBITED CYCLOCONVERTER: MAGNITUDES OF THE LOW FREQUENCY COMPONENTS. DISTORTED REFERENCE WAVE WITH PRE-INHIBITION
shown in tables 4.1a - 4.1f. These magnitudes, together with the corresponding results for the modified control techniques, are plotted against the $pf_i/f_o$ Ratio in figures 4.10 - 4.16. Each figure shows the magnitude of one general harmonic for the four methods of control as follows:

1 - Natural sampling of a sinusoidal reference wave;
2 - Regular sampling of a sinusoidal reference wave;
3 - Regular sampling of a distorted reference wave (as described earlier in this chapter);
4 - As 3 but with pre-inhibition applied.

The results shown in figures 4.10 - 4.16 are for a 3-pulse inhibited cycloconverter operating at 0.9 modulation factor and supplying a load having time constant 0.005 s. They cover the sub-harmonic frequency range together with a limited extension into the super-harmonic frequencies, corresponding to the frequency range in tables 4.1a - 4.1f.

The pronounced peaks and troughs in the general trend of the magnitude of each harmonic are due to its coincidence with one or more other harmonics at particular frequencies.

The replacement of natural by regular sampling of a sinusoidal reference wave is seen to attenuate those sub-harmonic components which are due to the $(3f_i - 4f_o)$ harmonic, from a mean level of 20% to 14% of the wanted
FIG 4.10 MAGNITUDE OF THE \((3f_1 - 4f_o)\) HARMONIC
FIG 4.11 MAGNITUDE OF THE \((3f_i - 6f_o)\) HARMONIC
FIG 4.12 MAGNITUDE OF THE \((3f_1-8f_0)\) HARMONIC
FIG 4.13 MAGNITUDE OF THE (6f₁ - 7f₀) HARMONIC
FIG 4.14 MAGNITUDE OF THE $(6f_i - 9f_o)$ HARMONIC
FIG 4.15 MAGNITUDE OF THE \((6f_c - 11f_o)\) HARMONIC
FIG 4.16 MAGNITUDE OF THE \((6f_1 - 13f_0)\) HARMONIC
component. By introducing second harmonic distortion in the reference wave, further attenuation can be obtained to a mean level of 6%. Regular sampling of a sinusoidal reference wave does not result in attenuation of the other harmonics shown in figures 4.10 – 4.16; for attenuation of these harmonics, a distorted reference wave is shown to be required. For example, the \((6f_1 - 9f_0)\) harmonic is attenuated from a mean level of 6% of the wanted component to 3%; the \((6f_1 - 11f_0)\) is attenuated from 4% to 2% (with pre-inhibition applied) and the \((6f_1 - 13f_0)\) harmonic is attenuated from 3½% to 2%. Therefore, for these three harmonics, a 50% reduction in the magnitude of the sub-harmonic component is achieved by employing regular sampling of a distorted reference wave in place of natural sampling of a sinusoidal reference wave.

By employing a distorted reference wave, significant attenuation of the \((3f_1 - 6f_0)\), \((3f_1 - 8f_0)\) and the \((6f_1 - 7f_0)\) harmonics is obtainable over only parts of their sub-harmonic ranges.

Figure 4.17 shows the variation in output voltage (at wanted output frequency) with the \(pf_1/f_0\) Ratio. The horizontal axis is also shown in terms of the ratio of output to input frequency for a 3-pulse cycloconverter. The figure shows the general trend of the variation in voltage and also the voltages at specific Ratio values which do not fit the general trend. It is seen that the general variation in output voltage for natural sampling
FIG 4-17 VARIATION IN OUTPUT VOLTAGE WITH OUTPUT FREQUENCY
of a sinusoidal reference wave is within a band of only 2% of the input voltage and is within 2% of the theoretical output voltage (74.3% of the input voltage). The other control methods under consideration all show a drop in output voltage when the $\text{pf}_1/f_0$ Ratio is decreased. The drop in output voltage is about 6% - 7% of the input voltage.

At specific values of the $\text{pf}_1/f_0$ Ratio, the output voltage shows discontinuities from the general trend. Figure 4.17 shows that the discontinuity may be as high as 8% of the input voltage. Figures 4.10 - 4.16 give an indication of the reason for this: it is seen in these figures that all of the significant harmonics coincide with the wanted component for integer values of the $\text{pf}_1/f_0$ Ratio. As discussed in section 2.6.4, this is also the condition when the fundamental repetition frequency is equal to the wanted output frequency and when no sub-harmonic components can exist in the output (except for the possibility of a d.c. component). In each of these cases the following equality holds:

$$ mpf_1 - nf_0 = f_0 $$

where $m$ and $n$ are integer numbers.

or

$$ \frac{\text{pf}_1}{f_0} = \frac{l + q}{m} $$

For the $3f_1$ family of harmonics, $m = 1$; for the $6f_1$ family $m = 2$. Therefore for each of the harmonics shown in figures 4.10 - 4.16, the right hand side of this expression is always an integer, thus confirming the observation that, in these cases, the harmonic coincides with the wanted component when the $\text{pf}_1/f_0$ Ratio is an integer value.
In figure 4.17 it is also seen that discontinuities occur when the $pf_1/f_0$ Ratio is midway between integer values. This is due to coincidence with the wanted component of harmonics other than those shown in figures 4.10 - 4.16. An example is the $(12f_1 - 17f_o)$ harmonic for a 3-pulse cycloconverter for which

$$\frac{1 + q}{m} = 4.5$$

Such harmonics as the $(12f_1 - 17f_o)$ harmonic can be extracted from the tabulated values given in tables 4.1a - 4.1f, and the magnitudes of some of them are sufficiently high to have a noticeable effect on the magnitude of the output voltage when they coincide at the same frequency with the wanted output component.

Harmonics with a higher value of 'm' coincide with the wanted output component at intermediate values of the $pf_1/f_0$ Ratio, but their magnitudes are generally low enough to have insignificant effect on the output voltage.
An experimental cycloconverter was constructed to test the validity of the computer program. The design and construction of the cycloconverter will be described in three parts:

a) Power circuit
b) Firing control circuits
c) Inhibition control circuit

Although experimental results for 3-pulse operation of the cycloconverter were required, the cycloconverter was designed for alternative 3- or 6-pulse operation to allow for future flexibility. Details of the main components are given in Appendix 5.

(5.2) Power Circuit

The thyristors were mounted on heat sinks and protected by water-immersed fuses. Figure 5.1 shows the thyristor connections for 6-pulse operation, and the thyristors are labelled in the figure for identification in subsequent diagrams.

The 10kΩ resistors shown in figure 5.1 between the thyristor common connections and the supply neutral provide a leakage path for starting and for no-load operation.

The power circuit is supplied from the 415 V
FIG. 5.1 EXTERNAL CONNECTIONS OF ONE OUTPUT PHASE OF THE EXPERIMENTAL CYCLOCONVERTER (SHOWN CONNECTED AS A 6-PULSE CYCLOCONVERTER)
3-phase mains supply via a variable ratio autotransformer.

Operation of the cycloconverter in the inhibited mode permits the circuit shown in figure 5.1 to be redrawn as 6 inverse-parallel pairs of thyristors as shown in figure 5.2. This shows that the snubber circuits, which attenuate the high frequency transients resulting from thyristor and inhibition switching operations, are required across pairs of thyristors rather than across individual thyristors.

(5.3) Firing Control Circuits

(5.3.1) General

The basic principles used in ordinary rectifier practice were applied to the control of the firing angles of the thyristor in the experimental cycloconverter. Instead of fixed firing angles, however, variable or modulated angles are required for the cycloconverter. As discussed in section 2.2, the usual method of deriving the modulated firing angles is by comparing the instantaneous magnitudes of the reference voltage and the cosinusoidal timing voltages by the process identified in section 3.3 as natural sampling. The firing control circuits were designed, in the first instance, for control of the cycloconverter in the circulating-current mode using natural sampling. The additional control circuits for operation in the inhibited mode were then added.
FIG. 5·2 R-C SNUBBER CIRCUITS FOR THE INHIBITED CYCLOCONVERTER

\[ R = 330 \Omega \]
\[ C = 0.1 \mu F \]
For the construction of the firing control circuits, the use of discrete components was kept to a minimum and encapsulated modules or integrated circuits were used wherever possible.

(5.3.2) Cosine Timing Waves

The cosine timing waves are obtained from the 415 V 3-phase mains supply via three 240/120-0-12 V transformers (see figure 5.3). The centre-taps of the secondary windings are commoned to give a 6-phase output from the transformers, and the line-neutral voltages are reduced to about 3 V with resistive potential-dividers connected to the secondary terminals of the transformers. The cosine timing waves are then led to the pulse-forming circuits.

Figure 5.4 shows those portions of the 6-phase output which constitute the cosine timing waves. The positive-group thyristors require negative-going portions (figures 5.4b and 5.4c) and the negative-group thyristors require positive-going portions (figures 5.4d and 5.4e). One complete wave (i.e., positive-going plus negative-going portions) is utilised for forming the gate pulses for two thyristors, one in the positive group and one in the negative group. This is shown in figure 5.3 where, for example, the pulse-forming circuits for thyristors P2 and N5 are controlled by the R-phase voltage.
FIG. 5-3 INPUT STAGE OF FIRING CONTROL CIRCUIT FOR 6-PULSE CYCLOCONVERTER WITH SINGLE PHASE OUTPUT
FIG 5.4 DERIVATION OF THE COSINE TIMING WAVES FROM THE SUPPLY VOLTAGES
(5.3.3) Reference Wave and End-Stop Control

The reference wave, the voltage of which is compared with the cosine timing voltages in the pulse-forming circuits, is obtained from an external signal generator. The frequency setting of the signal generator gives the required output frequency of the cycloconverter whilst its amplitude setting controls the magnitude of the wanted component of the output voltage.

The ratio of the peak values of the reference and cosine timing waves is defined as the modulation factor. If the modulation factor exceeds unity, the firing angle may exceed $180^\circ$ resulting in commutation failure and probably a short circuit. To prevent the firing angle reaching $180^\circ$, an end-stop control was connected between the signal generator and the input to the pulse-forming circuits. The circuit diagram is shown in figure 5.5. The two 100pF electrolytic capacitors are charged to a voltage equal to the difference between the peak value of the cosine timing wave and the forward voltage drop of the two IN4148 silicon diodes in series. One of the capacitors is charged by the positive cosine timing wave, the other by the negative cosine timing wave. The voltages at the bases of the two transistors are equal to the voltages across the electrolytic capacitors. The instantaneous voltages at the emitters of the transistors are determined by the output of the 3-phase bridge rectifier.
FIG. 5.5 END-STOP CONTROL OF FIRING ANGLE
(6 AAZ15 germanium diodes) which in turn is determined by the magnitude of the 3-phase reference voltage.

Thus, if the instantaneous reference voltage exceeds the voltage to which the electrolytic capacitor is charged, then the associated transistor will conduct. The p-n-p transistor (BCY70) controls the positive half cycle of the reference voltage, and the n-p-n transistor (BC107) controls the negative half cycle. When either transistor conducts, the reference voltage is held at a value equal to the total voltage drop across one AAZ15 diode and the associated transistor. The reference voltage outputs from the end-stop control are taken via operational amplifier buffers in order to give a low impedance output.

(5.3.4) **Pulse-Forming Circuits**

The pulse-forming circuit, shown as a block in figure 5.3, is shown in detail in figure 5.6. Each pulse-forming circuit initiates the gate-pulses for one thyristor in the positive group and one thyristor in the negative group, as discussed in section 5.3.2. The waveforms are shown in figure 5.7.

The first stage of the pulse-forming circuit is a differential amplifier, the basic component of which is a 741 operational amplifier. The input signals are the reference wave and the cosine timing wave. Whilst the instantaneous voltage of the reference wave is less than
REFERENCE VOLTAGE

COSINE TIMING VOLTAGE

MONOSTABLE CIRCUIT

BISTABLE CIRCUIT

INVERTER

INTERCONNECTION WITH OTHER CONTROL CHANNELS.

FIG. 5.6  PULSE-FORMING CIRCUIT FOR ONE THYRISTOR IN EACH GROUP
FIG. 5.7 WAVEFORMS FOR FIG 5.6
that of the cosine timing wave, the output from the operational amplifier is at +12 V; when the instantaneous voltage of the reference wave becomes greater than that of the cosine timing wave, the output switches to -12 V.

The output from the operational amplifier is split into two control channels, one for a positive group thyristor and the other for a negative group thyristor. The separate control channels are identical except that the output from the operational amplifier for the positive group thyristor is taken via an inverter.

The next stage of the pulse-forming circuit consists of a monostable circuit in each control channel. This circuit, details of which are given in Appendix 6, is constructed from NOR-gates and an R-C circuit, the values of resistance and capacitance being selected to give an output pulse of approximate width 0.4 ms for every positive-going input transition. Thus, short-duration pulses are produced at the output of the monostable circuit in the positive group control channel at the instant when the reference voltage becomes greater than that of the cosine timing wave, as shown in figure 5.7, and at the output of the monostable circuit in the negative group control channel at the instant when the reference voltage becomes less than that of the cosine timing wave.

Each monostable circuit is connected to a bistable circuit (Appendix 6) consisting of two NOR-gates. The monostable and bistable circuits are interconnected, as
shown in figure 5.8, to form 3-stage ring counters. By this means, when a pulse is produced from one of the monostable circuits, the output from the associated bistable circuit switches to logic 1 level. The output stays at that level until a pulse is received from one of the other monostable circuits. In this way, rectangular pulses are produced at the pulse-forming circuit outputs having a duration equal to the required conduction period of the associated thyristor in the power circuit.

The output stage shown in figure 5.9 provides the required gate-pulse power for driving the thyristor, and also isolation between the power and control circuits. The two NOR-gates are for the inhibition control which is discussed in detail later in this chapter. If the input from the inhibition control circuit is at logic 0 level, then the two NOR-gates have no effect on the thyristor gate-pulses.

The rectangular pulse produced by each pulse-forming circuit is fed to a UPA61 thyristor drive oscillator. This is a standard encapsulated free-running oscillator, with an externally connected 0.01 μF capacitor, giving a 10kHz output when its input is at logic 1 level. The rectangular pulses are therefore converted into 10kHz oscillatory pulses having the same duration as the rectangular pulses. The oscillatory pulses are then passed to the thyristor gates via TT61 thyristor output transformers. These are also standard encapsulated components.
FIG 58 DIAGRAM SHOWING INTERCONNECTIONS BETWEEN THE MONOSTABLE AND BISTABLE CIRCUITS OF THE FIRING CONTROL CIRCUITS FOR EITHER THE POSITIVE GROUP OR NEGATIVE GROUP.
FIG 5.9 OUTPUT STAGE OF FIRING CONTROL CIRCUIT FOR ONE THYRISTOR
5.3.5) Regular Sampling Circuits

The requirement for the regular sampling circuits is to sample the reference wave at regular intervals and to hold the sampled value for a specific duration. As discussed in section 3.4.1, it is convenient to sample the reference wave at the starts of the cosine timing waves, and to hold the samples for the duration of the timing waves; that is, one half cycle of the input frequency. Since there is an overlap between successive periods during which the samples are held constant, separate regular sampling control channels are required for each input phase.

Each sampling instant coincides with the intersections of the two commutation phase voltages. Sampling instants for the positive group of thyristors occur when the two phase voltages are positive, whilst those for the negative group occur when they are negative. In each case, the samples are to be held for 180° periods. The sampling instants for both groups are detected by full-wave rectification of the output voltages from the mains transformers shown in figure 5.3, and feeding them to an operational amplifier connected as a comparator. The circuit is shown in figure 5.10 and the input and output waveforms of the Y-phase comparator are shown in figures 5.11a and 5.11b. The transitions of the square-wave output occur at all intersections of the rectified phase voltages, but only the positive-going transitions...
FIG. 5.10 CIRCUIT FOR SAMPLING THE REFERENCE VOLTAGE (FOR ONE OUTPUT PHASE OF THE CYCLOCONVERTER)
FIG. 5-11 WAVEFORMS FOR ONE CHANNEL OF CIRCUIT SHOWN IN FIG. 5-10 WHEN SWITCHED FOR REGULAR SAMPLING
correspond to the sampling instants. The negative-going transitions are not required and have no effect on the sample-and-hold part of the circuit.

The output from the comparator is differentiated by an R-C circuit to give a train of positive and negative pulses which are then offset to -12 V when the switch shown in figure 5.10 is at -12 V (for regular sampling). The resultant train of pulses (figure 5.11c) is used to control the field-effect transistor FET1 in the sample-and-hold part of the circuit in figure 5.10.

The field effect transistor FET1 acts as a sampling switch for the reference wave which is fed to it from a signal generator via an operational amplifier buffer. This transistor is held in the non-conducting state if its gate voltage is less than about -3 V. It is therefore in its conducting state only during the short durations of the positive pulses shown in figure 5.11c. By this means, the instantaneous value of the reference wave is sampled at the required instant, and held for the 180° period.

If the switch in figure 5.10 is switched to 0 V, then the gate voltage of FET1 is at 0 V causing it to conduct for practically the whole of the 180° period. The reference wave therefore passes through with no sample-and-hold action. In this way, the cycloconverter can be operated with natural or regular sampling simply by operating the switch shown in figure 5.10.
The circuit described above was connected between the end-stop control and the pulse-forming circuit of each control channel of the experimental cycloconverter. For operation with regular sampling, the thyristor firing pulses occur at intersections of the cosine timing waves and the held samples of the reference wave, whilst by switching to natural sampling, they occur at intersections of the cosine timing waves and the reference wave.

(5.4) Inhibition Control Circuits

(5.4.1) General

There is no established method of controlling changeover of conduction of load current from one group of thyristors to the other in practical cycloconverters, and descriptions in the literature give few details of the individual methods adopted for their control. An inhibition control system was therefore developed for the experimental cycloconverter, and the circuit shown in figure 5.12 was constructed as the first step.

The purpose of this circuit is to transfer the load current from one thyristor group to the other at each reference voltage zero. The part of the circuit containing the 741 operational amplifier acts as a zero-crossing detector giving a square wave output as in figure 5.13. The IN4148 diode allows only the positive half cycles to pass to the pair of cascade-connected ST241
FIG. 5.12 INHIBITION CONTROL USING REFERENCE VOLTAGE
FIG 5.13 WAVEFORMS FOR FIG 5.12
transistors. The waveforms of the signals at the respective collectors of these transistors are rectangular and in antiphase with each other as shown in figure 5.13. The collector terminal $T_p$ was connected to the inhibition terminals of all the Firing Control Circuits (see figure 5.9) of the positive thyristor group, and $T_n$ was connected to all the negative group Firing Control Circuits.

This method of inhibition control does not allow for the possibility that the load current may not be zero at the instant when the reference voltage passes through zero. Firing pulses may then be released to the incoming thyristor group whilst the outgoing group is still conducting. In order to prevent a short circuit then occurring across the input phases, the circulating-current reactors were connected as in figure 5.1 and the load current was restricted to a few milliamperes. This enabled the cycloconverter to be operated in the inhibited mode and provided valuable experience in developing an inhibition control scheme.

In order to avoid the possibility of a short circuit due to the release of firing pulses to the incoming thyristor group before the outgoing group has ceased conducting, some form of current sensing is essential. There will generally be several current zeros at the end of each half-cycle of load current, and it is necessary to select one at which inhibition control will be initiated. Therefore a reliable method of inhibition
control must combine information about both the reference voltage and the load current. The method developed for the experimental cycloconverter is described in the following sections.

(5.4.2) Detection of Load Current Zeros

Load current zeros may be detected by monitoring the load current directly with, for example, d.c. current transformers (Fallside et al.\textsuperscript{8}, Arikan\textsuperscript{20}). However, d.c. current transformers require auxiliary supplies of frequency of about 2 kHz which can cause a delay of up to about ¼ ms in detecting a current zero. A further problem is that this method does not reliably distinguish between low current and zero current.

Another method of detecting load current zeros, requiring two pairs of back-to-back connected diodes, was used by Takahashi and Miyairi\textsuperscript{(11)}. Whilst there is a voltage drop (of about 1 V) across a diode, then current must be flowing; when there is no voltage drop, then the current must be zero. The voltage across each pair of diodes is monitored and used to control the inhibition process of the cycloconverter.

The diodes in this case have to be fully rated for the maximum load current. A cheaper method which does not incur the heat-loss penalty of the diodes is to monitor the voltages across each of the thyristors. Whilst a thyristor is conducting, the voltage across it is practically constant at about 1 V for any magnitude of
current down to the holding current. As soon as the thyristor ceases conduction, the voltage across it becomes dependent on the supply voltage. Thus, when the absolute voltages across all the thyristors are simultaneously greater than about 1 V, then the load current must be zero.

The voltages across the thyristors can be monitored directly using combinational logic circuits as described by Hamblin and Barton (19). These circuits operate at a 'floating voltage' with respect to earth which would be expected to present problems regarding the power supplies of the integrated circuits. Isolating transformers connected across the thyristors would not necessarily overcome this problem without introducing other problems concerning the distortion of the thyristor voltage waveform.

However, optocouplers and differential amplifiers present solid-state alternatives to isolating transformers, and as there is no published evidence of these devices having been used, successfully or otherwise, for cycloconverter applications, a series of laboratory tests were carried out.

The optocoupler consists of a light-emitting diode and a phototransistor encapsulated in a common package. The diode is the input to the device and is connected in series with a resistor to limit the diode current to its rated value when subjected to the maximum anticipated voltage. The phototransistor provides a low-
level output operating in the switching mode between saturation and the off-state. The phototransistor was connected to a level detector, the basic component of which was a 301 integrated circuit operational amplifier, and both components were selected for fast response times giving a total response time of about 10-20 μs. This is fast enough to give close control over the inhibition of the cycloconverter.

The other alternative, the differential amplifier, has a response time dependent on only an operational amplifier. Its response time is therefore slightly faster than the optocoupler/level detector combination. However, it does not provide electrical isolation between the power and control circuits, and some difficulty was experienced in setting the external resistances to avoid distortion of the output waveform due to common-mode rejection. Common-mode rejection could occur when the thyristor is in the conducting state resulting in similar voltages (with respect to earth) at the two input terminals of the differential amplifier.

Optocouplers and level detectors were finally selected for detection of current zeros because (i) no additional components are required in the power circuit, (ii) electrical isolation is provided between the power and control circuits, (iii) they are compatible with the complementary-M.O.S. integrated circuits which were used in the inhibition control circuits, (iv) the response
time was fast enough for close control of inhibition.

In order to detect current zeros in the inhibited cycloconverter it is not in fact necessary to monitor the voltages across individual thyristors. As shown in figure 5.2, any thyristor in the positive group is effectively connected in inverse-parallel with a thyristor in the negative group. Thus, an optocoupler connected across a thyristor in one group will also be connected across a thyristor in the other group. Furthermore, in the 6-pulse cycloconverter, as with any 6-pulse bridge rectifier, load current flowing through a thyristor in one half of the bridge must inevitably be flowing through a thyristor in the other half. It is therefore only necessary to connect the optocouplers across pairs of thyristors in one half of the bridge. Thus only 3 optocouplers are required for each output phase of a 6-pulse cycloconverter. The same number is also required for 3-pulse operation.

Thus the requirements of optocouplers for zero current detection in the inhibited cycloconverter reduce to one per input phase for each output phase. The connections to the power circuit are shown in figure 5.14. The problem of detecting load current zeros is now that of monitoring the voltages between the input phases and one of the load terminals rather than across individual thyristors. If any one of these voltages is held to the forward voltage drop of a thyristor, then load current
3-PHASE SUPPLY

TO ZERO CURRENT DETECTION CIRCUIT
(SEE FIG. 5.15)

FIG. 5.14 POWER CIRCUIT CONNECTIONS FOR ZERO CURRENT DETECTION
must be flowing; if all three voltages are greater than this value, then the load current must be zero.

The zero current detection circuit shown in figure 5.15 was designed to monitor the three voltages as described above and to deliver a digital signal at logic 1 level at its output terminal whilst load current flows. When there is no load current, the output is required to be at logic 0 level. The four IN 4148 diodes full-wave rectify the voltage across each pair of back-back thyristors, and the maximum instantaneous current passing through the input diode of the TIL 111 optocoupler was limited to 60 mA for the full 415 V supply voltage. The input diode current is zero whilst the associated thyristor is conducting load current. Under this condition, the photo-transistor is in the off-state and its emitter is at about 6 V. The inverting terminal of the 301 operational amplifier in the level detector circuit is therefore also at about 6 V. The input preset potentiometer is set so that the non-inverting terminal of the operational amplifier is at a voltage sufficiently greater than that at the inverting terminal to give a positive output from the operational amplifier.

Under the alternative condition of zero thyristor current, current flows through the input diode of the optocoupler and the phototransistor is in the saturated state. Its emitter, and the inverting terminal of the operational amplifier, are raised to almost 12 V.
FIG. 5.15  ZERO CURRENT DETECTION CIRCUIT FOR INHIBITION CONTROL
The inverting terminal is now at a higher voltage than the non-inverting terminal so that the output of the operational amplifier is zero.

The output of the NOR-gate in figure 5.15 can be at logic 1 level only when all of its inputs are simultaneously at logic 0 level; this is the condition for zero load current in the cycloconverter. To prevent operation of the inhibition control on spurious or insignificant current zeros, the output is taken through a time-delay circuit consisting of an R-C circuit and a NAND-gate. The 1 MΩ pre-set potentiometer was adjusted so that the short duration current zeros did not appear in the waveform at the output of the NAND-gate.

The output from the zero current detection circuit is taken to the changeover logic circuit which will be described in section 5.4.4.

(5.4.3) Detection of Reference Voltage Zeros

The level detector circuit forming part of the zero current detection circuit (figure 5.15) was also used in the zero reference voltage detection circuit as shown in figure 5.16. The mid-point of the ±12 V power supply was commoned with the zero volts terminal of the signal generator (reference voltage), and the 1 kΩ potentiometer was set to give a rectangular waveform of equal positive and negative periods at the output of the operational amplifier. This is the waveform for position(A) as shown in figure 5.17.
FIG. 5.16 ZERO REFERENCE VOLTAGE DETECTION CIRCUIT FOR INHIBITION CONTROL
FIG 5.17 WAVEFORMS RELATING TO FIG 5.16
The output from the operational amplifier is taken to two monostable circuits. As shown in greater detail in Appendix 6, these circuits produce short duration pulses for every negative-going transition at their respective inputs. An inverter NAND-gate is connected between the operational amplifier and one of the monostable circuits which therefore effectively responds to positive-going transitions developed by the operational amplifier. One monostable circuit thus produces pulses for negative-going voltage zeros of the reference voltage, and the other produces pulses for positive-going voltage zeros, as shown in figure 5.17.

The monostable circuits consist of NAND-gates and R-C circuits, and the 470 kΩ pre-set potentiometers were set to give output pulses of about 0.5 ms duration. These were then fed to the changeover logic circuit which is described in the next section.

(5.4.4) Changeover Logic

The changeover logic circuit combines the outputs from the zero current and the zero reference voltage circuits to control the instant of changeover of load current from one thyristor group to the other. The circuit is shown in figure 5.18. The design philosophy adopted is that for prospective changeover from, for example, positive to negative group, all current zeros \( A_o \) in the figure occurring before the instant of the negative-going reference voltage zero \( V_{opn} \) are to be ignored. The first
FIG 5.18 CHANGEOVER LOGIC CIRCUIT FOR INHIBITION CONTROL
current zero occurring after the reference voltage zero initiates the changeover procedure. This involves immediate inhibition (neglecting propagation delays) of the positive group and then, after a short delay, the release of firing pulses to the negative group. Changeover of load current back to the positive group is prevented until the next positive-going reference voltage zero. The detailed description of the operation of the changeover logic circuit follows, and the waveforms around the circuit are shown in figure 5.19.

The voltages at the terminals $V_{onp}$ and $V_{opn}$ are normally at logic 1 level, and reference voltage zeros are indicated by short zero-level pulses. The outputs of the bistable circuits RS1 and RS2 change state from logic 0 to logic 1 level when the respective $V_{onp}$ or $V_{opn}$ pulses are produced. The output remains at logic 1 level until the bistable circuit is reset by a logic 0 level pulse at the other input terminal. This occurs at the instant of changeover (see below). The truth table for an RS bistable circuit is given in appendix 6.

The voltage at the terminal $A_0$ is normally at logic 1 level and current zeros are indicated by logic 0 level pulses. These are inverted through a NAND-gate inverter. The output of NAND1 (or NAND2) remains at logic 1 level whilst either, or both, of its inputs is at logic 0 level, and switches to '0' when both inputs are at '1'. This occurs at the first current zero indication after the reference voltage zero.
REFERENCES
VOLTAGE

V_{ONP}
V_{OPN}

A_0 (TYPICAL)
A_0 (INVERTED)

R\text{S1 OUTPUT}
R\text{S2 OUTPUT}

NAND1 OUTPUT
NAND2 OUTPUT

R\text{S3 OUTPUTS}
\{ Q, \bar{Q} \}

BASE OF T1
BASE OF T2

\text{T}I\text{M}E \text{DE}L\text{AY}

\text{FIG} 5.19 \quad \text{WAVEFORMS FOR FIG. 5.18}
The outputs of NAND1 and NAND2 are the inputs to the bistable circuit RS3. During steady-state conditions these inputs are both at logic 1 level, and the outputs provide two antivalent signals simultaneously. When one of the inputs switches to logic 0 level (at the first current zero after the reference voltage zero), the outputs switch simultaneously to their respective complementary states. The actual signal levels at the outputs of RS3 for all possible combinations of inputs is shown in the truth table in Appendix 6. RS3 is the core of the changeover logic control of the cycloconverter as it provides the necessary output signals for inhibition and conduction of the two thyristor groups.

Positive-going transitions at the outputs of RS3 are transmitted without time delay, and after inversion in the NOR-gates, cut off the associated BL48 transistors. The output of the transistor is then at 24 V. This is fed to the firing control circuits (see figure 5.9) of all the thyristors in the outgoing group and switches the output of the second NOR-gate to logic 0 level. This immediately cuts off the thyristor drive oscillators so as to inhibit the gate pulses. At the same instant, the zero-going transition at the base of the BL48 transistor is also fed back to RS1 (or RS2) which is thereby reset to give an output at logic 0 level. This ensures that further current zeros cannot cause another changeover until after the next reference voltage zero.
Zero-going transitions at the outputs of RS3 are delayed by the R-C circuit and this results in a time delay before the other EL48 transistor is turned on, and hence before firing pulses are released to the other thyristor group. The required time delay is the minimum to avoid a short circuit occurring and was set by adjustment of the 1 MΩ pre-set potentiometer to about 1 ms.

The changeover logic circuit shown in figure 5.18 includes additional NOR-gates associated with the terminals A_{sc1} and A_{sc2}. These terminals are for connection to a short-circuit detection circuit. For operation of the inhibited cycloconverter without short-circuit detection, the terminals A_{sc1} and A_{sc2} are commoned to 0 V of the control circuit power supply. When the cycloconverter was put into operation, it was found that the delay circuits in the changeover logic circuit provided adequate protection against short circuits. It was also found that short-circuits occurring whilst the delay circuits were being adjusted were transient and cleared before any fuse had time to blow.

(5.5) Experimental Results

(5.5.1) General

In order to make direct comparisons between the computer and experimental results, the experimental cycloconverter was connected for 3-pulse operation in the inhibited mode and connected to a load consisting of 25Ω.
resistance in series with 0.125 H inductance, the time constant being 0.005 s. Spectral analyses and an examination of the waveforms were carried out for natural and regular sampling of a sinusoidal reference wave.

(5.5.2) Spectral Analysis

The magnitude of the components of the output voltage of the experimental cycloconverter at selected frequency ratios were measured with a Muirhead Kl34A wave analyser. The specified bandwidth of this wave analyser was claimed to be 2% of the in-tune frequency (which has a minimum scale reading of 3 Hz), and was therefore more suitable for low frequency measurements than the available spectrum analyser which had a bandwidth of 10 Hz. However, in spite of manufacturer's claims, clear readings were not possible with the wave analyser when a large number of components were present.

It was therefore necessary to restrict measurements to particular frequency ratios at which only one sub-harmonic component was expected. The tabulated computer results (tables 4.1a - 4.1f) show that this condition can be obtained by operating the cycloconverter at output frequencies which give $\frac{f_1}{f_0}$ Ratio values of 3.5, 4.5, 5.5, etc. The sub-harmonic frequency is then 0.5 of the output frequency and, as shown in chapter 2, this is also the fundamental frequency. The second harmonic is therefore the wanted component, whilst the third harmonic gives the lowest order super-harmonic component at 1.5
of the output frequency. These frequency relationships were confirmed experimentally using the wave analyser.

The theoretical and experimental results for natural sampling are compared in figure 5.20 and for regular sampling in figure 5.21, and it is seen that the computer results are closely verified experimentally.

A comparison of figures 5.20 and 5.21 shows that the \(0.5 f_0\) sub-harmonic component is attenuated by employing regular in place of natural sampling when the cycloconverter is operated at a \(pf_1/f_0\) Ratio of 4.5. However, at a \(pf_1/f_0\) Ratio of 5.5, the magnitude of the \(0.5 f_0\) sub-harmonic component is increased when regular sampling is employed, whilst at Ratios of 6.5, 7.5 and 8.5, the differences are insignificant.

(5.5.3) **Output Voltage and Current Waveforms**

(5.5.3.1) **General**

Typical waveforms of the output voltage and current of the experimental cycloconverter operated in the inhibited mode with natural sampling are discussed in the following sections. In each figure, the top trace is the output voltage and the bottom trace is the load current. The input frequency is 50 Hz, the modulation factor is 0.9 and the time constant of the load is 0.005 s. The effect of switching to regular sampling had negligible effect on the waveforms, and therefore the following discussion refers equally to natural and regular sampling.
FIG 5.20 COMPARISON OF EXPERIMENTAL AND COMPUTER RESULTS: NATURAL SAMPLING
FIG 5.21 COMPARISON OF EXPERIMENTAL AND COMPUTER RESULTS: REGULAR SAMPLING
(5.5.3.2) **Operation at Integer Values of \( \frac{f_1}{f_0} \) Ratio**

**Output frequency = 16\( \frac{1}{3} \) Hz and 50 Hz**

The waveforms for output frequencies of 16\( \frac{1}{3} \) Hz and 50 Hz are shown in figures 5.22 and 5.23 respectively. These are equivalent to \( \frac{f_1}{f_0} \) Ratios of 9.0 and 3.0 which are the extremities of the tabulated computer results shown in table 4.1. In both cases, the waveforms for successive cycles of the wanted component are identical, and the fundamental frequency is therefore equal to the wanted output frequency. There are therefore no sub-harmonics, thus confirming the relevant computed results shown in tables 4.1a and 4.1f.

(5.5.3.3) **Operation at a Non-Integer Value of \( \frac{f_1}{f_0} \) Ratio**

**Output Frequency = 17.6 Hz**

The waveforms for an output frequency of 17.6 Hz, corresponding to a \( \frac{f_1}{f_0} \) Ratio of 8.5 are shown in figure 5.24. The two successive cycles, at wanted output frequency, of these waveforms are dissimilar. This is due to the presence of a sub-harmonic component whose magnitude and frequency, by wave analyser measurement, was 7% of the wanted component and 8.8 Hz (50% of the output frequency).

The waveforms shown in figure 5.24 for the experimental cycloconverter are very similar to those obtained from the computer (figure 4.6 in chapter 4), thus validating the computer programming technique.
FIG 5.22 \( f_0 = 16\frac{2}{3} \text{ Hz} \)

FIG 5.23 \( f_0 = 50.0 \text{ Hz} \)
FIG 5.24 $f_0 = 17.6$ HZ

FIG 5.25 $f_0 = 38.0$ HZ
(5.5.3.4) Effect of a Sub-harmonic Component having a Frequency considerably less than the Wanted Output Frequency

The computed results in table 4.1a in chapter 4 showed that for operation of the cycloconverter at a \( pf_1/f_o \) Ratio of 3.9, there is a high magnitude sub-harmonic component at 10% of the output frequency, whilst at a Ratio of 4.0, there is a high magnitude d.c. component. By interpolation between these results, there is a high magnitude sub-harmonic component of frequency less than 10% of the output frequency when the \( pf_1/f_o \) Ratio is between 3.9 and 4.0. For the experimental cycloconverter, these Ratios correspond to output frequencies of 38.5 Hz and 37.5 Hz respectively, and the waveforms for output frequencies of 38.0 Hz, 37.75 Hz and 37.5 Hz are discussed below.

Output Frequency = 38.0 Hz

Figure 5.25 shows the output waveforms of the experimental cycloconverter for an output frequency of 38.0 Hz, which corresponds to a \( pf_1/f_o \) Ratio of 3.95. It is seen that the waveform oscillates about the zero axis, and a sufficient number of output cycles is shown for a repetition period to be identified. By counting the number of cycles of the wanted component in one repetition period, it is seen that the frequency of oscillation of the waveforms is about 5% of the output frequency, or about 1.9 Hz. The computed results given in
chapter 4 confirm that this is the sub-harmonic component, and it is concluded that a low frequency sub-harmonic component causes an oscillation of the waveform such as is shown in figure 5.25.

**Output Frequency = 37.75 Hz**

With decrease in output frequency from 38.0 Hz to 37.75 Hz (corresponding to an increase of $\frac{f_1}{f_0}$ Ratio from 3.95 to 3.97), the computed results given in chapter 4 predict a decrease in the frequency of the high magnitude sub-harmonic component. Thus, in the waveforms for 37.75 Hz shown in figure 5.26, the number of output cycles for a repetition period to be identified is greater than in figure 5.25. By again counting the number of cycles of the wanted component in one repetition period, the frequency of the sub-harmonic component is seen to be about 2.8% of the output frequency, or about 1 Hz.

In both figures 5.25 and 5.26, the alternating pattern of light and dark areas is due to the variation in nett areas enclosed by the waveform during successive cycles of the wanted component. As discussed in chapters 3 and 4, this is a direct result of the presence of the sub-harmonic component.

**Output Frequency = 37.5 Hz**

When the output frequency of the cycloconverter is further reduced to 37.5 Hz (corresponding to a $\frac{f_1}{f_0}$ Ratio of 4.0), the waveforms are as shown in
FIG 5.26 $f_0 = 37.75$ HZ

FIG 5.27 $f_0 = 37.5$ HZ
figure 5.27. The variation in nett area enclosed by the waveform over a large number of cycles is negligible, but the lighter top half of the current trace is due to a consistent positive nett area indicating the presence of a d.c. component. The computed results confirm this observation.

Output Frequency = 35.0 Hz

With further reduction of cycloconverter output frequency to 35 Hz (pf₁/f₀ Ratio = 4.28), the output waveforms are as shown in figure 5.28. The variation in the nett areas over one repetition period indicate the presence of a sub-harmonic at about \( \frac{1}{3} \) of the output frequency, or about 12 Hz. This is again confirmed by the computer results.

(5.5.3.5) Effect of a Sub-harmonic Component having a Frequency near the Wanted Output Frequency

The computed results in tables 4.1b and 4.1c show that at a pf₁/f₀ Ratio of 5.0, there is no d.c. component and no sub-harmonic component. However, there is a significant general harmonic which coincides with the wanted output component at this pf₁/f₀ Ratio, and which emerges as a sub-harmonic component when the Ratio is increased, and which emerges as a super-harmonic component when the Ratio is decreased.

The effect of this harmonic on the output waveforms of the experimental cycloconverter when the output frequency is increased from 30.0 Hz (pf₁/f₀ Ratio = 5.0)
FIG 5.28  $f_0 = 35.0$ Hz

FIG 5.29  $f_0 = 30.0$ Hz
to 31.0 Hz \((pf_1/f_0\text{ Ratio } = 4.84)\) are shown in figures 5.29 - 5.31 and are discussed in detail below. A small decrease in output frequency gave similar waveforms.

**Output Frequency = 30.0 Hz**

Figures 5.29 and 5.30 both show the output waveforms of the experimental cycloconverter when the output frequency is 30.0 Hz. Any variation in the nett area enclosed by successive cycles of the wanted component is insignificant, either in figure 5.29 or for the larger number of output cycles in figure 5.30. There is therefore no sub-harmonic component in the output, as confirmed by the computed results in Chapter 4.

**Output Frequency = 31.0 Hz**

Figure 5.31 shows the effect of the emergence of the sub-harmonic component at a frequency just below the output frequency of the cycloconverter. The output frequency is 31.0 Hz, and the current waveform in particular, is characteristic of a waveform which results from amplitude modulation of two signals. The side-band frequency in figure 5.31 is about 4 Hz, and therefore the frequency of the modulating signal (the sub-harmonic component) is about \(31 \text{ Hz } - 4 \text{ Hz } = 27 \text{ Hz}\). This is very close to the computed frequency of 26.4 Hz.

The output waveforms of the cycloconverter for an output frequency of about 29 Hz were similar to those shown in figure 5.31. The computed results show that a super-harmonic component of the frequency just above
FIG 5.30  \( f_0 = 30.0 \text{ Hz} \)

FIG 5.31  \( f_0 = 31.0 \text{ Hz} \)
the output frequency is then present.

It is concluded from these observations concerning figure 5.31 that a sub-harmonic or a super-harmonic, having a frequency close to that of the output frequency, causes amplitude modulation of the output waveforms.
CHAPTER 6

CONCLUSIONS AND SUGGESTIONS FOR FURTHER WORK

The object of this thesis is to show that the application of sampling techniques enables a new approach to be made to the control of the inhibited cycloconverter, and thereby to show that the sub-harmonic components of the output voltage can be attenuated by appropriate modification of the control technique.

It was shown in chapter 3 that, although replacement of natural by regular sampling in the control of the circulating-current cycloconverter gave very significant attenuation of the sub-harmonic components, this was not generally the case for the inhibited cycloconverter. It is concluded that the inhibition process interferes with the sequence of commutations, defined by the thyristor switching functions, in such a way as to negate the improvements obtained when the cycloconverter is operated in the circulating-current mode. It was shown that significant improvements can be obtained in the inhibited cycloconverter by adding ramp functions to the regular samples in a similar manner to the process of pulse position modulation in communications engineering. However, this improvement was only obtainable whilst the load current was sinusoidal, that is whilst the inhibition switching instants are regular and the conducting periods of both thyristor groups are equal.
It was shown in chapter 4 that if the load current is non-sinusoidal, the inhibition switching instants are generally irregular, and the conducting periods of the thyristor groups are unequal. This was shown to have the effect of increasing the magnitude of the sub-harmonics, and it was then necessary to introduce a further modification to the control of the cycloconverter to reduce this irregularity, and thereby to attenuate the sub-harmonics. Ramp functions of variable slope in the pulse position modulation method referred to above gave some improvements, but the most significant improvements were obtained by distorting the reference wave with its second harmonic. The application of this modification was shown to result in very significant attenuations of the sub-harmonic components, and it would therefore make a useful contribution to improving the performance of a cycloconverter-motor system when the cycloconverter is operated over a wider range of input/output frequency ratios than at present.

It is suggested that the work be extended by further investigation into the use of distorted reference waves in the control system in order to optimize these improvements. Since the effect of the inhibition process, particularly the effect of irregular inhibition switching instants, is shown to accentuate the sub-harmonics, then it would also be useful to investigate the operation of the cycloconverter in a hybrid inhibited/
circulating-current mode. This would allow circulating-current to flow for a limited period whilst the load current is commutated from one thyristor group to the other, and could provide greater flexibility over the control of the inhibition switching instants than in the pure inhibited cycloconverter.

It is shown that the operation of the cycloconverter can be identified with pulse width modulation. Pulse width modulation is commonly used in communications engineering and it is suggested that the mathematical analysis in this context be related to the cycloconverter, with the possibility of deriving further benefits from the application of communications techniques.

Computer programs were written for both the circulating-current and inhibited cycloconverters as described in chapters 3 and 4. The computer analysis discussed in chapter 4 showed that a sub-harmonic component is identifiable as a general harmonic whose frequency depends on the value of the \( \text{pf}_1/f_0 \) Ratio, and that the general harmonic is a sub-harmonic component over a limited range of the \( \text{pf}_1/f_0 \) Ratio values; for even integer values of the \( \text{pf}_1/f_0 \) Ratio it takes the form of a d.c. component, whilst at odd integer values there is no sub-harmonic component and no d.c. component.

The value of the \( \text{pf}_1/f_0 \) Ratio was also shown to be significant in the prediction of the fundamental
repetition frequency, and two methods of determining its value for the computer analysis were developed and described in chapter 2.

It was shown in chapter 3 that the nett areas enclosed by successive cycles of the output waveform (at wanted output frequency) are zero if no sub-harmonic component is present. If a sub-harmonic is present, then the nett areas vary from one complete cycle to the next. This criterion was found to be of particular value for assessing the sub-harmonic contents of the computer plots of the output waveforms.

Practical tests on the experimental cycloconverter, which is described in chapter 5, confirmed the validity of the computer programs. The inhibition control system of the experimental cycloconverter incorporated a novel method of zero current detection involving optocouplers. The practical tests also showed that the effect of a sub-harmonic component on the shape of the output waveform depends on its frequency in relation to the wanted output frequency. If the difference between these frequencies is small, the envelope of the waveform is characteristic of amplitude modulation; if the difference is large, the waveform oscillates about its mean level at the sub-harmonic frequency.

Although the objectives of this thesis are concerned with the distortion of the output waveforms, distortion of the supply, or input, waveform is often
queried in practical rectifier installations. Future work should therefore include an investigation into the input side distortion components. This would inevitably introduce consideration of the source, or commutation, impedance which if of significant magnitude, could also affect the output waveforms. The effect of source impedance on the output distortion components should therefore be investigated.

Existing cycloconverter control systems invariably rely on such methods as the cosinusoidal control method in which two analogue input signals are used. The results discussed in this thesis have shown that such methods do not necessarily give optimum cycloconverter performance unless appropriate modifications are made. There are in fact an infinite number of combinations of thyristor switching instants if the constraint of the cosinusoidal control method is removed. The sequence of switching instants must nevertheless be controlled to give a predominant a.c. output component at the wanted output frequency, but there are still a very large number of possible combinations. It is therefore recommended that the requirements for an all-digital control system be investigated in order to provide greater flexibility of control over the thyristor switching functions and the inhibition switching functions. Such a system is now viable for practical cycloconverters with the availability of low-cost pre-programmable micro-processors. The thyristor switching instants can then be pre-programmed with a wide range of possible values.
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APPENDIX 1

COMPUTER PROGRAM FOR THE
CIRCULATING-CURRENT CYCLOCONVERTER
THYRISTOR CYCLOCONVERTER WITH SINGLE PHASE OUTPUT
CIRCULATING CURRENT MODE
COSINUSOIDAL CONTROL (NATURAL AND REGULAR SAMPLING)
RESISTIVE LOAD

REQUIRED INPUT DATA
NP=NUMBER OF INPUT (2 OR 3 IN THIS PROGRAM)
FR=INPUT FREQUENCY (Hz)
FO=OUTPUT FREQUENCY (Hz)
FF=FREQUENCY OF FUNDAMENTAL (LOWEST FREQUENCY IN OUTPUT)
FP/FF MUST BE EQUAL AN INTEGER
NH=HIGHEST ORDER OF HARMONIC IN THE OUTPUT

DUAL MODULATION DEPTH

ADDITIONAL SYMBOLS USED IN THIS PROGRAM
N=ORDER OF HARMONIC REFERRED TO FF
IC=COUNTER FOR SAMPLING
IC=COUNT FOR COMPUTATION FOR NATURAL SAMPLING FOLLOWED
BY REGULAR SAMPLING
KP=COMPUTATION POINT (=1 AT FIRST COMUTATION AFTER TH0)
NT=TOTAL NUMBER OF COMPUTATION POINTS IN THE COMPUTATION
TIME=INSTANT AT WHICH COMPUTATION IS TO BE CARRIED OUT
TS=TIME SECS AT START OF SINUSOIDAL VRTROL VOLTAGE
T=TIME OF START OF VC
C=TIME OF COMPUTATION KP
DT=INCREMENTAL TIME INTERVAL FOR COMPUTATION
VR=REFERENCE VOLTAGE
VC=COSINUSOIDAL CONTROL VOLTAGE
LH2=FOR POSITIVE GROUP
=1 FOR NEGATIVE GROUP
A/I=FOURIER COEFFICIENT FOR COSINE TERMS
D/I=FOURIER COEFFICIENT FOR SINE TERMS
C/I=FOURIER COEFFICIENT FOR COSINE TERMS OF POSITIVE GROUP
D/I=DITTO FOR SINE TERMS
E/I=FOURIER COEFFICIENT FOR COSINE TERMS OF NEGATIVE GROUP
F/I=DITTO FOR SINE TERMS
H/I=AMPLITUDE (PEAK, P.U.) OF THE NTH HARMONIC
W/DIFFERENCE IN ANGULAR FREQUENCY BETWEEN FI AND HARMONIC
WS=SUM OF DITTO

1P.U.=PEAK OF INPUT VOLTAGE

DIMENSION T(1000), A(1000), B(1000), C(1000), D(1000), E(1000), F(1000)
DIMENSION H(1000), NP(100), H(100), ANH(100), ANH(100), ANG(100)
READ(5, 199) HZ
199 FORMAT(15)
DO 90 JK=1, HZ
READ(5, 200) F, FC, FF, NH, NP, DH
200 FORMAT(3F10.0, 211D10.0)
DO 201 IC=1, 2
WRITE(6, 201)
201 FORMAT(*11, 5X, 'INPUT DATA')
WRITE(6,205)Np
205 FORMAT('"",NUMBER OF PULSES=',I3)
WRITE(6,205)F1
202 FORMAT('"",INPUT FREQUENCY=',F10.5,'Hz')
WRITE(6,203)F0
203 FORMAT('"",OUTPUT FREQUENCY=',F10.5,'Hz')
WRITE(6,204)FE
204 FORMAT('"",FUNDAMENTAL FREQUENCY=',F10.5,'Hz')
WRITE(6,207)NK
207 FORMAT('"",HIGHEST ORDER OF HARMONIC=',I3)
WRITE(6,206)DM
206 FORMAT('"",INFORMATION DEPTH=',F10.5,'p.u.')
IF(IN-1)96,06,97
C NATURAL SAMPLING
96 WRITE(6,209)
209 FORMAT('"",5X,"OUTPUT DATA FOR CIRCULATING CURRENT CYCLO, NATURAL 2 SAMPLING'!)
GO TO 98
C REGULAR SAMPLING
97 WRITE(6,210)
208 FORMAT('"",5X,"OUTPUT DATA FOR CIRCULATING CURRENT CYCLO, REGULAR 2 SAMPLING'!)
98 WRITE(6,215)
210 FORMAT('"",5X,"KP","Sx","TIME(SECS)'!)
215 FORMAT('"",5X,"NEGATIVE GROUP'!
C
PI=2*3.14159/7.0
F1=8.1/300.0/F1
NKP=1.0F1*(NP/F1+0.01)
NKPI=NP+1
C COMMUTATION POINTS
DO 5 L=1,2
DO 10 J=1,NKP1
6 TIME((NKP1=(NKP1+1)*NKPI+1)/F1
5 CONTINUE
10 CONTINUE
TIME=(I-1)*DT+TIME
IF(IN=1)93,03,64
C NATURAL SAMPLING
93 VR=V1*SIN(2.0*PI*FO*TIME)
GO TO 95
C REGULAR SAMPLING
94 VR=V1*SIN(2.0*PI*FO*TIME)
95 VC=(2*X-3)*COS(2.0*PI*FO*(I-1)*DT)
C IS THIS FOR POSITIVE OR NEGATIVE GROUP
IF(IN=1)15,15,16
C IS VC_EQUAL TO OR GREATER THAN, VR
15 IF(VC=VR)20,21,21
C IS VR_EQUAL TO OR GREATER THAN, VC
16 IF(VR=VC)20,21,21
20 CONTINUE
C
21 T(KP)=TIME
WRITE(6, 300) KP, TIME
300 FORMAT(*1, 4X, 15, 9X, F9.4)
CONTINUE

C FOURIER COEFFICIENTS
DO 40 N=1,NN
A(N)=0.0
B(N)=0.0
C DO 50 KP=2, NKP1
P=(C(NP-2)/4.0-CN+L)+2.0*PI/NP
WD=2.0*PI*(FI+H*FF)
WS=2.0*PI*(FI+H*FF)
C IS N=FI/FF
YH=N=FI/FF
ABSY=ABS(Y)
IF(ABSY.0.01) GO TO 48
47 A(N)=A(N)+COS(KP*(TP(KP-1))+R)-COS(KP+R))**FF/WS
1+COS((WD+T*(KP-1)+R))**FF/WD
B(N)=B(N)+SIN(KP*(TP(KP-1)+R))**FF/ND
2-(SIN(KP+R))**FF/WS
GO TO 50
48 A(N)=A(N)+(T*(KP)-T*(KP-1))+(SIN(R))**FF/1.0*(COS(4.0*PI*FI*TP(KP)+R)
3+COS(4.0*FI*TP(KP)+R))**FF**/4.0/IN/FF1
B(N)=B(N)+(T*(KP)-T*(KP-1))+(COS(KP))**FF/1.0*(SIN(4.0*FI*TP(KP)+R)
4+SIN(4.0*FI*TP(KP)+R))**FF/4.0/IN/FF1
C CONTINUE
C IS THIS FOR POSITIVE OR NEGATIVE GROUP
IF(L.0.51,51,52
C FOURIER COEFFICIENTS FOR NEGATIVE GROUP
51 E(N)=A(N)
F(N)=B(N)
C HARMONICS DUE TO NEGATIVE GROUP
HH(N)=SQRT(E(N)**L(N)+F(N)**L(N))
ANGH(N)=180.0/PI*ATAN2(E(N), F(N))
GO TO 40
C FOURIER COEFFICIENTS FOR POSITIVE GROUP
52 C(N)=A(N)
D(N)=B(N)
C HARMONICS DUE TO POSITIVE GROUP
HP(N)=SQRT(C(N)**L(N)+D(N)**L(N))
ANGP(N)=180.0/PI*ATAN2(C(N), D(N))
40 CONTINUE
C WRITE(6, 315)
315 FORMAT(*1, 'POSITIVE GROUP')
5 CONTINUE
C DO 60 H=1, NN
C FOURIER COEFFICIENTS OF COMPLETE CYCLOCONVERTER
A(N)=0.0
B(N)=0.0
C HARMONICS DUE TO COMPLETE CYCLOCONVERTER
H(N)=SORT(A(N)**2*B(N)**2)
ANGC(N)=180.*1I+ATAN2(A(N),B(N))
60 CONTINUE
   IF(11.1)30,80,81
C   NATURAL SAMPLING
80 WRITE(0,340)
340 FORMAT('1', 'CIRCULATING CURRENT CYCLO', ' NATURAL SAMPLING')
   GO TO 82
C   REGULAR_SAMPLING
81 WRITE(0,341)
341 FORMAT('1', 'CIRCULATING CURRENT CYCLO', ' REGULAR SAMPLING')
C   OUTPUT FREQUENCY EXPRESSED AS HARMONIC OF THE FUNDAMENTAL
82 NO=FIX(F0/FF+0.01)
C   OUTPUT VOLTAGE (AT OUTPUT FREQUENCY)
   VN=HR(NO)*100.0
   VP=HP(NO)*100.0
   VO=H(NO)*100.0
   WRITE(0,347)
347 FORMAT('1', 'VOLTAGES AT OUTPUT FREQUENCY')
   WRITE(0,348)VN
348 FORMAT('1', 'YV=1.1', F0, '.2', 1X, '0/0 OF INPUT VOLTAGE')
   WRITE(0,349)VP
349 FORMAT('1', 'YP=1.1', F0, '.2', 1X, '0/0 OF INPUT VOLTAGE')
   WRITE(0,350)VO
350 FORMAT('1', 'YV=1.1', F0, '.2', 1X, '0/0 OF INPUT VOLTAGE')
   WRITE(0,360)
360 FORMAT('1', 'HARMONICS IN OUTPUT(0/0 OF VOLTAGE),3X, PHASE ANGLE (DEGS) OF HARM')
   WRITE(0,361)
361 FORMAT('1', '7X,AT OUTPUT FREQUENCY'), 18X, 'AT INPUT FREQUENCY', 1X, 'FROM T=0, 1 REFERRED TO')
   WRITE(0,370)
370 FORMAT('1', '2X', 'FREQ', '2X', 'POSITIVE', 'X', 'NEGATIVE', 'X', 'COMPLETE', 'X', 'INPUT FREQ')
   WRITE(0,360)
380 FORMAT('1', '3X', 'FREQ', '4X', 'GROUP', '4X', 'GROUP', '4X', 'Cyclo', '6X', 'GROUP', '4X', 'Cyclo')
C   EXPRESS HARMONICS AS 0/0 OF OUTPUT VOLTAGE
   DU .9, N=1,10
   HN(N)=H(N)*100.0/VN*100.0
   HP(N)=HP(N)*100.0/VP*100.0
   H(N)=H(N)*100.0/VO*100.0
   FREC=NPFF
C   PHASE ANGLES REFERRED TO INPUT FREQUENCY
   ANGPF=FI/FRE*ANGP(N)
   ANGPF=FI/FRE+ANGP(N)
   ANGPF=FI/FRE-ANGP(N)
   WRITE(0,400)FRE, HP(N), HN(N), H(N), ANGPF(N), ANGPF(N), ANGPF(N), ANGPF(N), ANGPF(N), ANGPF(N)
400 FORMAT('1', 'F0, ', 7X, '5X', 'F0, ', 5X, 'F0, ', 5X, 'F0, ', 5X, 'F0, ', 5X, 'F0, ', 5X, 'F0, ', 5X, 'F0, ')
70 CONTINUE
91 CONTINUE
C   90 CONTINUE
STOP
END
APPENDIX 2

APPLICATION OF FOURIER ANALYSIS TO THE OUTPUT VOLTAGE WAVEFORM OF THE CYCLOCONVERTER

Basic Principle

The basic principle underlying Fourier Analysis is that any non-sinusoidal waveform can be separated into a number of sinusoidal and cosinusoidal components.

If a non-sinusoidal waveform is defined as a periodic function $F(t)$ during the period $t_1$ to $t_2$, then

$$F(t) = A_0 + \sum_{n=1}^{\infty} A_N \cos(2\pi N f_f t) + \sum_{n=1}^{\infty} B_N \sin(2\pi N f_f t) \quad (A2.1)$$

where $A_0 = \frac{1}{t_2-t_1} \int_{t_1}^{t_2} F(t) \, dt \quad (A2.2)$

$$A_N = 2f_f \int_{t_1}^{t_2} F(t) \cos(2\pi N f_f t) \, dt \quad (A2.3)$$

$$B_N = 2f_f \int_{t_1}^{t_2} F(t) \sin(2\pi N f_f t) \, dt \quad (A2.4)$$

$f_f$ = fundamental repetition frequency $= \frac{1}{t_2-t_1}$

Adaption of the Basic Principle of the Cycloconverter

(a) Fourier Coefficient, $A_0$

If the voltage of the input supply to the cycloconverter is 1 p.u.,

$$F(t) = \sin(2\pi f_1 t + R) \quad (A2.5)$$

where $f_1$ = input frequency

and $R$ depends on which input phase is being
utilised at any particular instant to fabricate the output voltage waveform.

From Equations (A2.2) and (A2.5)

\[ A_o = \frac{f_1}{2\pi f_i} \int_{t_1}^{t_2} \sin(2\pi f_i t + R) \, dt \]

\[ = - \frac{f_1}{2\pi f_i} \left[ \cos(2\pi f_i t + R) \right]_{t_1}^{t_2} \]

The output waveform consists of segments of the input sine waves. Each segment can be defined between the time limits \( t(K-1) \) and \( t(K) \) where \( K \) successively takes the values 1, 2, 3, ..., up to the total number \( N_K \) of switching instants in one cycle of fundamental frequency.

\[ \therefore A_o = \sum_{K=1}^{N_K} \frac{f_1}{2\pi f_i} \left[ \cos(2\pi f_i T(K-1) + R) - \cos(2\pi f_i T(K) + R) \right] \quad (A2.6) \]

(b) Fourier Coefficients of the Cosine Terms \( A_N \)

From equation (A2.3), quantity to be integrated

\[ = F(t) \cos(2\pi N f_i t) \]
\[ = \sin(2\pi f_i t + R) \cos(2\pi N f_i t) \quad \text{(from (A2.5))} \]
\[ = \frac{1}{2} \sin(W_D t + R) + \frac{1}{2} \sin(W_S t + R) \quad (A2.7) \]

where
\[ W_D = 2\pi (f_i - N f_i) \]
\[ W_S = 2\pi (f_i + N f_i) \]
\[ A_N = \int_{t_1}^{t_2} \left[ \sin(W_D t + R) + \sin(W_S t + R) \right] \, dt \]

This expression can be re-written as:

\[ A_N = \int_{t_1}^{t_2} \left[ \frac{1}{W_D} \cos(W_D t + R) + \frac{1}{W_S} \cos(W_S t + R) \right] \, dt \]

In the special case when \( f_1 = N f_f \) giving \( W_D = 0 \), the above expression cannot apply.

\[ A_N = \sum_{K=1}^{N_K} \left[ \cos(W_D t(K-1) + R) - \cos(W_D t(K) + R) \right] \]

\[ + \sum_{K=1}^{N_K} \left[ \cos(W_S t(K-1) + R) - \cos(W_S t(K) + R) \right] \]

In the special case when \( f_1 = N f_f \) giving \( W_D = 0 \), the above expression cannot apply.

\[ A_N = \sum_{K=1}^{N_K} f_f \left[ t(K) - t(K-1) \right] \sin(R) \]

\[ + \sum_{K=1}^{N_K} \frac{f_f}{4 \pi f_1} \left[ \cos(4 \pi f_1 t(K-1) + R) - \cos(4 \pi f_1 t(K) + R) \right] \]
(c) Fourier Coefficients of the Sine Terms ($B_N$)

A similar procedure to that set out in (b) above gives the following expressions for $B_N$ derived from (A2.4) and (A2.5):

For $N \neq \frac{f_0}{f_1}$,

$$B_N = \sum_{K=1}^{N_K} \frac{f_1}{w_D} \left[ \sin(w_D t(K) + R) - \sin(w_D t(K-1) + R) \right]$$

$$- \sum_{K=1}^{N_K} \frac{f_1}{w_S} \left[ \sin(w_S t(K) + R) - \sin(w_S t(K-1) + R) \right]$$

(A2.11)

For $N = \frac{f_0}{f_1}$,

$$B_N = \sum_{K=1}^{N_K} f_1 [t(K) - t(K-1)] \cos (R)$$

$$- \sum_{K=1}^{N_K} \frac{f_1}{4 f_1} \left[ \sin(4 \pi f_1 t(K) + R) - \sin(4 \pi f_1 t(K-1) + R) \right]$$

(A2.12)

(d) Magnitudes and Phase Angles of the Harmonics ($H_N$, $\Theta_N$)

$$H_N = (A_N^2 + B_N^2)^{\frac{1}{2}}$$

(A2.13)

$$\Theta_N = \tan^{-1} \frac{B_N}{A_N}$$

(A2.14)
APPENDIX 3

CURRENT FOR AN INHIBITED CYCLOCONVERTER FEEDING A RESISTIVE-INDUCTIVE LOAD

The voltage across the load at any instant is given by

\[ v = K \sin(wt + \phi) \]

where \( K \) = amplitude of input voltage
\( \phi \) = phase angle defining which input phase voltage is conducting
\( w \) = frequency of the input voltage to the cycloconverter

also \( v = iR + L \frac{di}{dt} \)

where \( R \) = resistance of the load
\( L \) = inductance of the load
\( i \) = instantaneous load current

\[ \therefore K \sin(wt + \phi) = iR + L \frac{di}{dt} \]

Taking Laplace Transforms:

\[ \frac{K(w \cos \phi + s \sin \phi)}{s^2 + w^2} = (R + Ls) i - Li(0) \]

\[ \therefore i = \frac{K(w \cos \phi + s \sin \phi) + (s^2 + w^2)Li(0)}{(s^2 + w^2)(R + Ls)} \]

Putting (A3.1) = \( \frac{As + B}{s^2 + w^2} + \frac{C}{R + Ls} \)

(A3.2)

gives: \( Kw \cos \phi + w^2Li(0) + Ks \sin \phi + s^2Li(0) \)

\[ = (As + B)(R + Ls) + C(s^2 + w^2) \]

\[ = ARs + BLS + BR + ALS^2 + Cs^2 + Cw^2 \]
equating $s^0$, $s^1$ and $s^2$:

$$K \cos \phi + w^2 L(0) = BR + Cw^2$$  \hspace{1cm} (A3.3)

$$K \sin \phi = AR + BL$$  \hspace{1cm} (A3.4)

$$L(0) = AL + C$$  \hspace{1cm} (A3.5)

from (A3.3) and (A3.5):

$$K \cos \phi + w^2 L(0) = BR + (L(0) - AL)w^2$$

$$\therefore \quad K \cos \phi = BR - ALw^2$$  \hspace{1cm} (A3.6)

from (A3.4) and (A3.6):

$$K R \cos \phi + w^2 LK \sin \phi = BR^2 + BL^2w^2$$

$$\therefore \quad B = \frac{K(wR \cos \phi + w^2 L \sin \phi)}{R^2 + w^2 L^2}$$  \hspace{1cm} (A3.7)

$$\therefore \quad A = \frac{K}{R} (\sin \phi - \frac{wL \cos \phi + w^2 L^2 \sin \phi}{R^2 + w^2 L^2})$$  \hspace{1cm} (A3.8)

from (A3.5) and (A3.8):

$$C = L(0) - \frac{K}{R^2 + w^2 L^2} (RL \sin \phi - wL^2 \cos \phi).$$  \hspace{1cm} (A3.9)

Substituting (A3.7), (A3.8) and (A3.9) in (A3.2), and rearranging, gives:

$$I = \frac{K}{(R^2 + w^2 L^2)} \left[ \frac{R(w \cos \phi + s \sin \phi)}{s^2 + w^2} + \frac{wL(w \sin \phi - s \cos \phi)}{s^2 + w^2} + \frac{wL \cos \phi - R \sin \phi + (R^2 + w^2 L^2) i(0)}{s + R/L} \right]$$

A3/2
transforming back to time function gives:

\[
1 = \frac{K}{R^2 + w^2L^2} \left[ R \sin (wt + \phi) - wL \cos (wt + \phi) \\
+ \exp(-\frac{R}{L}t). (wL \cos \phi - R \sin \phi + (R^2 + w^2L^2)i(0)) \right]
\]

Re-arranging and putting \( \frac{L}{R} = t_c \):

\[
1 = \frac{K}{R(1 + w^2t_c^2)} \left[ \sin (wt + \phi) - wt_c \cos (wt + \phi) \\
+ \exp(-\frac{t}{t_c}). (wt_c \cos \phi - \sin \phi) \right] \\
+ \exp(-\frac{t}{t_c}). (i(0))
\]

This expression gives the current in a load having time constant \( t_c \), due to the voltage \( K \sin (wt + \phi) \).
APPENDIX 4

COMPUTER PROGRAM FOR THE INHIBITED CYCLOCONVERTER
THYRISTOR CYCLOCONVERTER WITH SINGLE PHASE OUTPUT
CIRCULATING CURRENT-FREE MODE
COSINUSOIDAL CONTROL
SINUSOIDAL REFERENCE WAVE
NATURAL SAMPLING
R=1 LOAD
GROUP CHANGEOVER AT COMPUTATION POINT NEAREST
INSTANT OF ZERO OUTPUT CURRENT AFTER ZERO VR
REQUIRED INPUT DATA
NP=NUMBER OF INPUT (=2, 3 OR 6 IN THIS PROGRAM)
F1=INPUT FREQUENCY (Hz)
F2=OUTPUT FREQUENCY (Hz)
FF=FREQUENCY OF FUNDAMENTAL (LOWEST FUNDAMENTAL IN OUTPUT)
FO/FF MUST EQUAL AN INTEGER (=FA)
RP#1/FF MUST EQUAL AN INTEGER
RS=HIGHEST ORDER OF HARMONIC IN THE OUTPUT
DM=MODULATION DEPTH

ADDITIONAL SYMBOLS USED IN THIS PROGRAM
ND=NUMBER OF DATA CARDS (EXCLUDING CARD STATING ND)
NH=ORDER OF HARMONIC REFERRED TO FF
NPCI=COMPUTATION POINT (=1 AT FIRST COMPUTATION AFTER T=0)
NPK=NUMBER OF KP IN CC MODE
K=COMPUTATION POINT FOR CHANGEOVER IN INHIBITED CYCLOCONVERTER
KCI IS FOR CHANGEOVER FROM POSITIVE TO NEGATIVE GROUP
FOURIER ANALYSIS IS FROM KC=2 TO AVOID START ERRORS
KCL=LAST KC
KCI(KC)=POSITIVE GROUP KP NEXT TO KC
KCI(KC)=NEGATIVE GROUP KP NEXT TO KC
KP=Dummy KP FOR START OF FOURIER ANALYSIS FOR ONE GROUP
KP=Dummy KP FOR FINISH OF BITTO
TIME=INSTANT AT WHICH COMPUTATION IS TO BE CARRIED OUT
TILE=.05 SECS AT START OF SINUSOIDAL REFERENCE VOLTAGE
TINC=TIME OF START OF VCC
TINC=TIME INTERVAL FOR COMPUTATION
TP(KP)=TIME OF COMPUTATION OF POSITIVE GROUP
TN(KP)=TIME OF COMputation of NEGATIVE GROUP
TC(KC)=TIME OF START OF COMPUTATION AT
CHANGEOVER COMPUTATION POINT KC
VR=REFERENCE VOLTAGE
AL=PROPORTIONAL TO THE LOAD CURRENT
ANG(AO)=COMPUTED INPUT PHASE SHIFT OF AO FROM VR
VC=COSINUSOIDAL CONTROL VOLTAGE
L=2 FOR POSITIVE GROUP
L=1 FOR NEGATIVE GROUP
A(N)=FOURIER COEFFICIENT FOR SINE TERMS
A(N)=FOURIER COEFFICIENT FOR COSINE TERMS
H(N)=MAGNITUDE PEAK, P.U., OF THE NTH HARMONIC
WS=Difference in angular frequency between F1 AND HARMONIC
WS=SUM OF BITTO

1. PEAK OF INPUT VOLTAGE

DIMENSION TP(500), TH(500), TC(120), KPP(120), KPN(120), T(500)
DIMENSION R(500), AINC(300), ALOAD(5000), TLOAD(5000)
READS(5, 159, 22)

199 FORMAT(15)
DO 90 J=1, NP
READS(5, 200) FT, RA, FA, IJ, HP, 00, TIMCO
90 FORMAT(5F10.0, F10.0)
F0N=F1/RA
IF=FO/FA

C

WRITE(6, 123)
218 FORMAT(6, 123(START))
WRITE(6, 201)
201 FORMAT(6, 15X, 'INPUT DATA')
WRITE(6, 205) NP
205 FORMAT(0, 'NUMBER OF PULSES' = 13)
WRITE(6, 202) FT
202 FORMAT(1, 'INPUT FREQUENCY' = F10.5, 'Hz')
WRITE(6, 213) RA
213 FORMAT(1, 'RATIO' = F10.5)
WRITE(6,203) FO
203 FORMAT(' ', 'OUTPUT FREQUENCY= ',F10.5, ', 'Hz')
WRITE(6,204) FF
204 FORMAT( ' ', 'FUNDAMENTAL FREQUENCY= ',F10.5, ', 'Hz')
WRITE(6,206) HH
206 FORMAT( ' ', 'HIGHEST ORDER OF HARMONIC=',I3)
WRITE(6,206) HH
206 FORMAT( ' ', 'MODULATION DEPTH=',F10.5, ', 'P.u.')
WRITE(6,212) TMC
212 FORMAT( ' ', 'TIME CONSTANT= ',F10.7, ', 'SECS')
WRITE(6,214) TMC
214 FORMAT( ' ', 5X, '*** COSINE CONTROL, SINUSOIDAL REFERENCE, NATURAL SAMPLING ***')
C
D 91 I C=1, 5
C
CONSTANTS
PI=3.141592653589793238462643
D=1.0/FI/1000.0
NKP=IFIX(KP*FI/FF+0.01)+IFIX(3*KP*FI/FO+0.01)
NKP=NKP-1
N=IFIX(FI/FF+0.01)
KCL=2+2*10
WP=0.0*PI*FI
W0=0.0*PI*FO
NA=IFIX((KCL/2.0/FO/DT/4.0)
IF(KC, G.E., 110) GO TO 66
IF(KP, G.E., 500) GO TO 66
IF(KP, G.E., 500) GO TO 66
GO TO 85
86 WRITE(6,215)
215 FORMAT(1X, ' 5('DIMENSION EXCEEDED', 5X))
GO TO 90
85 CONTINUE
C
WRITE(6,219)
219 FORMAT(' ', 12('NEXT'))
IF(CC-2), 26, 93, 3, 94
96 WRITE(6,205)
205 FORMAT(1X, 5X, 'OUTPUT DATA FOR ', 8('VR/..0', 3X))
GO TO 221
93 WRITE(6,219)
209 FORMAT(0, 5X, 'OUTPUT DATA FOR ', 8('VR/..120', 3X))
GO TO 221
94 WRITE(6,220)
220 FORMAT(0, 5X, 'OUTPUT DATA FOR ', 8('VR/..240', 3X))
221 WRITE(6,211)
211 FORMAT(0, 'COMPUTATION POINTS FOR UNINHIBITED CYCLOCONVERTER')
WRITE(6,216)
210 FORMAT(1X, 2X, 'GROUP', 5X, 'KP', 6X, 'TIME(SECS)', 5X, 'MODULATION(SECS)')
C
L=1, 2
GO TO 10
KP=1, KP1
IF(KP=3), 3, 77, 78, 79
77 TIME=(KP-1)/2.0/FI
GO TO 92
78 TIME=(2*KP-4*L)/6.0/FI
GO TO 92
79 TIME=(KP-3+1)/6.0/FI
92 GO TO 1
TIME=(1-1)+TIME
VR=(24*3)+COS(2.0*P1+I-1)/1000.0
VR=VR*SLN(2.0*P1+TIME/(1-1)*2.0*P1/3.0)
C
FOR POSITIVE OR NEGATIVE GROUP
IF(KP=1), 15, 15, 16
15 IF(VR=VR) 20, 21, 21
20 CONTINUE
WRITE(6,230)
230 FORMAT(1X, 'NO COMPUTATION')
1 IF(KP)=TIME
TIMOD=TIME=TIME
WRITE (6,300)KP,TIMED,TIMOD
300 FORMAT('I', 'POSITIVE', 6X, I3, 8X, F7.4, 8X, F7.4)
GO TO 10
22 TN(KP)=TIME
TIME=TIME-TIMOD
WRITE (6,310)KP,TIMED,TIMOD
310 FORMAT('I', 'NEGATIVE', 6X, I3, 8X, F7.4, 8X, F7.4)
C
10 CONTINUE
5 CONTINUE

TIMES FOR CHANGEOVER IN INHIBITED CYCLOCONVERTER

L=2
IA=0
TIMI=0.0
KCLI=KCL+1
DO 150 KC=1, KCLI
AQ=0.0
TIMVZ=KC/2.0/F0+(IC-1)/3.0/F0
IF(KC.EQ.1) GO TO 141
TIMI=TC(KC-1)
141 IF(L-1)154, 154, 151
C
FIND FIRST KP OF INCOMING(POSITIVE) GROUP
150 IF(1P(KP) .GE. TIMI) GO TO 152
151 CONTINUE
152 IF(KP(KP-1) .GE. 1) GO TO 142
KP(KP(KP-1))=KP
GO TO 142
C
FIND FIRST KP OF INCOMING(NEGATIVE) GROUP
154 IF(KP(KP-1), KPI)
IF(KP(KP) .GE. TIMI) GO TO 149
156 CONTINUE
149 KP(KP(KP-1))=KP
C
142 KP1=KP
GO 146. KP=KP1. 1000
146 IF(KP1)153, 155, 151
153 KP=KP1
GO TO 161
155 K=(4*IC(P-KP1)+L+PI/6.0)
GO TO 161
160 IC=IC(KP(KP-1)+PI/3.0
161 TIMI=TIMI
162 IF(TIMI .LT. 0.00001) GO TO 155
AL=(0.07/((T+1)*T+)2)*SIN(WT+T+R)+0.00001
165 CUS(WT+T+R)+(EXP(T+T+R-TIMI)+TIMI)*COS(WT+T+R-TIMI)+
164 2.0+T+R-TIMI)))+4.0+EXP(TIMI+T+R-TIMI)
GO TO 165
155 AL=SIN(WT+T+R+R)+0.00001
157 IF(KP(KP-1) .LT. 0.00001) GO TO 164
166 IF(AL=0.00001) 164, 170, 170
169 IF(AL<0.00001) GO TO 164
170 AL=(2*IC(P) .LT. 0.00001)
164 TIMI=TIMI+4.0
164 IF(KP(KP-1) .LT. 0.00001) GO TO 163
164 TIMI=TIMI+4.0
163 IF(KP(KP-1) .LT. 0.00001) GO TO 165
165 TIMI=TIMI+4.0
165 IF(KP(KP-1) .LT. 0.00001) GO TO 165
165 TIMI=TIMI+4.0
165 CONTINUE
C
165 IA=IA+1
ALOAD(IA)=AL
TL=LOAD(IA)=TIMI+DT+4.0
GO TO 162
C
CHANGEOVER OCCURS HOW

A4/4
C RECORD LAST KP OF OUTGOING GROUP AND THEN CHANGE TO OTHER GROUP
171 KPN(KC)=KP-1
GO TO 167
166 KPF(KC)=KP-1
L=1
167 CONTINUE
C 150 CONTINUE
C ILAST=IA
C FOURIER ANALYSIS
DO 67 N=1,NN
A=0.0
C=0.0
OCCUP=0.0
L=2
1KP1=2
C DO 53 KC=1,KC1
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)59,59,63
C POSITIVE GROUP
58 T(KP)=TP(KP)
T(KP-1)=TP(KP-1)
GO TO 64
C NEGATIVE GROUP
59 T(KP)=IN(KP)
T(KP-1)=IN(KP-1)
64 T(KPS)=TS
T(KPF)=TF
C CHECK FOR CURRENT DISCONTINUITIES
ISTART=1KP1-1
DO 60 IA=1START,ILAST
IF(LOAD(IA).GE.T(KP-1)) GO TO 61
60 CONTINUE
WRITE(6,328)
328 FORMAT('U','ERROR?')
61 1KP1=IA
GO 62 IA=1KP1,ILAST
IF(LOAD(IA).GE.T(KP)) GO TO 63
62 CONTINUE
WRITE(6,329)
329 FORMAT('U','ERROR??')
63 1KP2=IA-1
1KPP=1KPF+1
ITC(KP)=C1,1KP2) GO TO 69
IF(L.EQ.2) GO TO 42
IF(LOAD(1KPP)+0.00002)=43,43,44
42 IF(LOAD(1KPP)-0.00002)=44,44,45
GO 45 IA=1KPP,1KP2
IF(L.EQ.2) GO TO 46
IF(LOAD(IA)+0.00002)=45,45,45
46 IF(LOAD(IA)=0.00002)=45,45,45
45 CONTINUE
T(KP-1)=T(KP)
GO TO 69
49 T(KP-1)=LOAD(IA-1)
43 IF(L IA, 2) G0 TO 69
IF(LOAD(IA)+5.00002)65,65,68
66 IF(LOAD(IA) = 5.00002)65,65,68
65 CONTINUE
GO TO 69
68 T(KP)=LOAD(IA)

69 IF(NP=3)76,73,74
76 R=(L-KP)*PI
GO TO 75
73 R=(4*(L-KP)+1)*PI/6.0
GO TO 75
74 R=(2*(L-KP)+1)*PI/3.0
75 W=2.0*PI*(0.1+FF)
GCOMP=GCOMP+FF/2.0/PI/FI*(COS(2.0*PI*FI*1+T(KP-1)+R)-COS(2.0*PI*FI*1+T(KP)+R))

COEFFICIENTS

15 N=FI/FF
V=N=FI/FF
ABS=ABS(Y)

47 A=A+(COS(W)*T(KP-1)*R)-COS(W*F(T(KP)+R))*FF/WS
1+(COS(W)*T(KP)+R)-COS(W*F(T(KP)-R))*FF/WS
2=ABS(W)*T(KP+R)-ABS(W)*T(KP-1)*R)*FF/WS

48 A=A+(T(KP)-(-TP-1)+((SIN(R)+R)*FF/1.0-((COS(4.0*PI+FI*1+T(KP)+R)))+ FF/1.0))
67 CONTINUE

END FOURIER COEFFICIENTS

OUTPUT VOLTAGE (AT OUTPUT FREQUENCY)
V0=V(0)+100.0
WRITE(6,350)VO
350 FORMAT('15X', 'VOLTAGE AT OUTPUT FREQUENCY')

PHASE ANGLE (AT OUTPUT FREQUENCY)
WRITE(6,352)ANGC(0.0)
352 FORMAT('15X', 'PHASE DIFF BETWEEN NO AND VR')

PSF=ANGC(0.0)*FI/FO
WRITE(6,354)PSF
354 FORMAT('15X', 'PHASE DIFF BETWEEN NO AND VR')

DC COMPONENT
GCOMP=GCOMP+10000.0/VO
WRITE(6,359)GCOMP
359 FORMAT('C0', 'DC COMPONENT')

WRITE(6,360)
360 FORMAT('U', 'HARMONICS IN OUTPUT')
WRITE(6,370)
370 FORMAT('U', 'HARMONICS IN OUTPUT')
WRITE(6,371)

A4/6
371 FORMAT(' ',28X,'B(N)',7X,'F1',7X,'F0')
C C EXPRESS HARMONICS AS 0/0 OF OUTPUT VOLTAGE AND CALCULATE DISTORTION FACTOR
C SUSO=0.0
SUSQ1=0.0
DO 70 N=1,NH
H(N)=H(N)*100.0/V0*100.0
FREQ=N*F0
ANGF1=ANGC(N)+F1/FREQ
ANGFO=ANGC(N)+F0/FREQ
WRITE(6,400,FREQ,H(N),ANGC(N),ANGF1,ANGFO)
400 FORMAT(' ',5(1X,'F10.3'))
SUSO=SUSO+H(N)*H(N)
SUSQ1=SUSQ1+(H(N)*F0/FREQ)**2
70 CONTINUE
C DF=100.0/SQRT(SUSO)
DF=100.0/SQRT(SUSQ1)
WRITE(6,390)DF
390 FORMAT(' ',7X,'UNWEIGHTED DISTORTION FACTOR=',F10.4)
WRITE(6,391)DF
391 FORMAT(' ',7X,'WEIGHTED(F0/FH) DISTORTION FACTOR=',F10.4)
C C 91 CONTINUE
C 90 CONTINUE
STOP
END
APPENDIX 5

MAIN COMPONENTS USED IN THE EXPERIMENTAL CYCLOCONVERTER

Thyristors: BTY87 - 800R
Heat sinks: 5" x 4", 12 x 1" fins, aluminium
Fuses: Water immersed 2A fuse wire
Circulating-current Reactors: 200 mH, centre tapped
Diodes: IN4148
       AAZ15
       BAX13
Encapsulated Modules:
   4NOR60 logic gates: 2 x 2 input + 2 x 3 input
   UPA61 thyristor drive oscillator
   TT61 thyristor output transformer
CMOS Integrated Circuits:
   14001 NOR-gate, 4 x 2 input
   14011 NAND-gate, 4 x 2 input
   14025 NOR-gate, 3 x 3 input
Operational Amplifier Integrated Circuits:
   741 (general purpose)
   301 (high slew rate)
Optocouplers: TIL111
Transistors: BCY70
           BC107
           ST241
           B148
Transformers: 0-240V/20-12-0-12-20V
APPENDIX 6

INTERCONNECTION OF LOGIC GATES TO FORM MONOSTABLE AND BISTABLE CIRCUITS

(A6.1) Monostable Circuits

Monostable circuits forming part of the overall control circuit of the experimental cycloconverter are shown in figures A6.1 and A6.2. That shown in figure A6.1 consists of NAND-gates, and produces a normally logic 1 level output with short duration logic 0 level pulses coincident with negative-going transitions of the input voltage. The circuit shown in figure A6.2 consists of NOR-gates, and produces a normally logic 0 level output with short duration logic 1 level pulses coincident with positive-going transitions of the input voltage.

The waveforms at the different points @ to ¶ around the circuit are shown in the figures. The leading edges of the short duration pulses occur when the input voltage changes polarity (or changes state), whilst the trailing edges occur when the voltage across the capacitor of the R-C circuit reaches the threshold value of the input ¶ of the NAND-gate or of the input @ of the NOR-gate. The time taken to reach this threshold voltage is determined by the time constant of the R-C circuit, and therefore the time constant determines the width of the short duration pulses.

The discharge of the capacitor is not utilised
FIG A6.1 MONOSTABLE CIRCUIT
DIAGRAM AND WAVEFORMS:
NAND-GATES

A6/2
FIG A6.2 MONOSTABLE CIRCUIT
DIAGRAM AND WAVEFORMS:
NOR-GATES

A6/3
in the monostable circuit, and therefore the diode/resistance branch is added to accelerate it.

(A6.2) Bistable Circuits

Figures A6.3a and A6.3b show the interconnection of NOR-gates and NAND-gates respectively to form the bistable circuits used in the experimental cycloconverter. The truth tables show the logic state of the output terminal \( Q \) for all combinations of the logic states of the input terminals \( R \) and \( S \) and for changes in the logic state of \( R \) and \( S \). The logic state of \( Q \) is always the complement of \( \overline{Q} \).
a) USING NOR-GATES

b) USING NAND-GATES

FIGA6.3 RS BISTABLE CIRCUITS: CONNECTION DIAGRAM AND TRUTH TABLES