APPLICATION   OF   SAMPLING   TECHNIQUES   TO   THE

PHASE-CONTROLLED   THYRISTOR   CYCLOCONVERTER

by

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A thesis submitted to the CNAA for the Degree of

DOCTOR   OF   PHILOSOPHY

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June 1976

CONTENTS

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.

LIST OF FIGURES AND TABLES

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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
μF	Microfarads
Hz	Hertz
kHz	Kilohertz
p	Pulse number of the cycloconverter circuit
r	Modulation factor
s	Seconds
ms	Milliseconds
μs	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

∝	Firing angle (or trigger, or delay angle)
o	Degrees (angular)
Ω	Ohms
kΩ	Kilo-ohms
MΩ	Megohms

CHAPTER 1

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

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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 (Ford[12]) 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

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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
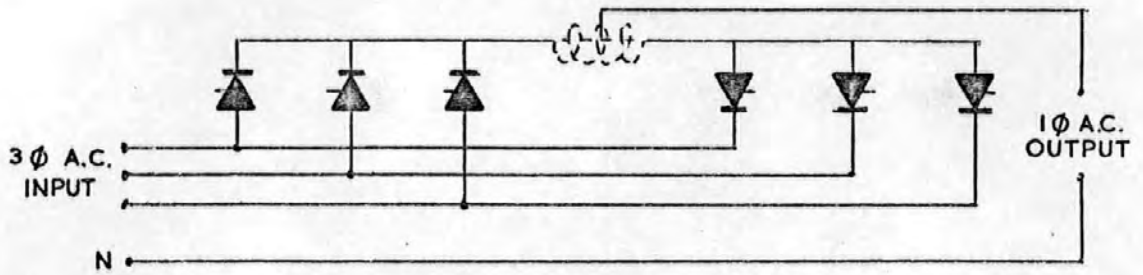
a) 3-PULSE CIRCUIT
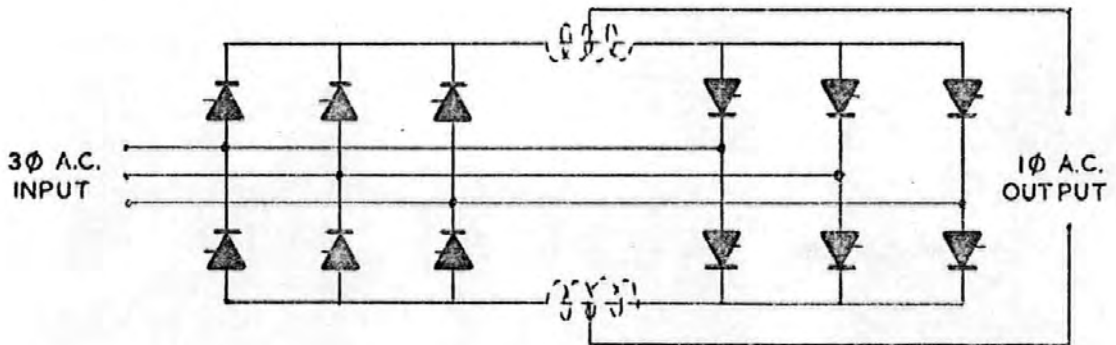
b) 6-PULSE CIRCUIT

FIG 2·1 THYRISTOR CONFIGURATIONS IN
3-PHASE CYCLOCONVERTERS

WANTED D.C. OUTPUT COMPONENT

a) RECTIFIER OPERATION

WANTED A.C. OUTPUT COMPONENT
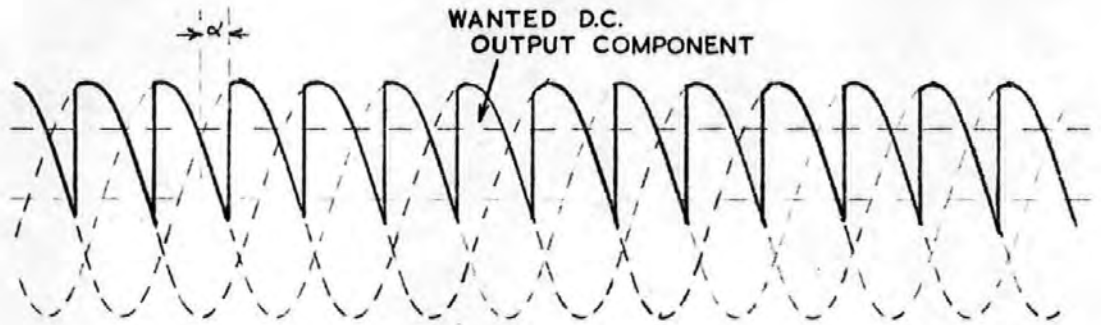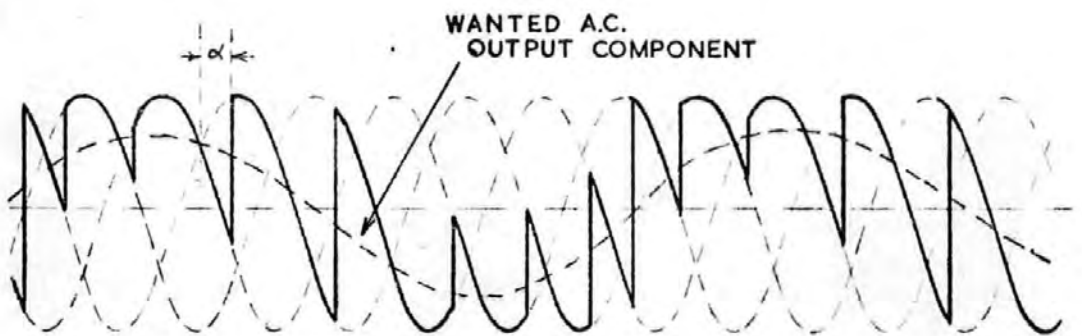
b) CYCLOCONVERTER OPERATION

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

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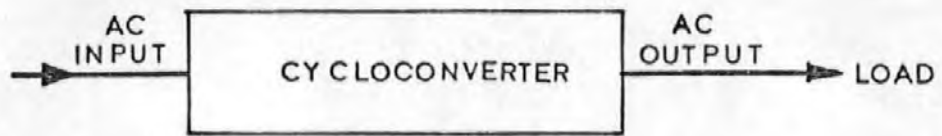
a) SIMPLEST DIAGRAM
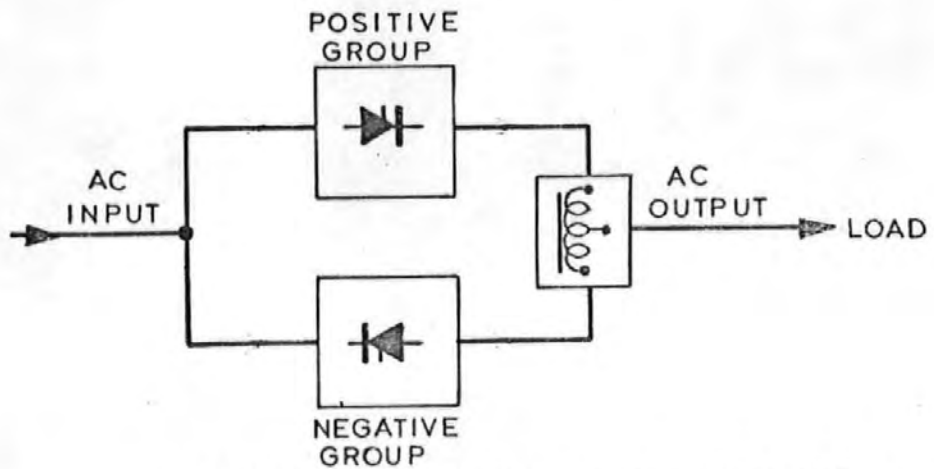
b) CIRCULATING-CURRENT CYCLOCONVERTER
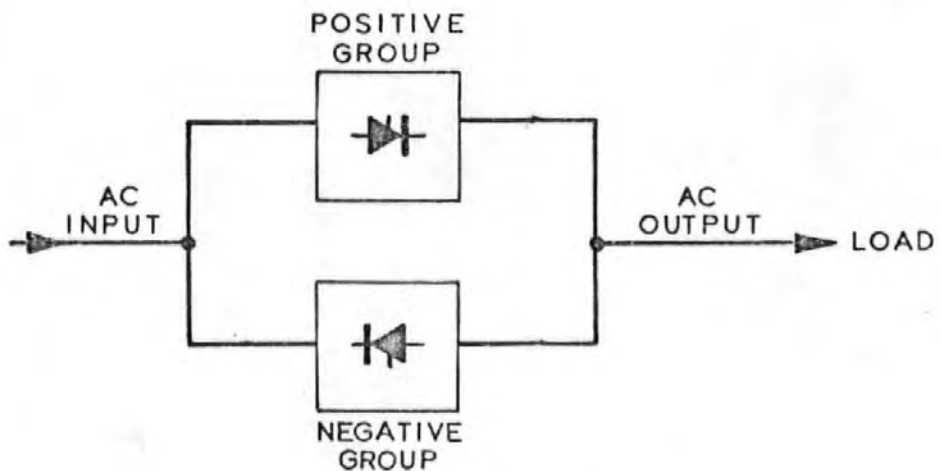
c) INHIBITED CYCLOCONVERTER

FIG 2·3   SCHEMATIC   DIAGRAMS OF
THE CYCLOCONVERTER

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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
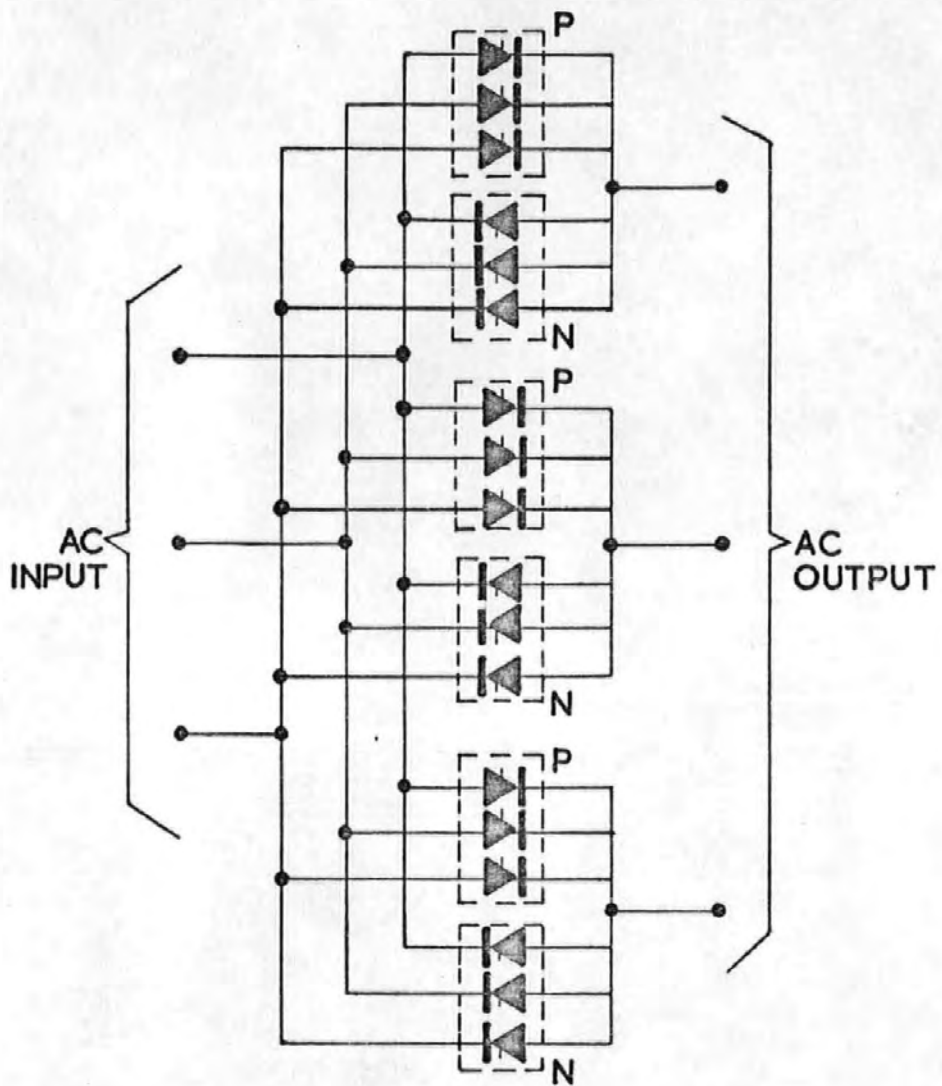
P = POSITIVE GROUP
N = NEGATIVE GROUP

FIG. 2·4   3-PULSE INHIBITED CYCLOCONVERTER
            WITH 3-PHASE OUTPUT

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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° and 180° 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° 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 biassed by the combined effect of the input voltage and the back-

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3—PHASE
SUPPLY
VOLTAGES

COSINE
TIMING WAVES
AND
SINUSOIDAL
REFERENCE
WAVE

COMMUTATION
POINTS

OUTPUT
VOLTAGE
WAVEFORM

WANTED
COMPONENT
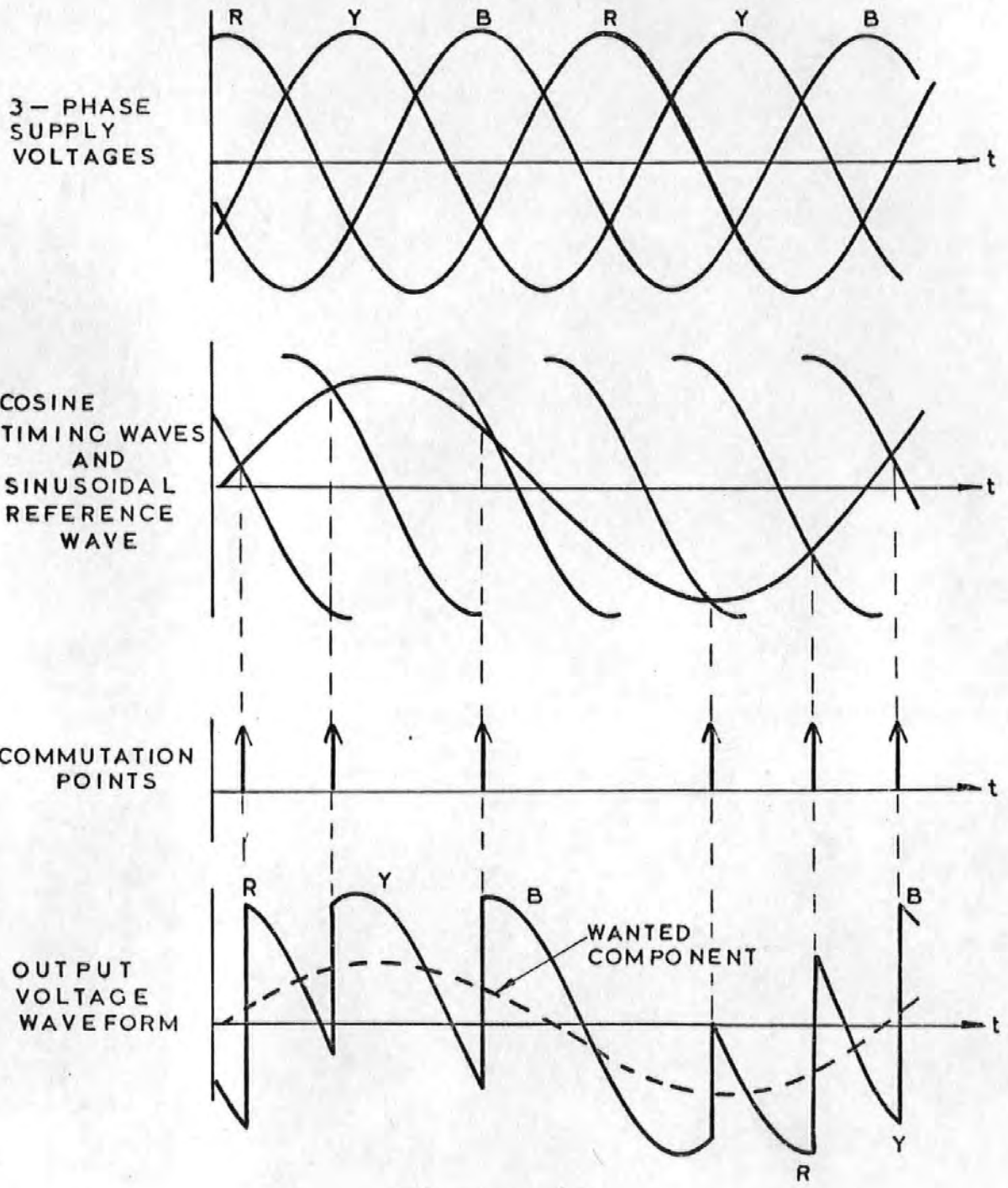
FIG 2·5 COSINUSOIDAL CONTROL OF A
THREE-PULSE CYCLOCONVERTER

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WANTED COMPONENT

a) POSITIVE GROUP

WANTED COMPONENT

b) NEGATIVE GROUP

WANTED COMPONENT
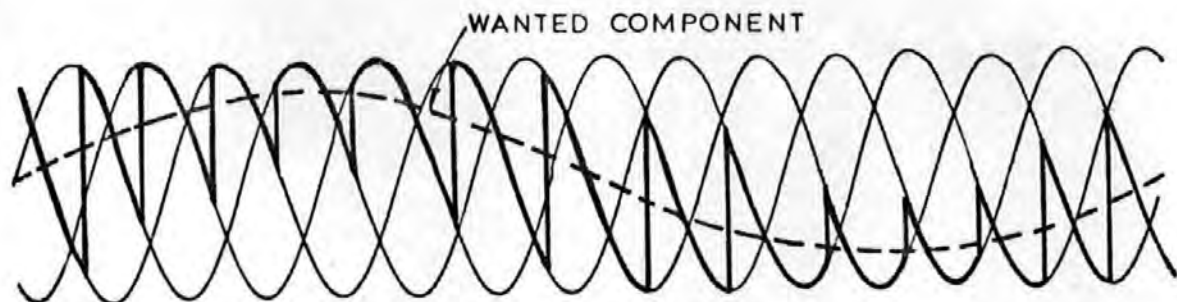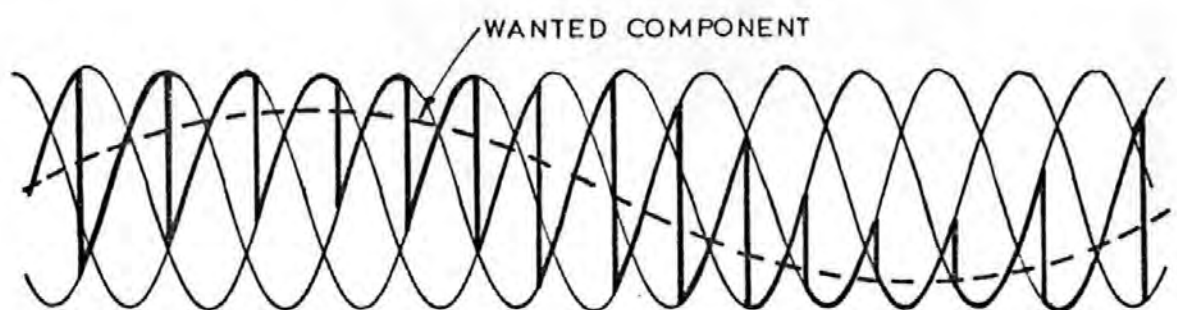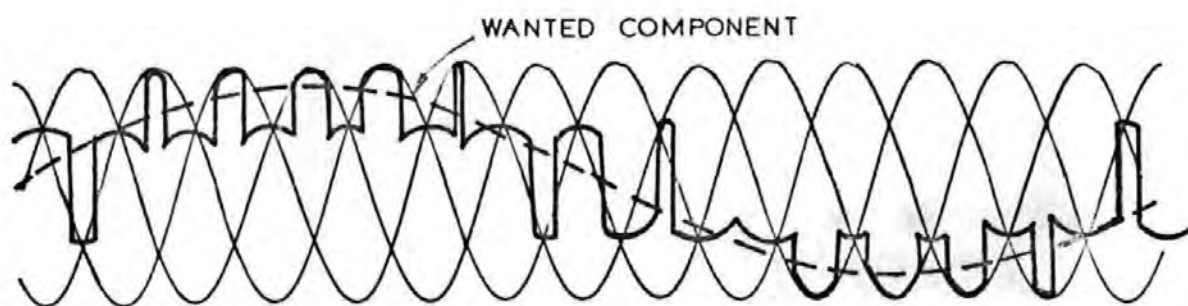
c) COMPLETE CYCLOCONVERTER

FIG 2·6 OUTPUT VOLTAGE WAVEFORMS FOR A
3-PULSE CIRCULATING- CURRENT
CYCLOCONVERTER.

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e.m.f. of the load.

The instantaneous differences between the
two voltage waveforms shown in figures 2.6a and 2.6b
cause instantaneous forward biassing 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
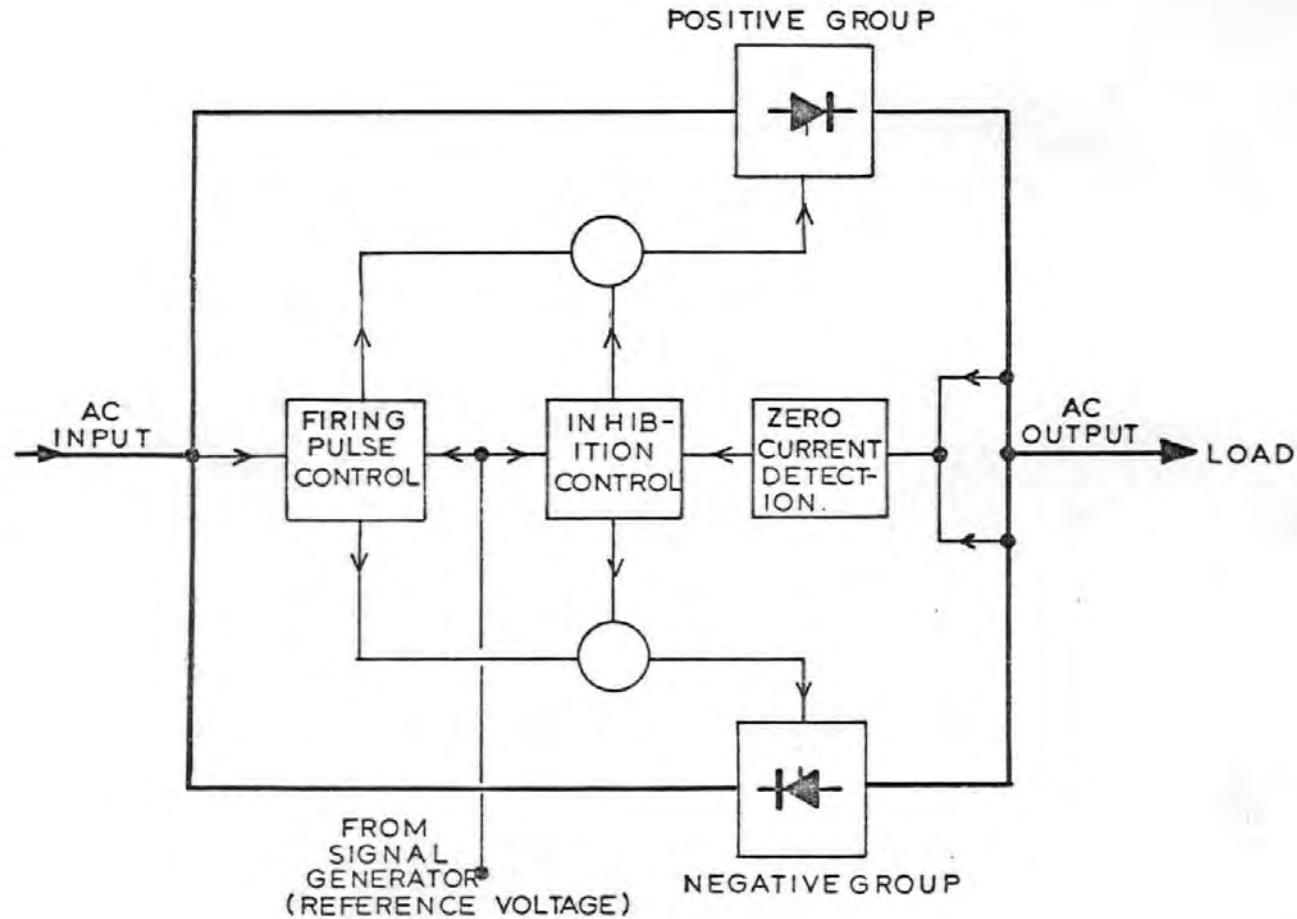
FIG. 2·7   SCHEMATIC DIAGRAM OF POWER AND CONTROL
CIRCUITS OF AN INHIBITED CYCLOCONVERTER

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

WANTED
COMPONENT

POSITIVE GROUP
CONDUCTION

NEGATIVE GROUP
CONDUCTION

CHANGEOVER AT
ZERO LOAD CURRENT

(a) UNITY DISPLACEMENT FACTOR LOAD

WANTED
COMPONENT

NEGATIVE
GROUP
CONDUCTION

POSITIVE GROUP
CONDUCTION

NEGATIVE GROUP
CONDUCTION

CHANGEOVER AT
ZERO LOAD CURRENT
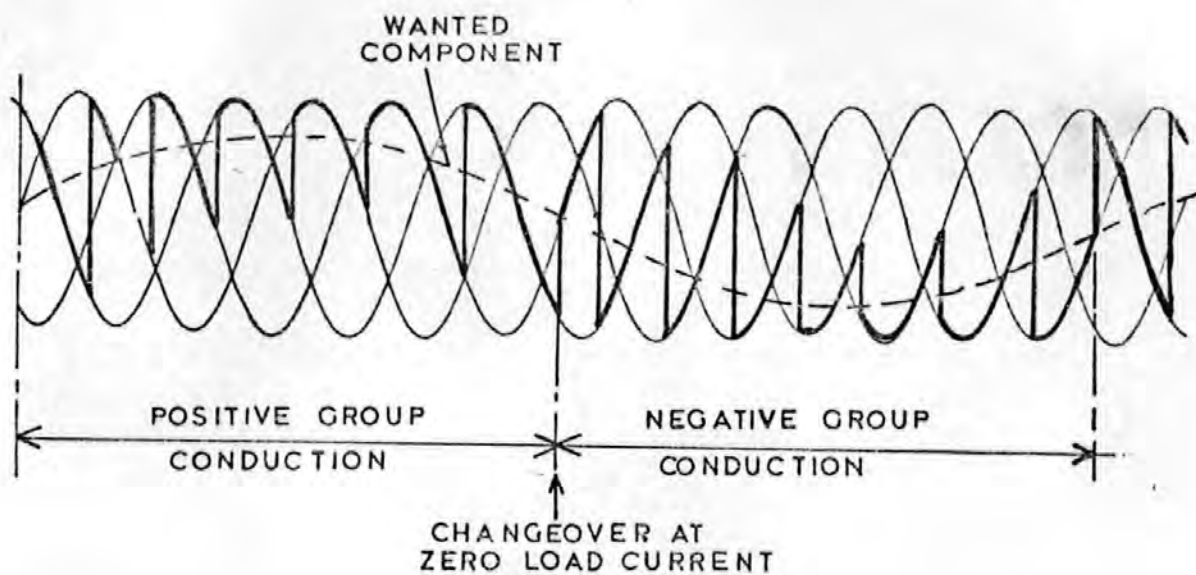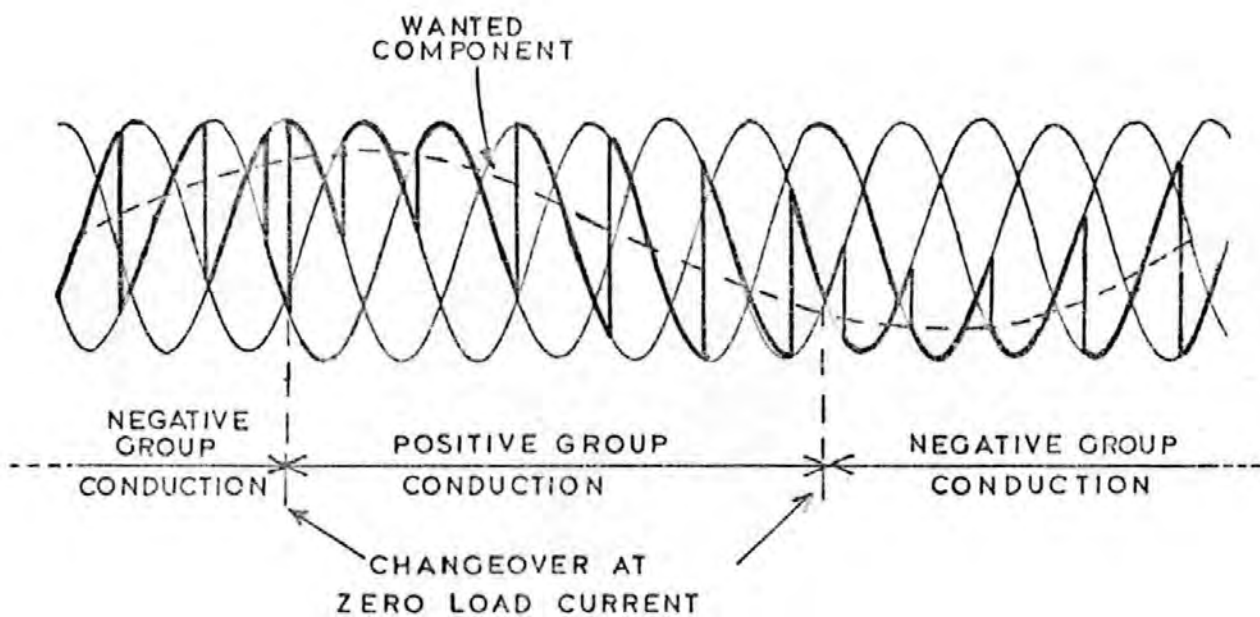
(b) LAGGING DISPLACEMENT FACTOR LOAD

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
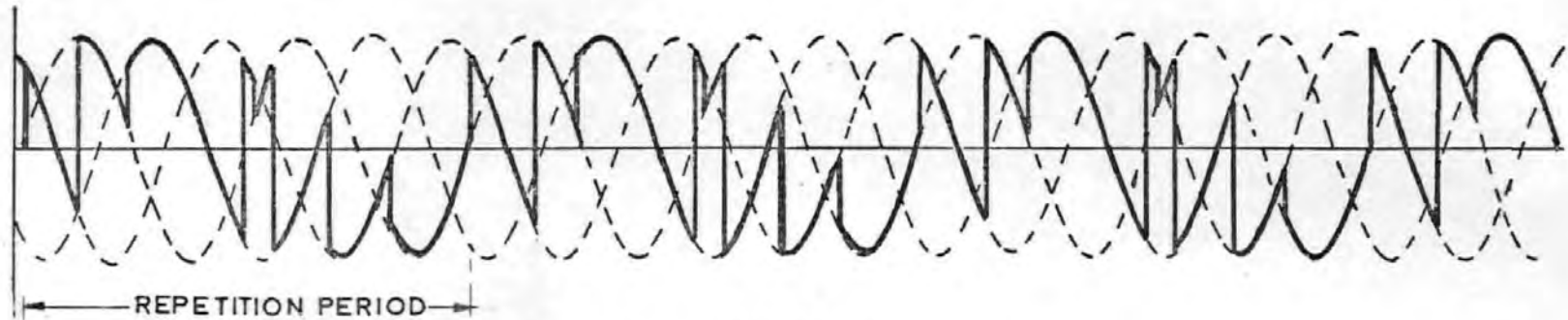
(a) OUTPUT FREQUENCY 25 Hz
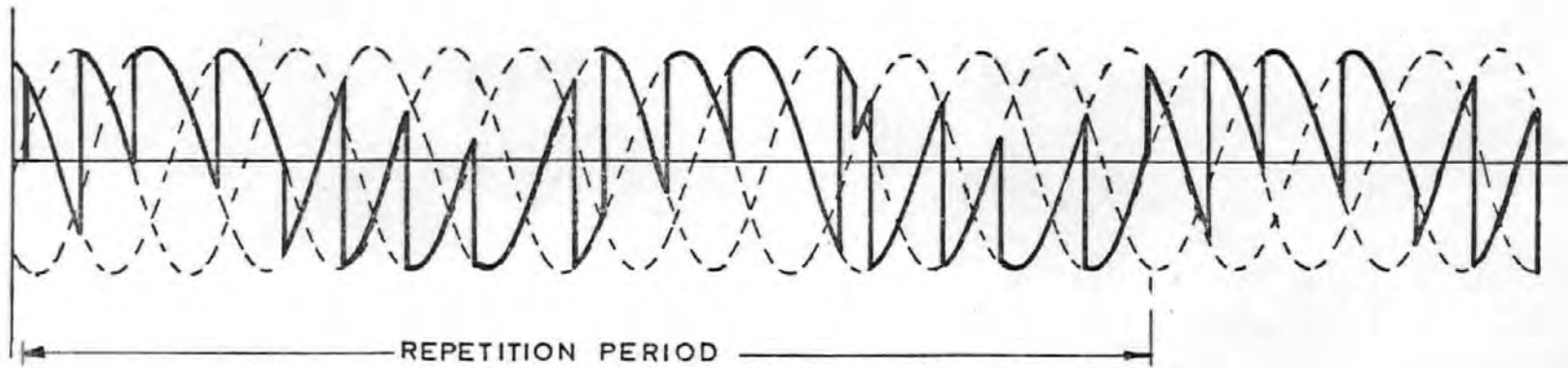
(b) OUTPUT FREQUENCY 20 Hz

FIG 2·9 OUTPUT VOLTAGE WAVEFORMS OF 50 Hz 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

$$\frac{pf_i}{f_f} \tag{2.1}$$

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.

$$\therefore \quad \frac{pf_i}{f_f} = \text{integer} \tag{2.2}$$

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_o) in each cycle of the fundamental
component. Therefore another necessary condition for
determining the fundamental frequency is that

$$\frac{f_o}{f_f} = \text{integer} \tag{2.3}$$

For a given set of values of p, f_i and f_o, the
highest value of f_f which satisfies the two conditions

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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:

$$\frac{pf_i}{f_f} \div \frac{f_o}{f_f} = \frac{pf_i}{f_o} \qquad (2.4)$$

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 pf_i/f_o (hereafter referred to as the 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 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 pf_i/f_o Ratio to a fractional number of the form:

$$M \frac{N}{D} \tag{2.5}$$

For example, if $p = 3$, $f_i = 50$ Hz and $f_o = 33\frac{1}{3}$ Hz, then the Ratio is 4.5. As an alternative, the pf_i/f_o Ratio is also 4.5 for $p = 6$, $f_i = 60$ Hz and $f_o = 80$ Hz. Thus in either case, the pf_i/f_o 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, $4\frac{5}{10}$ 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):

$$\frac{MD + N}{F_f} \div \frac{D}{F_f} \tag{2.6}$$

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:

$$\frac{MD + N}{1} \div \frac{D}{1} \tag{2.7}$$

Comparing expressions (2.4) and (2.7) gives:

$$f_o = D \text{ Hz} \tag{2.8a}$$

$$pf_i = MD + N \text{ Hz} \tag{2.8b}$$

$$f_f = 1 \text{ Hz} \tag{2.8c}$$

The significance of these expressions is that when the pf_i/f_o Ratio of the cycloconverter has the numerical value $M \frac{N}{D}$, 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:

$$f_f = \frac{f_o}{D} \qquad\qquad (2.9)$$

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 $p = 3$, $f_i = 50$ Hz and $f_o = 33\frac{1}{3}$ Hz, giving a pf_i/f_o Ratio of $4\frac{1}{2}$, the fundamental frequency is $\frac{33\frac{1}{3}}{2} = 16\frac{2}{3}$ Hz. In the example where $p = 6$, $f_i = 60$ Hz and $f_o = 80$ 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 α, 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 α; 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 cosinusoidal 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

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(a)
RECTIFIER
ACTION

SWITCHING
FUNCTIONS
R—PHASE

Y- PHASE

B- PHASE

(b)
CYCLO-
CONVERTER
ACTION

(CIRCULAT
ING
CURRENT
MODE)

SWITCHING
FUNCTIONS
R — PHASE

Y - PHASE

B - PHASE
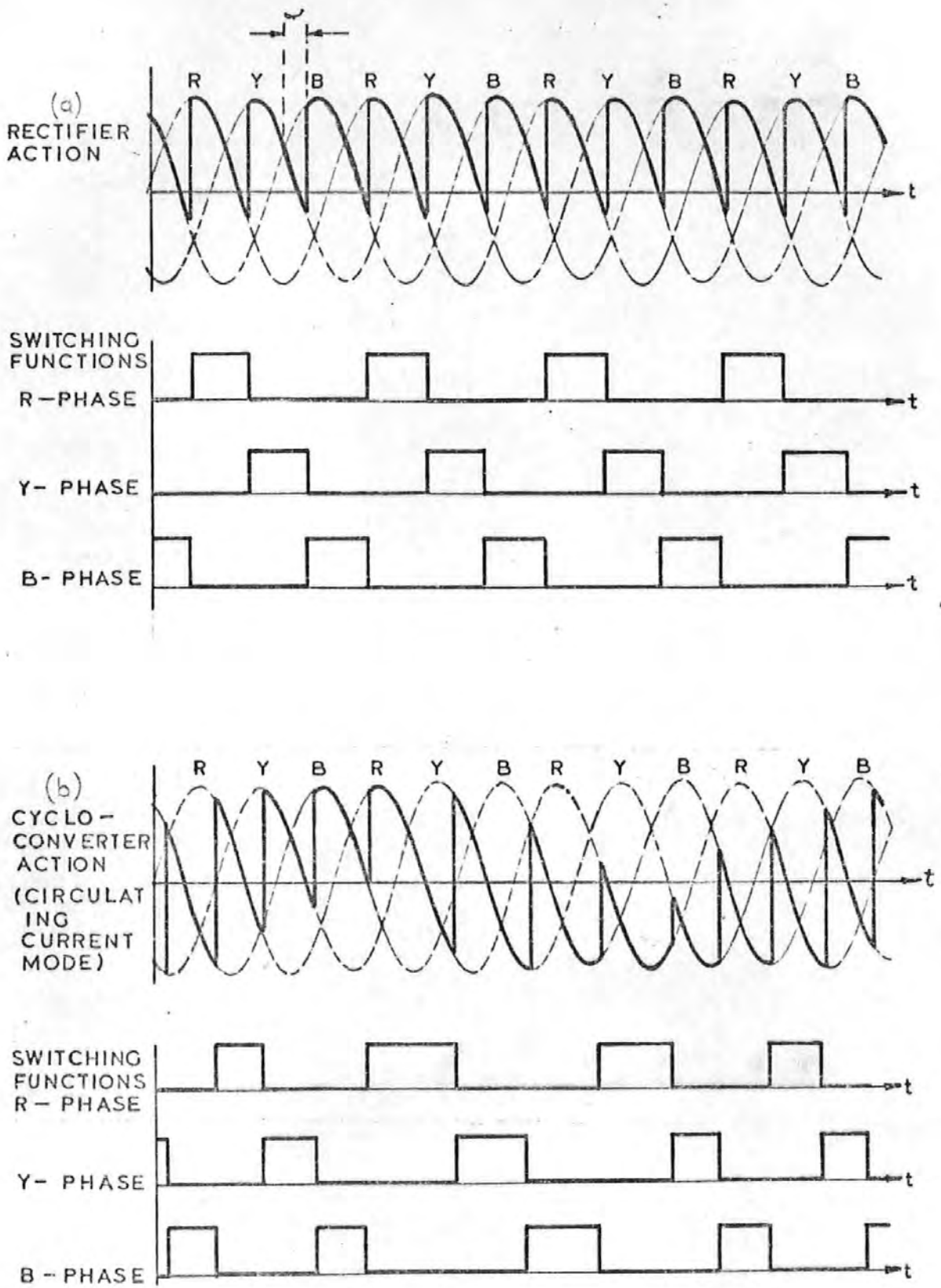
FIG 3·1 SWITCHING FUNCTIONS AND MODIFIED
SWITCHING FUNCTIONS

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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

$$(r \sin w_o t) = \cos (w_i t - \theta_k) \qquad (3.1)$$

where θ_k is the phase displacement of the kth cosine
timing wave from an arbitrary reference point.

The thyristor switching functions, $f(w_i t, w_o t)$,
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_i and the output
frequency w_o. 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
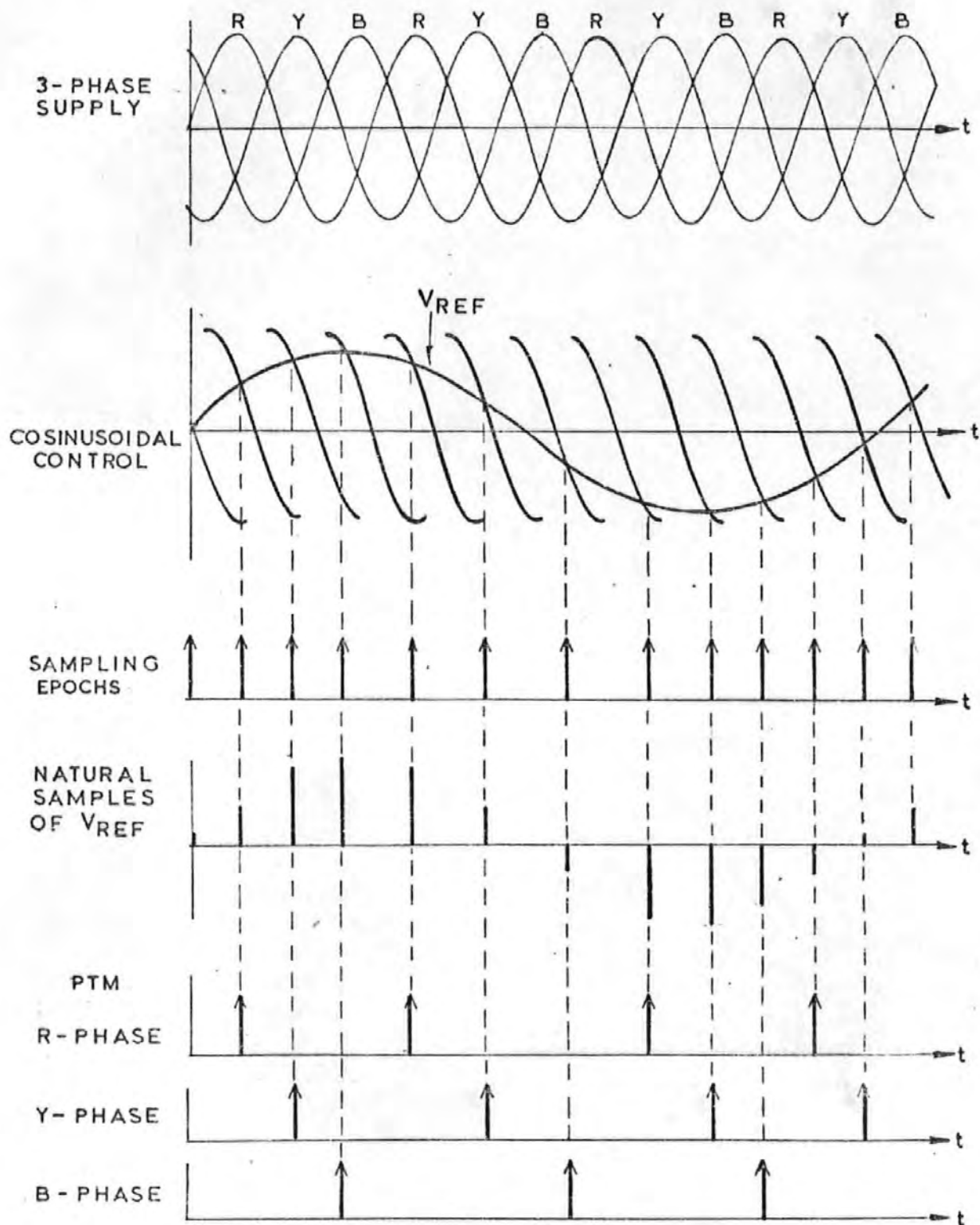
FIG 3·2 COSINUSOIDAL CONTROL WITH NATURAL
SAMPLING OF THE REFERENCE WAVE:
PULSE TIME MODULATED IMPULSES

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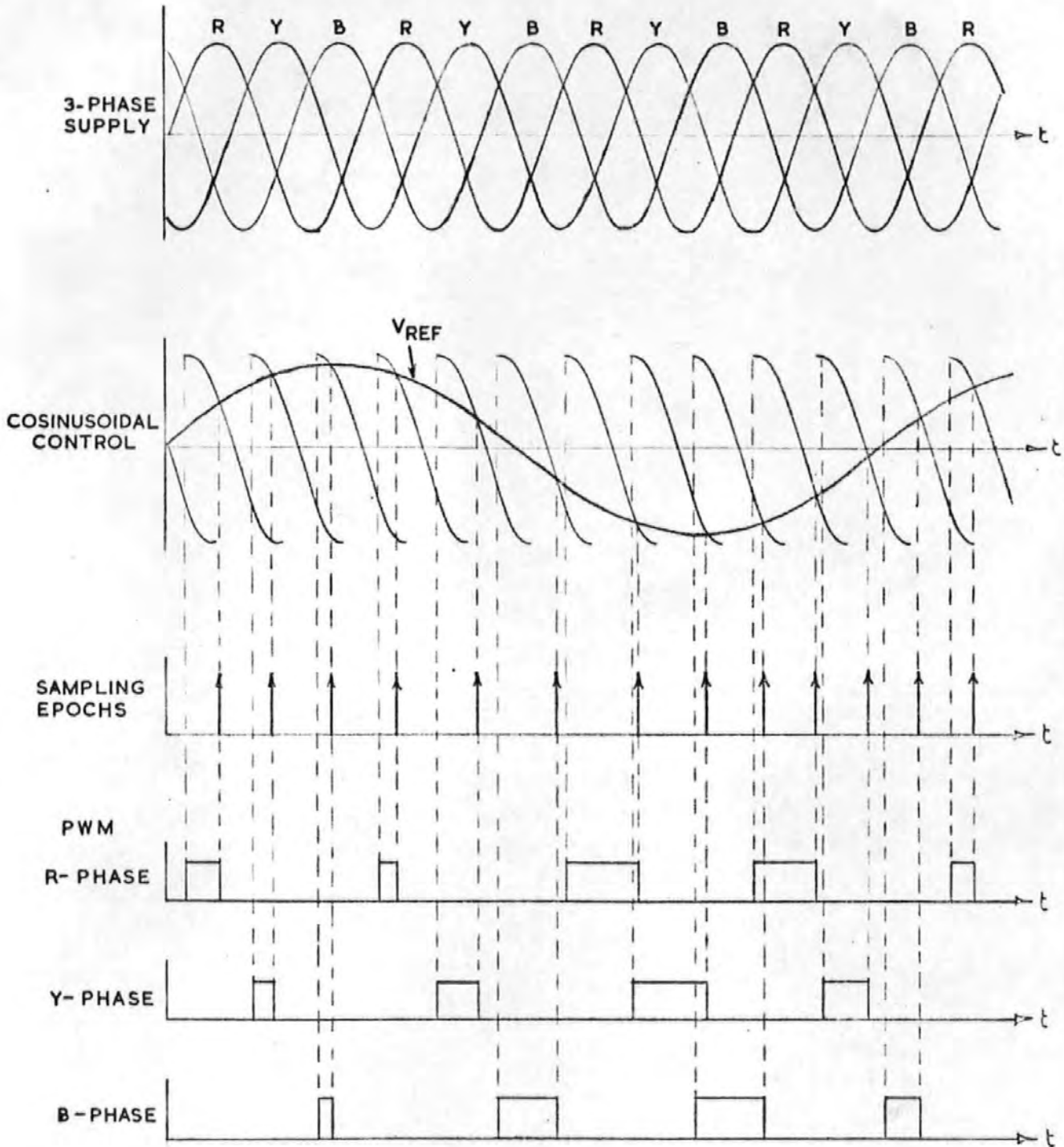
FIG 3·3 COSINUSOIDAL CONTROL WITH NATURAL
SAMPLING OF THE REFERENCE WAVE :
PULSE WIDTH MODULATED PULSES
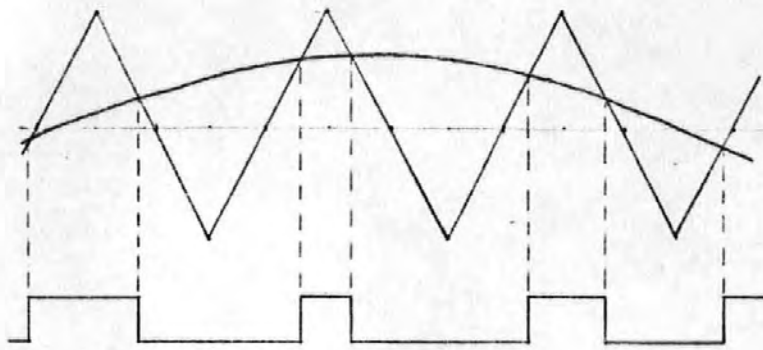
a) LINEAR MODULATION
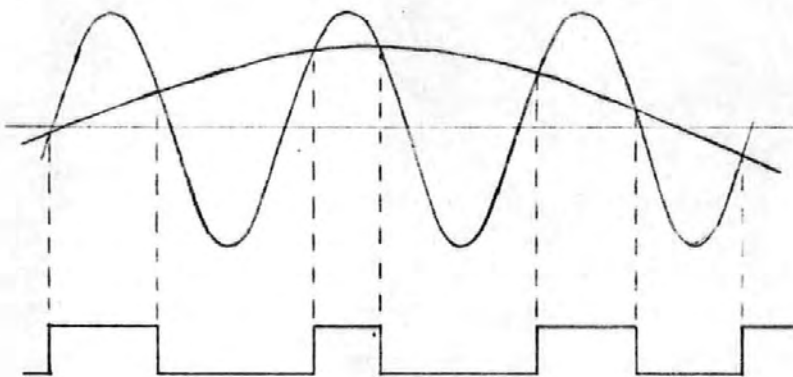
b) NON-LINEAR MODULATION (ONE COSINE WAVE)
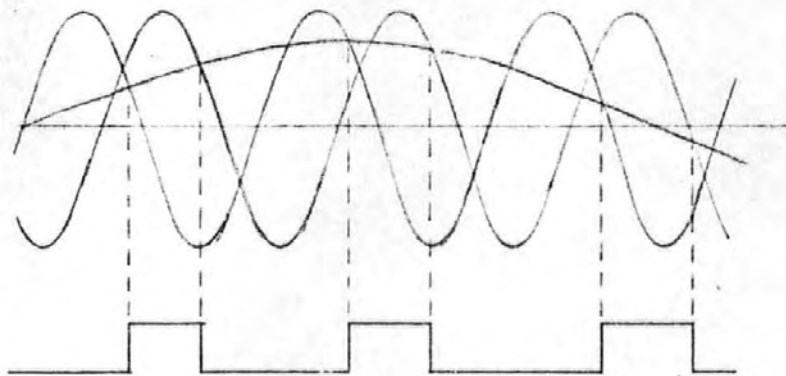
c) NON-LINEAR MODULATION (TWO COSINE WAVES)

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 2π etc., as shown in figure 3.4b. However, without recourse to forced commutation techniques, it is not possible to employ a thyristor firing angle (α) in the range $180° \leqslant \alpha \leqslant 360°$ 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.

(3.4) <u>Application of Regular Sampling to the Control</u>
       <u>of the Circulating-current Cycloconverter</u>

(3.4.1) <u>General</u>

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 Panter[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
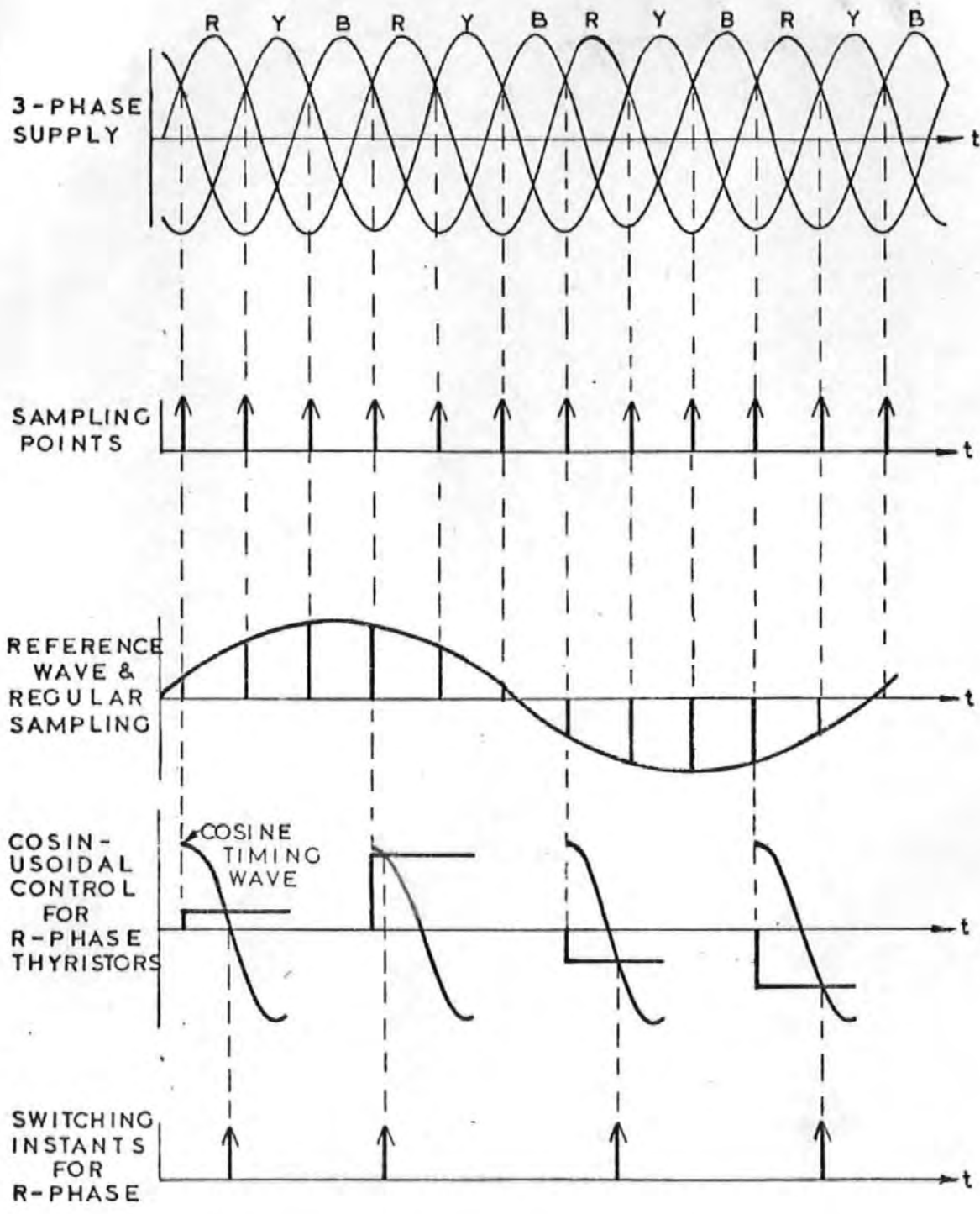
FIG 3·5 COSINUSOIDAL CONTROL WITH
REGULAR SAMPLING OF THE
REFERENCE WAVE

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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_o.

$$\therefore \text{ Nyquist sampling rate} = 2f_o \qquad (3.2)$$

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_i).

$$\therefore \quad \text{Actual sampling rate} = pf_i \qquad (3.3)$$

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

$$pf_i \geqslant 2f_o$$

$$\text{or} \quad f_o \leqslant \frac{pf_i}{2} \qquad (3.4)$$

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

$$\frac{3 \cdot 50}{2} \quad = \quad 75 \text{ Hz}$$

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

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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 pf_i/f_o Ratio of the cycloconverter is 2.0. It was shown in section 2.6.4 that the magnitude of the pf_i/f_o Ratio is significant in terms of the fundamental repetition frequency of the output voltage waveform. For example, when the value of the pf_i/f_o 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

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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) <u>Spectral Analysis by Computer Investigation</u>

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

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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.

OUTPUT FREQ. (HZ)	FREQUENCY OF COMPONENT (HZ)														
	0	5	10	15	20	25	30	35	40	45	50	55	60	65	70
5		100													
10			100												
15				100											
20					100										9
25						100					9				
30							100								
35			9					100				4			
40			9		4				100		2				31
45				5			10			100			31		
50											100				

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[10], and Ford[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 pf_i/f_o 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 pf_i/f_o Ratio. Output frequencies of 35, 40 and 45 Hz correspond to non-integer values of the pf_i/f_o 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 $\frac{3\sqrt{3}}{2\pi}$ (= 0.825) of the input phase voltage[10]. The modulation factor for the computed results is 0.9 and therefore the theoretical output voltage is 0.9 × 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

OUTPUT FREQ. (HZ)	FREQUENCY OF COMPONENT (HZ)														
	0	5	10	15	20	25	30	35	40	45	50	55	60	65	70
5		100													
10			100				2								
15				100						3					
20					100								6		1
25						100									
30							100								
35								100							
40									100						31
45										100			29		
50											100				

TABLE 3.2 3-PULSE 50HZ CIRCULATING-CURRENT CYCLOCONVERTER:
MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS
AS PERCENTAGES OF THE WANTED COMPONENT.
REGULAR SAMPLING
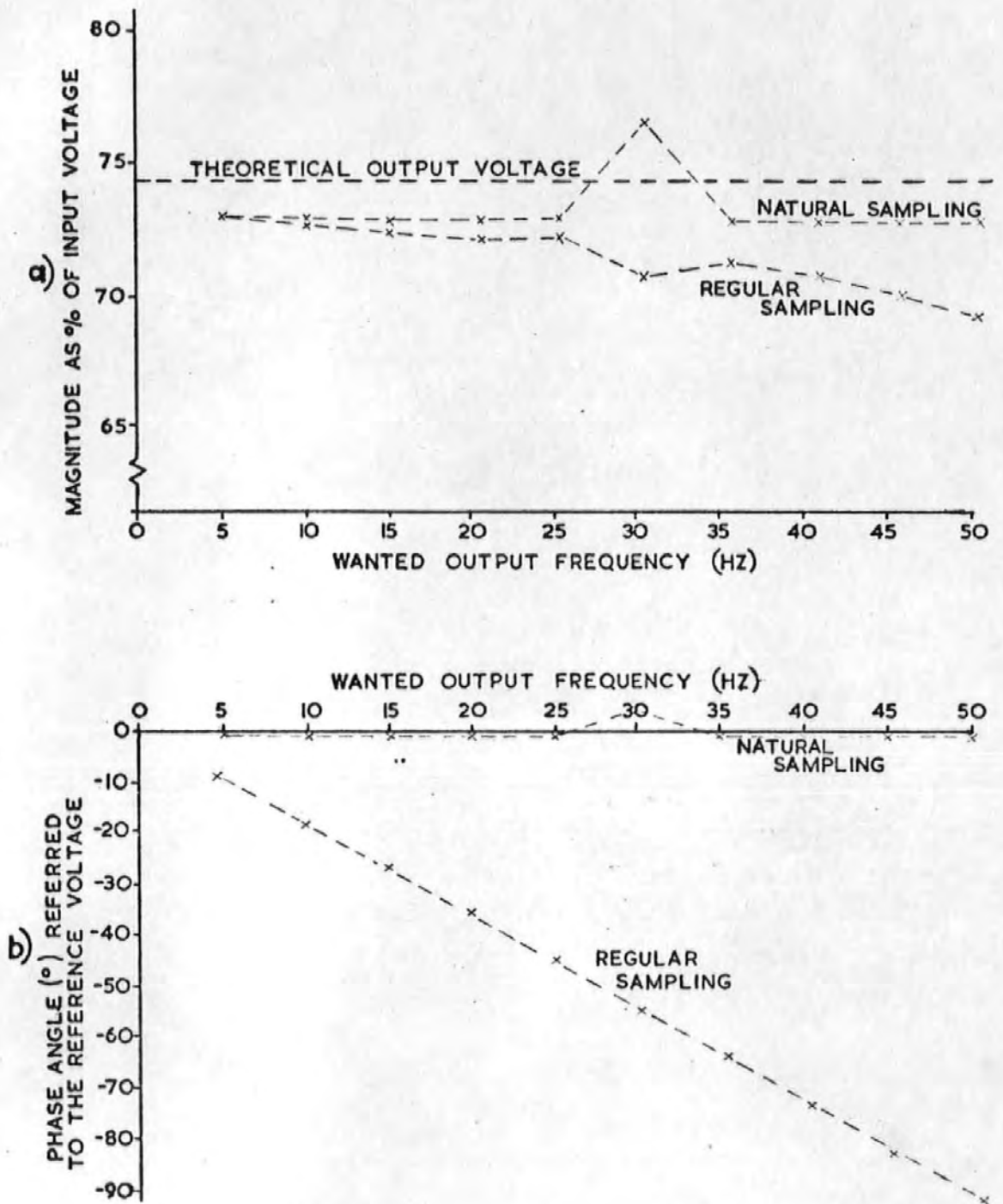
FIG 3·6 MAGNITUDE AND PHASE ANGLE OF THE WANTED
COMPONENT (CIRCULATING-CURRENT CYCLOCONVERTER)
NATURAL & REGULAR SAMPLING

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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 $+7^O$. With regular sampling, however, a phase-shift is introduced for all wanted output frequencies, increasing linearly from 0° at 0 Hz to -90° 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°, 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

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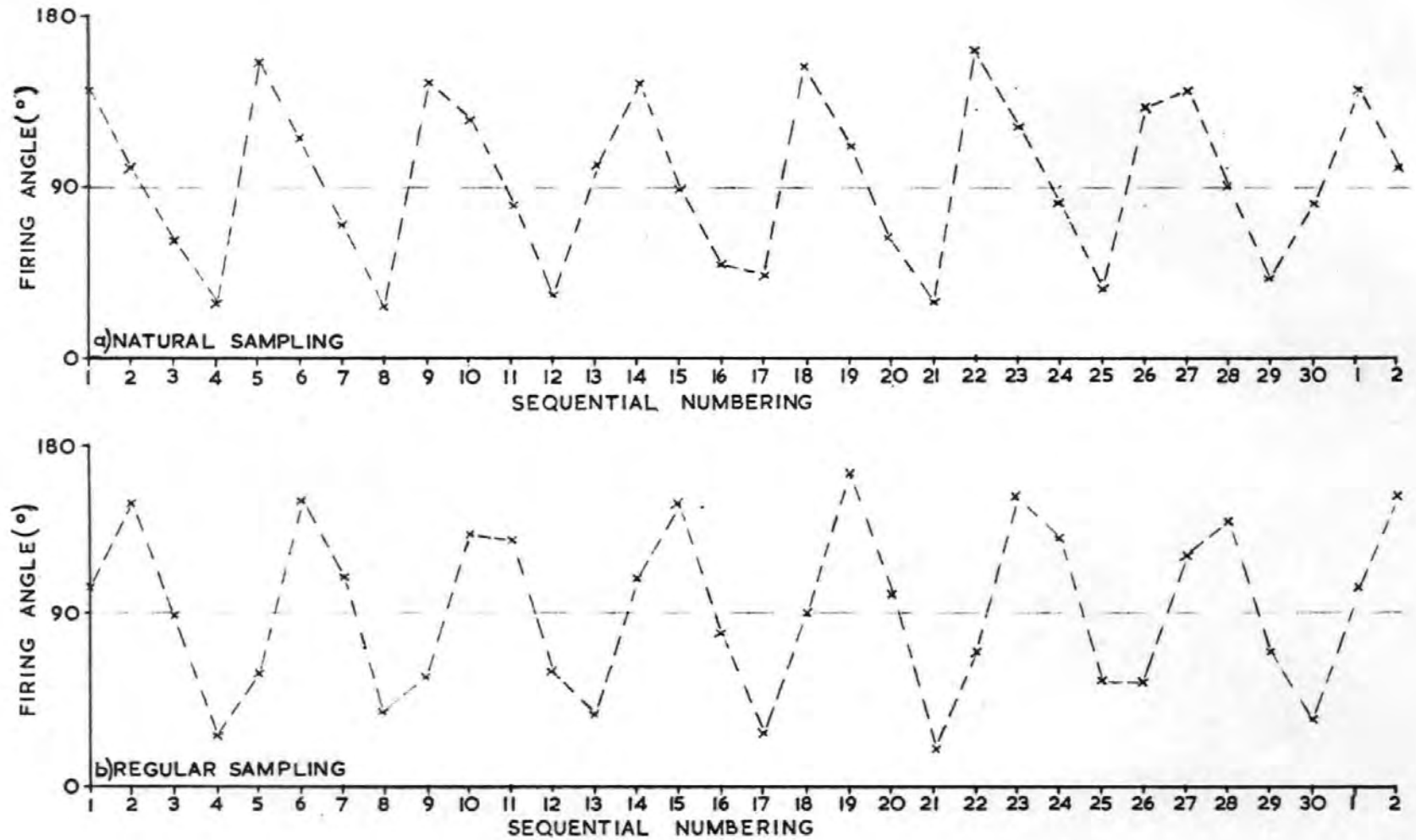
FIG 3.7 FIRING ANGLES FOR ONE REPETITION CYCLE OF WANTED OUTPUT COMPONENT
OF THE CIRCULATING-CURRENT CYCLOCONVERTER (POSITIVE GROUP)
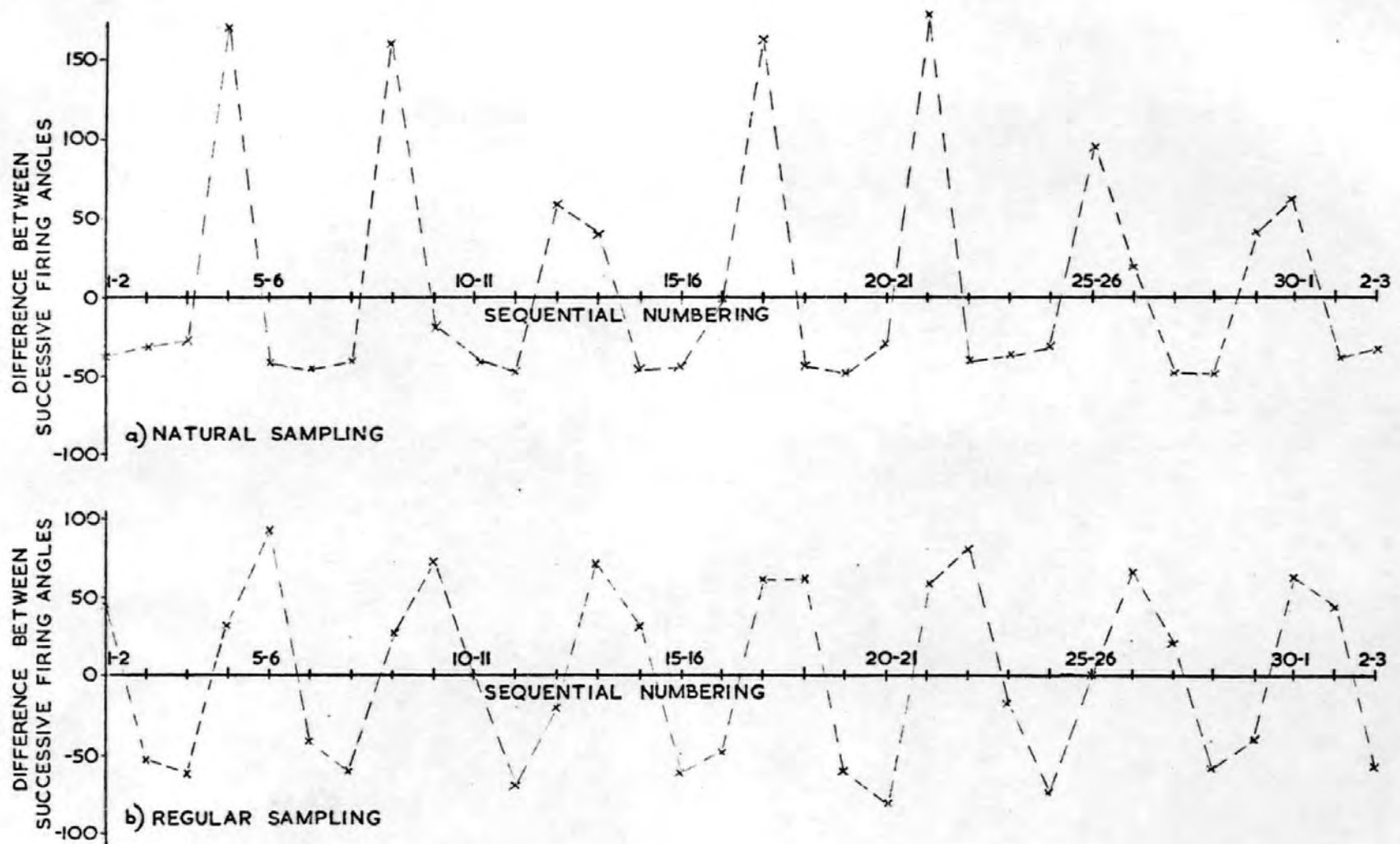
FIG 3·8 DIFFERENCES BETWEEN SUCCESSIVE FIRING ANGLES IN FIGURES 3·7a & 3·7b

58

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 pf_i/f_o Ratio value. In section 3.4.2 it was shown that if the pf_i/f_o 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 pf_i/f_o 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 pf_i/f_o 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 $0 - 180^{\circ}$) 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 biassed 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.

a) CONSTANT +VE BIAS CAUSING A D.C. COMPONENT
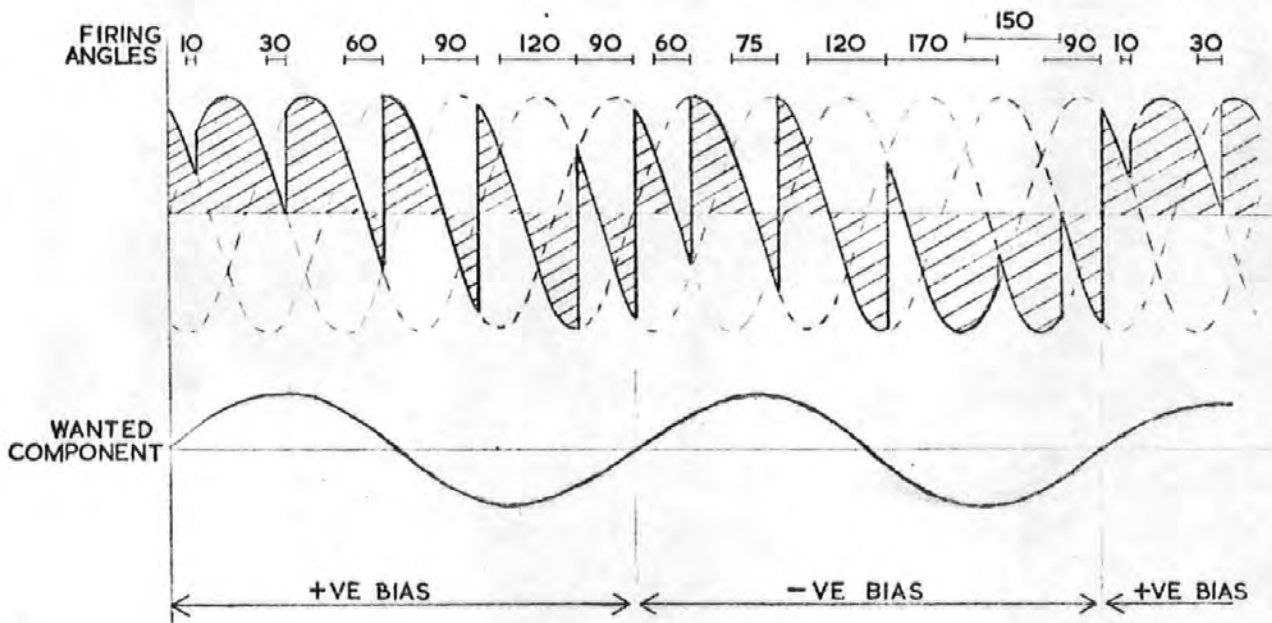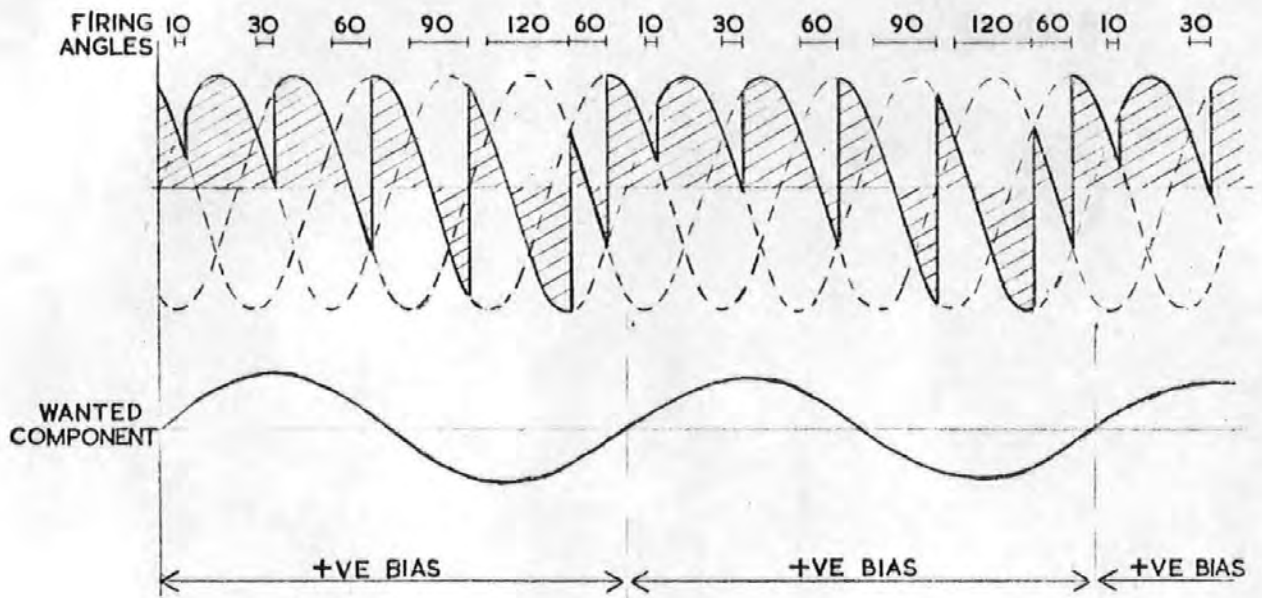
b) ALTERNATING +VE AND -VE BIAS CAUSING A SUB-HARMONIC COMPONENT

FIG 3·9 EFFECT OF BIASSED FIRING ANGLES
ON THE OUTPUT VOLTAGE WAVEFORM

(3.5.2) The Equal-area Criterion

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

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

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

$$\text{d.c. component} = \frac{1}{T_f} \int_0^{T_f} f(t)dt \qquad (3.6)$$

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

$$\frac{1}{T_o} \int_0^{T_o} f(t)dt \qquad (3.7)$$

Expression (3.7) gives the d.c. component of the cycloconverter output voltage only if the wanted output frequency of the cycloconverter is

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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_o. 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 $\int_0^{T_o} f(t)dt$
since the component $\frac{1}{T_o}$ of expression (3.7) is common.
This integral gives the nett area enclosed by the output
voltage waveform during the period T_o 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

OUTPUT
VOLTAGE
WAVEFORM

WANTED
COMPONENT

NETT
AREAS

65
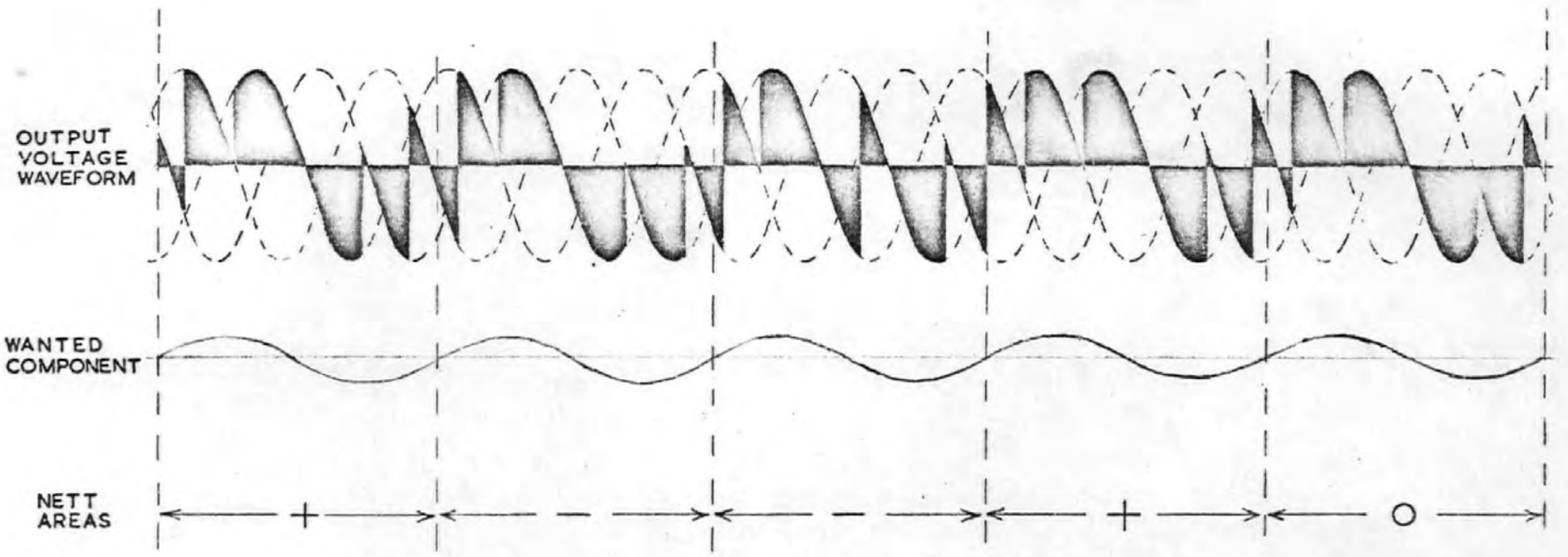
FIG 3·10 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF CIRCULATING-
CURRENT CYCLOCONVERTER (POSITIVE GROUP).
NATURAL SAMPLING

OUTPUT
VOLTAGE
WAVEFORM

WANTED
COMPONENT
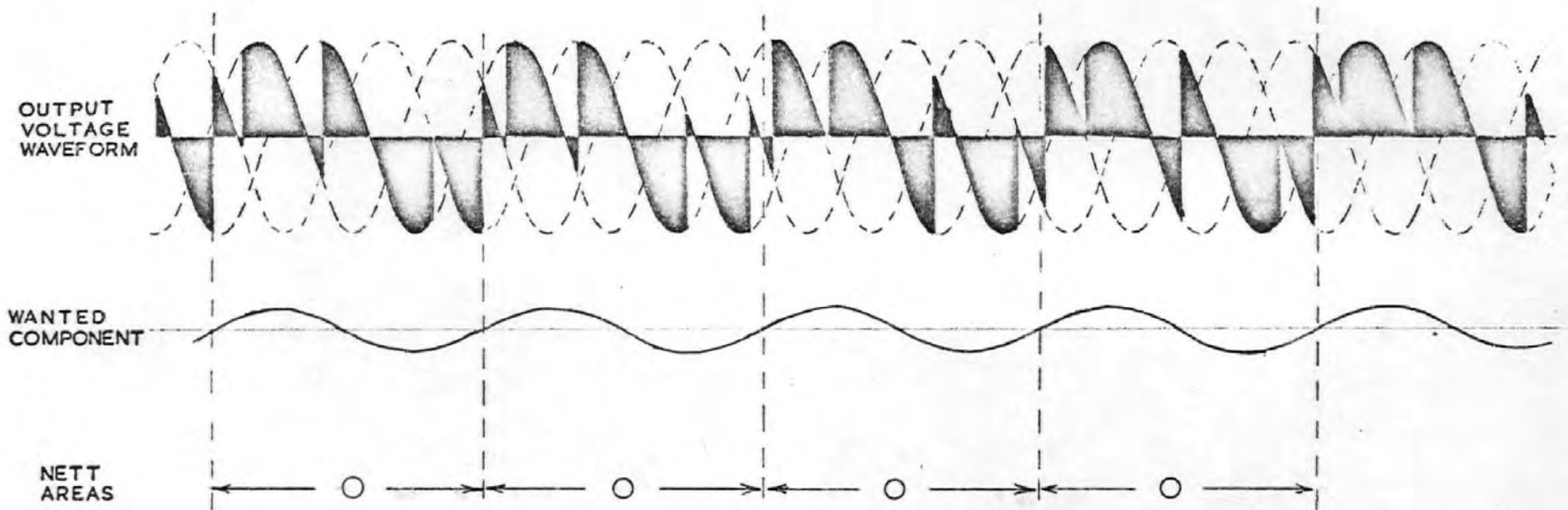
NETT
AREAS

FIG 3·II 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_o) of the cycloconverter. This leads to the
concept of inhibition switching functions of unity

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magnitude as shown in figure 3.12. If the inhibition switching function for one group is $f(w_o t)$, then the inhibition switching function for the other group is $f(w_o t - \pi)$. 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
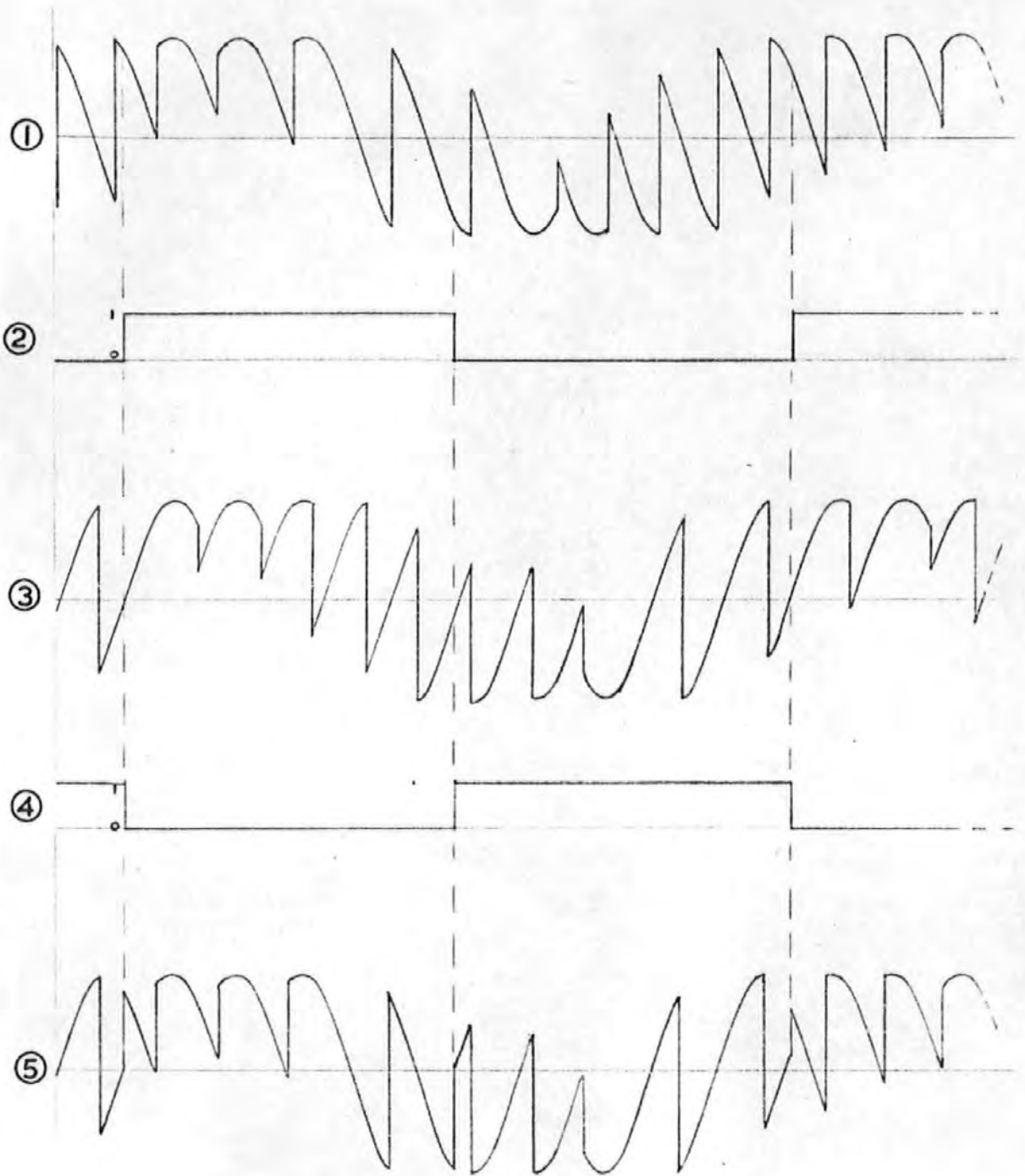
① POSITIVE GROUP OUTPUT VOLTAGE WAVEFORM
② POSITIVE GROUP INHIBITION SWITCHING FUNCTION
③ NEGATIVE GROUP OUTPUT VOLTAGE WAVEFORM
④ NEGATIVE GROUP INHIBITION SWITCHING FUNCTION
⑤ OUTPUT VOLTAGE WAVEFORM OF INHIBITED CYCLOCONVERTER
$$\left(= ① \times ② + ③ \times ④ \right)$$

FIG 3·12   INHIBITION SWITCHING FUNCTIONS OF THE OUTPUT
VOLTAGE WAVEFORM OF AN INHIBITED CYCLOCONVERTER

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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OUTPUT FREQ. (HZ)	FREQUENCY OF COMPONENT (HZ)														
	0	5	10	15	20	25	30	35	40	45	50	55	60	65	70
5		100		1		1									
10			100				2				2				2
15				100			1			3			4		
20			3		100		6		4		2		1		16
25	3					100					14				
30							100								
35			15	4			2	100	1			9	5		
40			16		10		3		100		5		6		32
45				7			14			100			32		
50											100				

TABLE 3·3 3-PULSE 50HZ INHIBITED CYCLOCONVERTER :
MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS
AS PERCENTAGES OF THE WANTED COMPONENT,
NATURAL SAMPLING

OUTPUT FREQ. (HZ)	FREQUENCY OF COMPONENT (HZ)														
	0	5	10	15	20	25	30	35	40	45	50	55	60	65	70
5		100		3		1									
10			100				8				4				4
15				100			4			10			5		
20			3		100		5		2		2		11		12
25	2					100					14				
30							100								
35			12	2			1	100	1			3	7		
40			13		4		1		100		2		3		41
45				3			12			100			41		
50											100				

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 12½%. 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
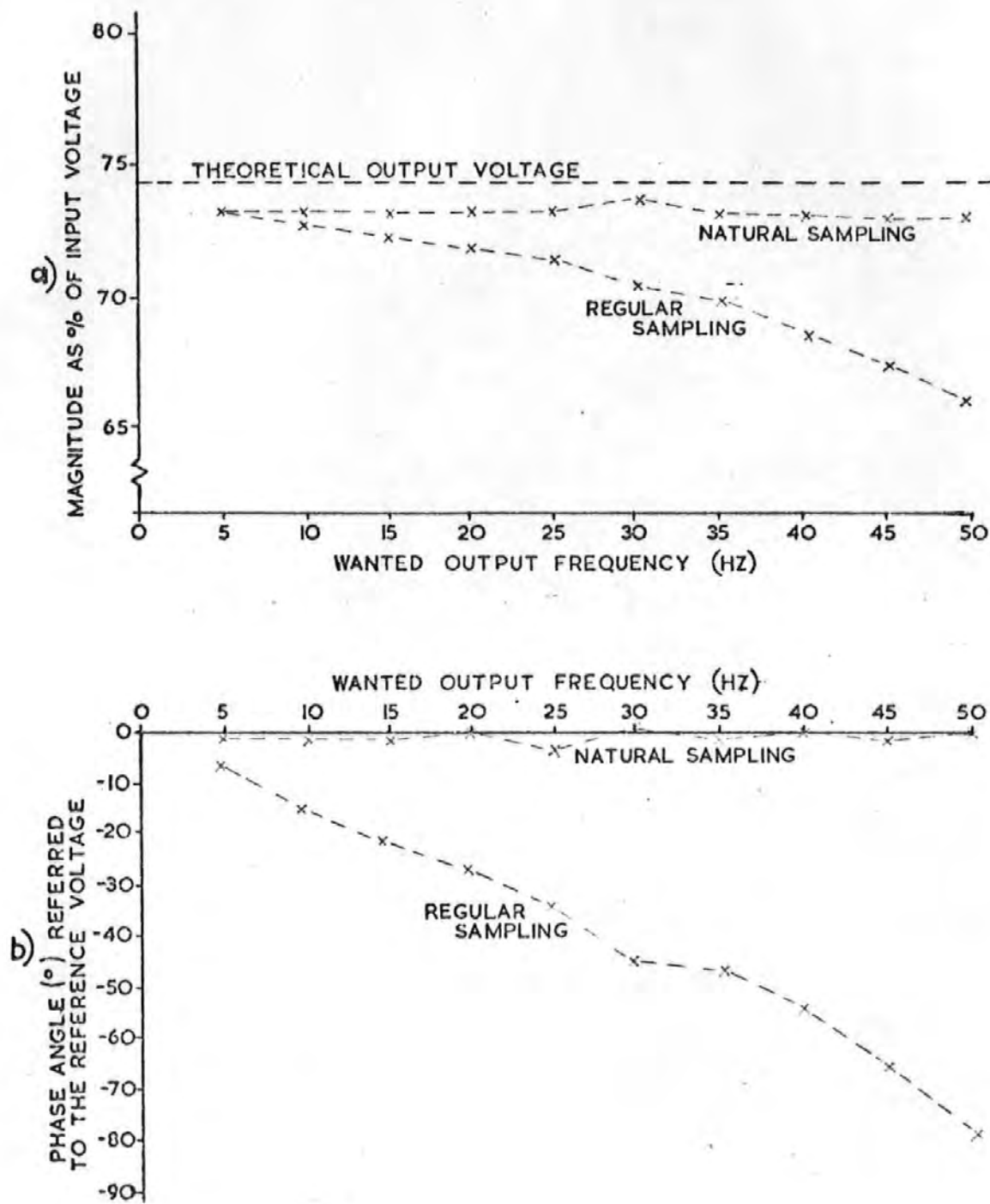
FIG 3·13 MAGNITUDE AND PHASE ANGLE OF THE WANTED
COMPONENT (INHIBITED CYCLOCONVERTER).
NATURAL & REGULAR SAMPLING

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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

OUTPUT
VOLTAGE
WAVEFORM

WANTED
COMPONENT

NETT
AREAS

78
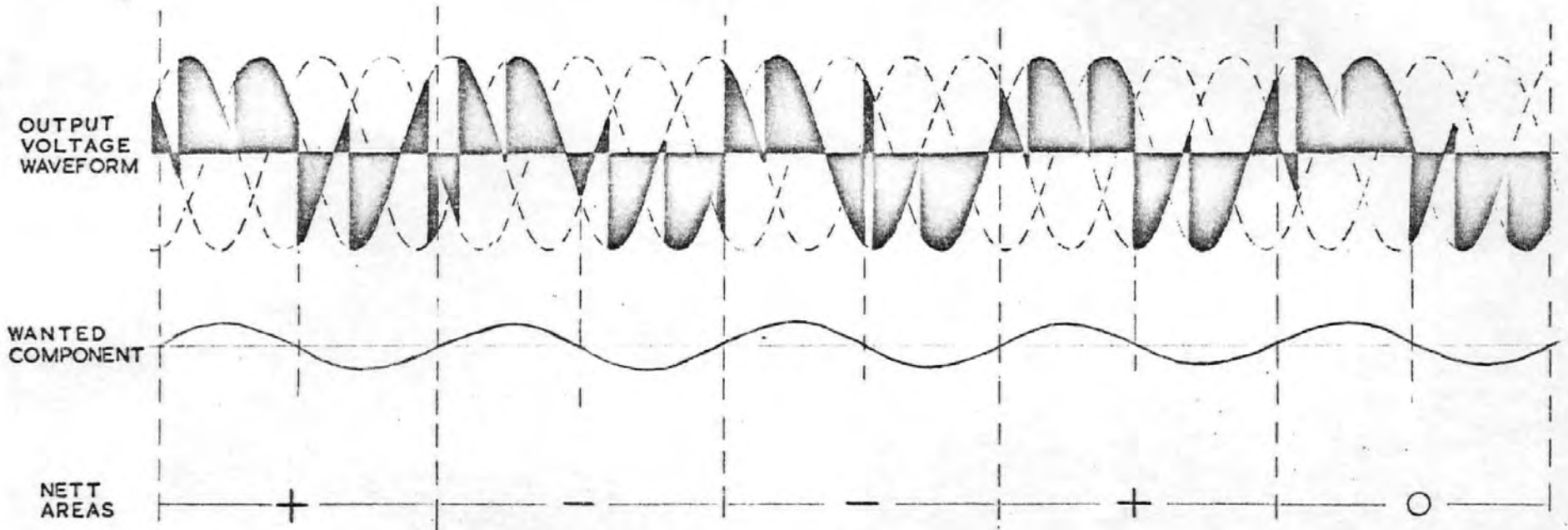
FIG 3·14 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF INHIBITED CYCLOCONVERTER.
UDF LOAD, SINUSOIDAL CURRENT ; NATURAL SAMPLING

OUTPUT
VOLTAGE
WAVEFORM

WANTED
COMPONENT
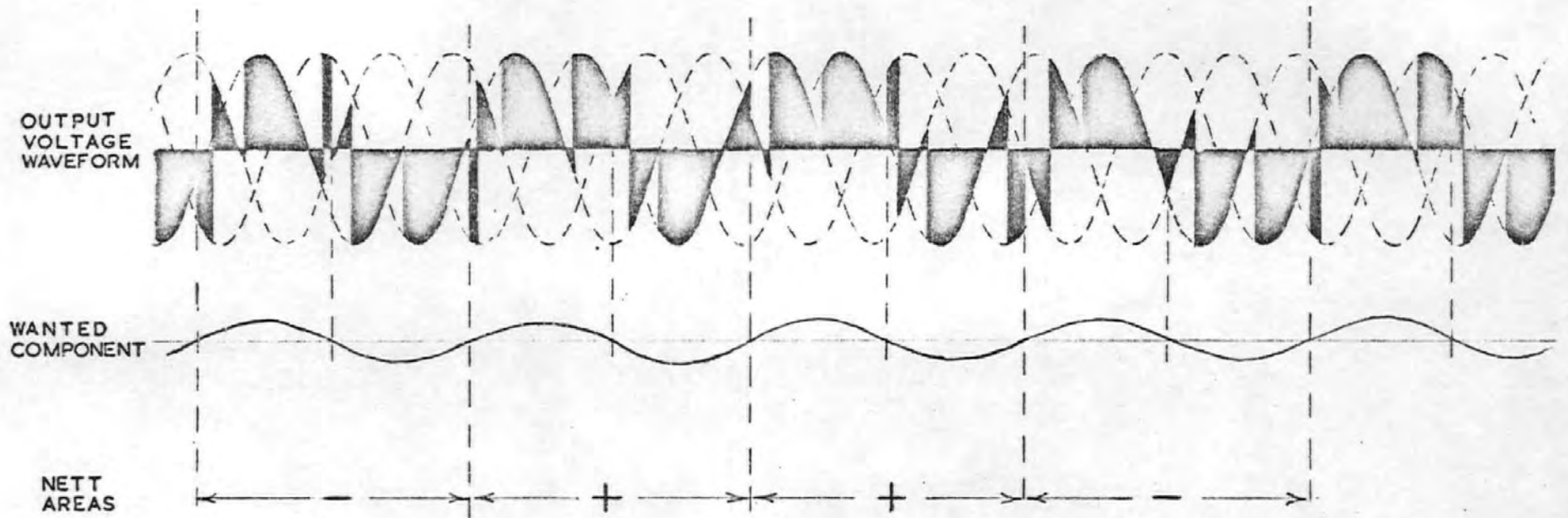
NETT
AREAS

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 $12\frac{1}{2}\%$ with regular sampling
(table 3.4). The 20 Hz sub-harmonic component for
40 Hz output frequency, and the 15 Hz sub-harmonic
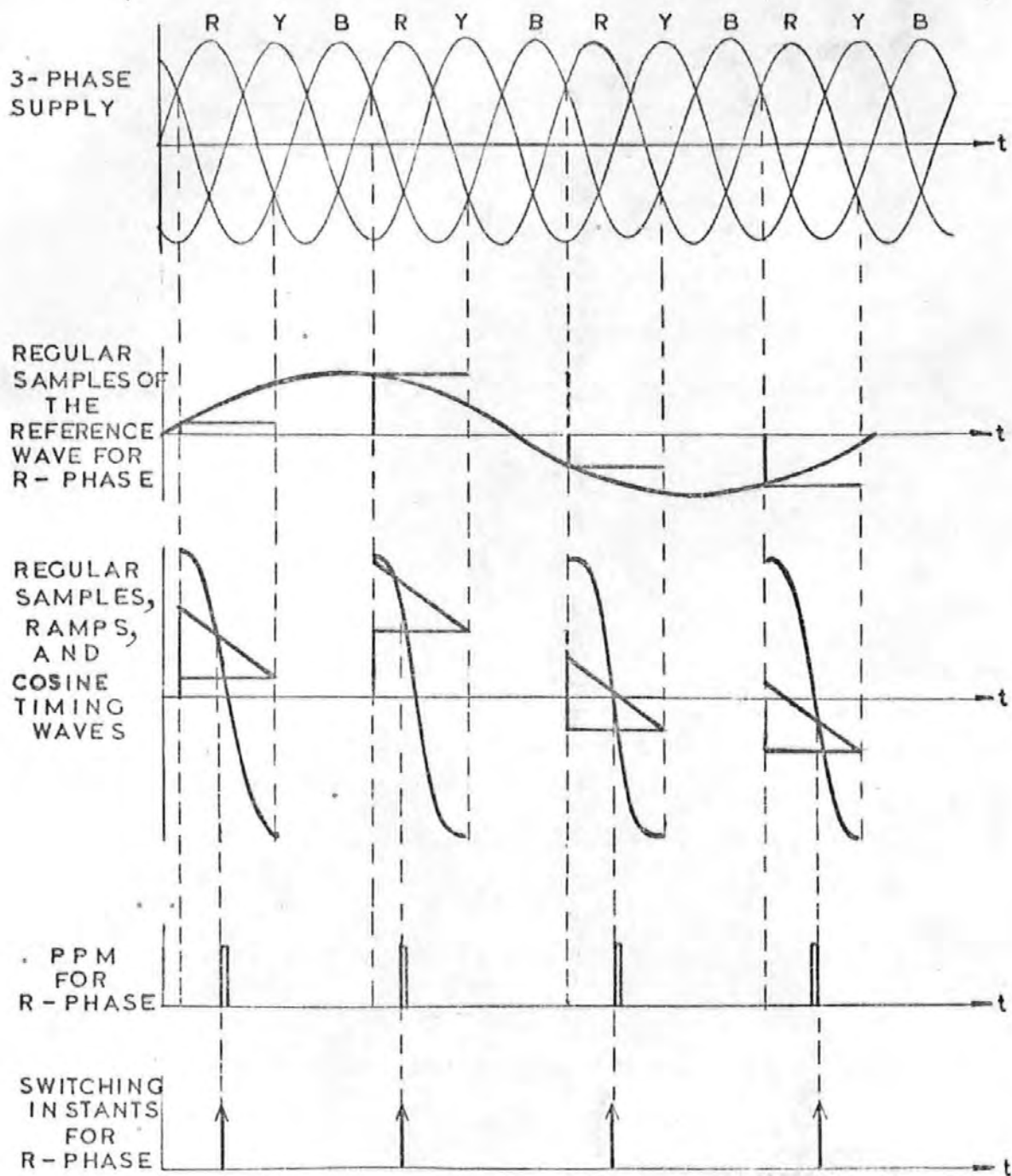
FIG 3·16 PULSE POSITION MODULATION FOR THE POSITIVE GROUP OF THE CYCLO – CONVERTER

OUTPUT FREQ. (HZ)	FREQUENCY OF COMPONENT (HZ)														
	0	5	10	15	20	25	30	35	40	45	50	55	60	65	70
5		100		13		8		6		5		4		3	
10			100				13				8				6
15				100			2			13			3		
20			2		100		2				1		14		6
25	1					100					7				
30							100								
35			6	2			1	100				3	3		
40			6		2		1		100		2		2		27
45				1			6			100			27		
50											100				

TABLE 3·5   3-PULSE 50HZ INHIBITED CYCLOCONVERTER :
MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS
AS PERCENTAGES OF THE WANTED COMPONENT.
PULSE POSITION MODULATION

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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.
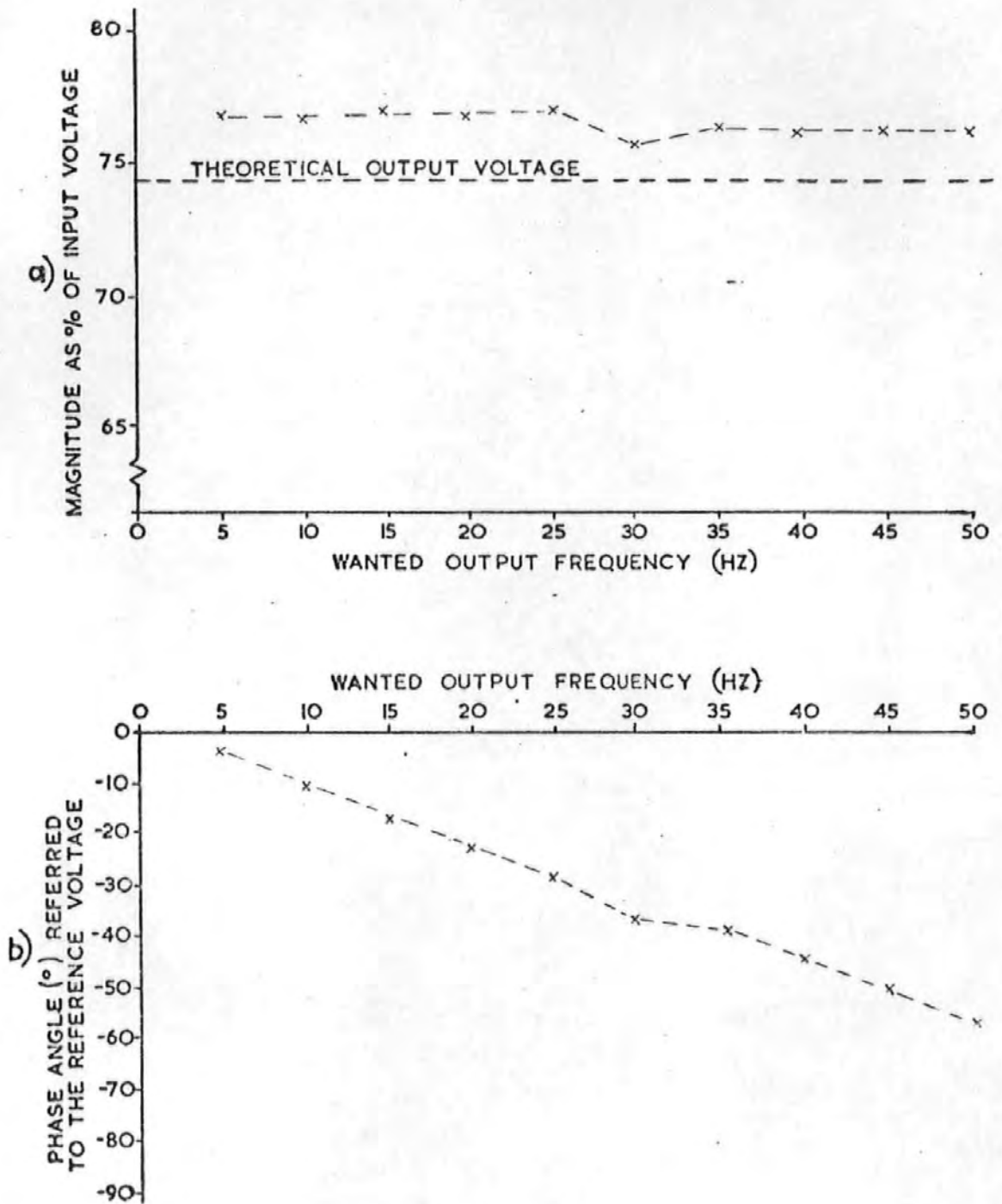
FIG 3·17 MAGNITUDE AND PHASE ANGLE OF THE WANTED
COMPONENT (INHIBITED CYCLOCONVERTER).
PULSE POSITION MODULATION

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OUTPUT
VOLTAGE
WAVEFORM

WANTED
COMPONENT

NETT
AREAS

$+$ $\circ$
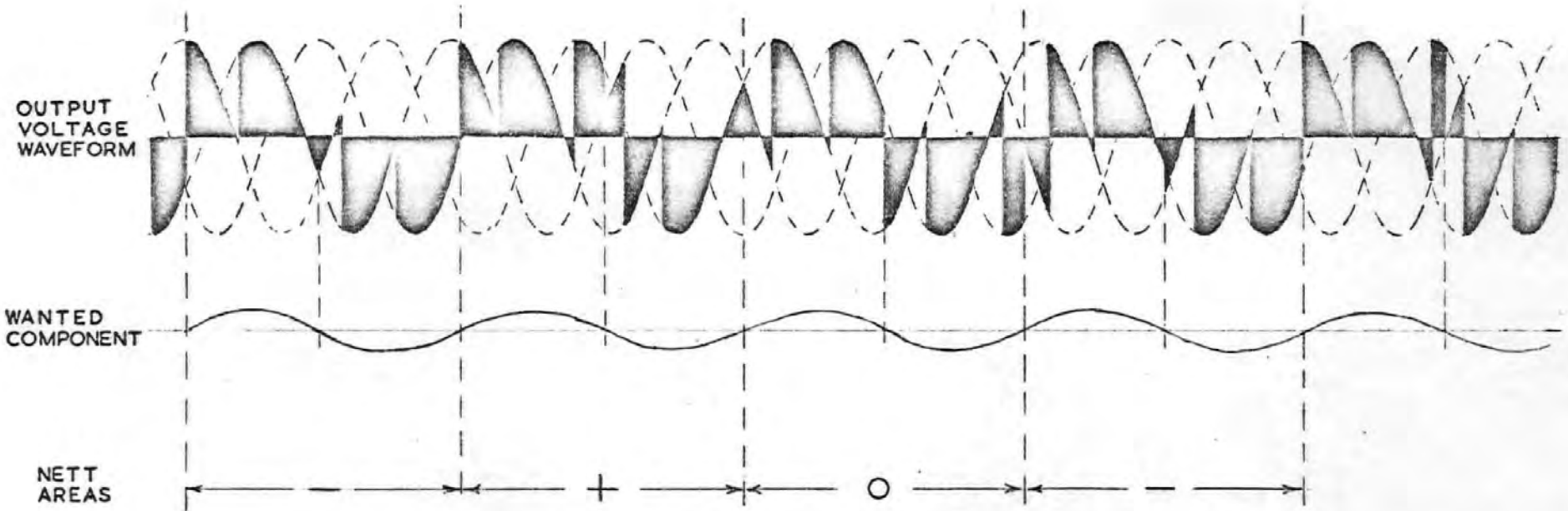
FIG 3·18 NETT AREAS ENCLOSED BY OUTPUT VOLTAGE WAVEFORM OF INHIBITED CYCLOCONVERTER.
UDF LOAD, SINUSOIDAL CURRENT ; PULSE POSITION MODULATION

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

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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[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[10] to be given by the following general relationships:

$$f_h = \left| 3(2m-1)f_i \pm 2nf_o \right| \qquad (4.1)$$

$$\text{and} \quad f_h = \left| 6mf_i \pm (2n+1)f_o \right| \qquad (4.2)$$

$$\text{where} \quad m = \text{any integer from 1 to infinity}$$

$$n = \text{any integer from 0 to infinity}$$

$$f_i = \text{input frequency}$$

$$f_o = \text{output frequency}$$

The frequencies of the components for $m = 1$ and $n = 0, 1, 2, 3, 4$, using expressions (4.1) and (4.2), are:

$$\left. \begin{array}{ll} n = 0 & f_h = 3f_i \\ n = 1 & f_h = 3f_i \pm 2f_o \\ n = 2 & f_h = 3f_i \pm 4f_o \\ n = 3 & f_h = 3f_i \pm 6f_o \\ n = 4 & f_h = 3f_i \pm 8f_o \end{array} \right\} \qquad (4.3)$$

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$$n = 0 \quad f_h = 6f_i \pm f_o$$
$$n = 1 \quad f_h = 6f_i \pm 3f_o$$
$$n = 2 \quad f_h = 6f_i \pm 5f_o \qquad \left. \right\} \quad (4.4)$$
$$n = 3 \quad f_h = 6f_i \pm 7f_o$$
$$n = 4 \quad f_h = 6f_i \pm 9f_o$$

and for $m = 2$ in expression (4.1):

$$n = 0 \quad f_h = 9f_i$$
$$n = 1 \quad f_h = 9f_i \pm 2f_o$$
$$n = 2 \quad f_h = 9f_i \pm 4f_o \qquad \left. \right\} \quad (4.5)$$
$$n = 3 \quad f_h = 9f_i \pm 6f_o$$
$$n = 4 \quad f_h = 9f_i \pm 8f_o$$

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 $6f_i$, $12f_i$ etc.

Thus, for a p-pulse inhibited cycloconverter, the frequencies of the harmonic components are given by

$$f_h = \left| mp \, f_i \pm (2n + k)f_o \right| \qquad (4.6)$$

where $k = 0$ for odd values of (mp)

$\qquad \qquad = 1$ 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 :

$$f_h = \left| mpf_i - (2n+k)f_0 \right| \tag{4.7}$$

This does not mean, however, that all components represented by expression (4.7) are sub-harmonic components.

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

$$f_h = \left| m\left(\frac{pf_i}{f_0}\right) - (2n+k) \right| f_0 \tag{4.8}$$

The factor (pf_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 $\left| m\left(\frac{pf_i}{f_0}\right) - (2n + k) \right| < 1.0$, 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 pf_i/f_0 Ratio.

(4.2.1) <u>Operation of the Cycloconverter at Integer</u>
         <u>pf_i/f_0 Ratio Values</u>

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

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pf_i/f_o Ratio is also an integer for the present discussion, then the coefficient of f_o can only be zero or an integer.

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

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

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

Therefore f_h is a d.c. component if the pf_i/f_o Ratio is an even-valued integer.

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

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

For _any_ integer value of pf_i/f_o, the coefficient of f_o is never zero, so that in this case f_h can never be a d.c. component.

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

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

In this case, f_h is a d.c. component if pf_i/f_o 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_i/f_o 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_i/f_o 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_i/f_o.

In terms of the output voltage waveform, operation of a cycloconverter at any integer pf_i/f_o 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_i/f_o Ratio Values

It has been shown in chapter 2 that the fundamental repetition frequency can be determined from the fractional form of the pf_i/f_o Ratio value; for any non-integer Ratio value, the fundamental repetition

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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_i/f_o, the coefficient of f_o in these expressions is always a non-integer, and that for certain combinations of values of (n) and (pf_i/f_o), 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_i/f_o Ratio Values

For decreasing values of pf_i/f_o (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 $\dfrac{1}{2f_o}$ 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

VOLTAGE

LOAD
CURRENT

INHIBITION
SWITCHING
FUNCTION

t_1

a) NO DELAY OF CURRENT ZERO

VOLTAGE

LOAD
CURRENT

INHIBITION
SWITCHING
FUNCTION

t_2

b) CURRENT ZERO DELAYED BY MODIFICATION
TO A THYRISTOR SWITCHING INSTANT
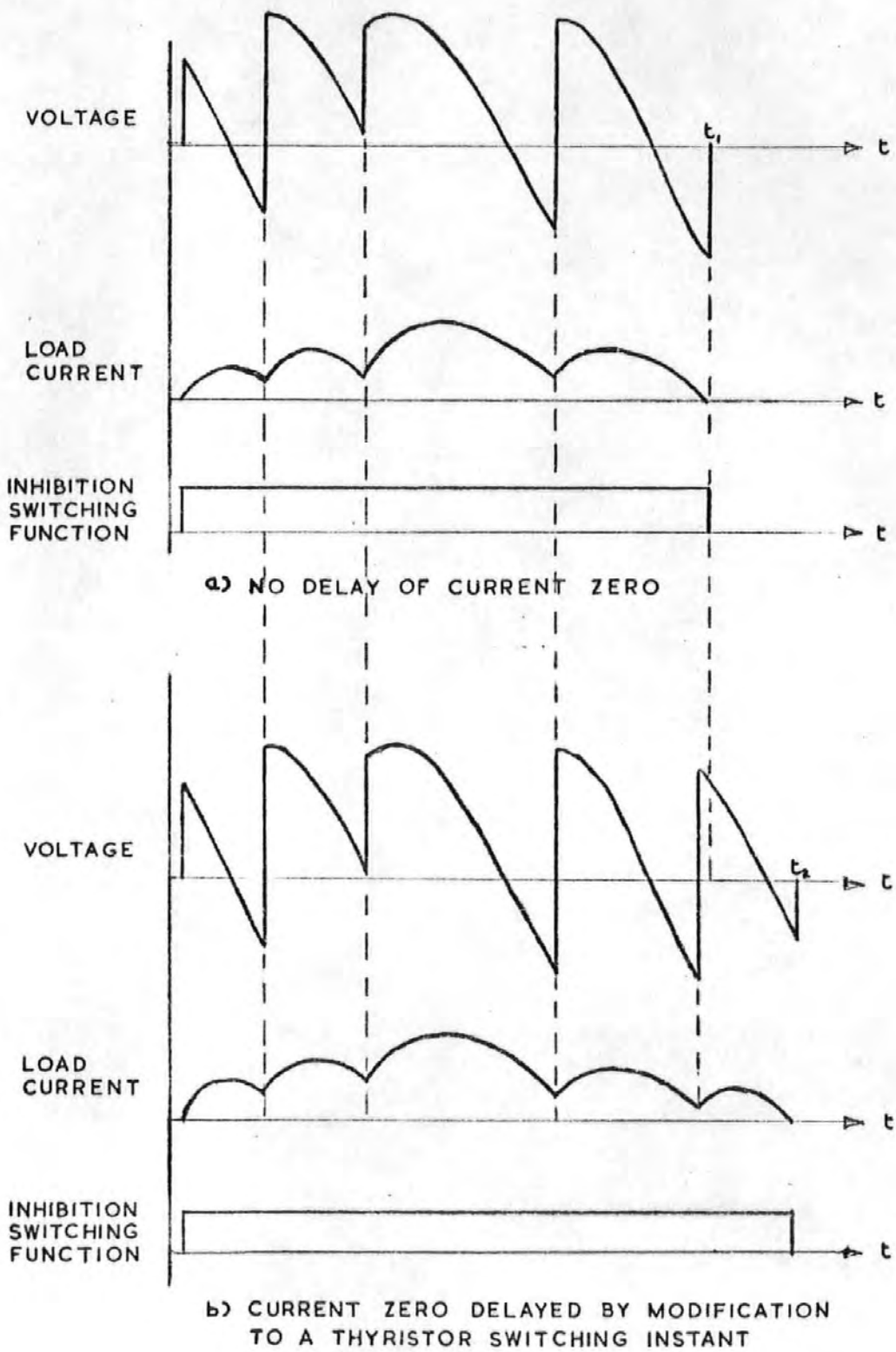
FIG. 4·1 MODIFICATION OF THE INHIBITION
SWITCHING FUNCTION

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_i/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_i/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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Ratio f_o/f_i	f_o/f_i %	0 (D.C.)	(Frequency of component) / f_o														
			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
3.0	100											100					
3.1	96.8	2	3	3	3	4	4	6	7	11	18	100	50	18	11	8	6
3.2	93.8	2		5		6		10		18		100		50		19	
3.3	91.0	1	7	6	1	11	5	2	19	4	2	100	3	2	49	3	2
3.4	88.2			16				21				100				46	
3.5	85.9	10					25					100				42	
3.6	83.4	2		11		19		3		8		100		4		6	
3.7	81.7		2	3	20	12	3	2	2	2	3	100	8	3	2	2	2
3.8	79.1			21				14				100				9	
3.9	77.0	1	19	2	4	2	3	4	2	13	2	100	2	4	3	2	4
4.0	75.0	15										100					

Right-side annotations: $6f_i-5f_o$ (row 3.3), $3f_i-2f_o$ (row 3.6), $6f_i-9f_o$ (row 3.7)

Bottom annotations: $3f_i-4f_o$, $6f_i-9f_o$, $6f_i-7f_o$

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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. pf_i/f_o RATIOS 3.0 TO 4.0

Ratio fo/fi	fo/fi %	O (D.C.)	(Frequency of component) / fo														
			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
4.0	75.0	15										100					
4.1	73.2	1	19	2	4	2	3	3	2	4	1	100	1	12	2	4	2
4.2	71.4			20				7				100				13	
4.3	69.8		2	3	20	4	2	2	2	3	5	100	3	2	2	2	3
4.4	68.2	1		4		20		3		3		100		5		3	
4.5	66.6	5					23					100					11
4.6	65.2			7				22				100				6	
4.7	63.8		3	3	2	6	2	1	22	2	2	100	2	3	4	1	3
4.8	62.4	1		3		4		6		22		100		4		3	
4.9	61.2		1	1	2	2	2	2	4	6	22	100	5	3	3	2	2
5.0	60.0											100					

Labels at right: $6f_i - 7f_o$ (row 4.3), $3f_i - 6f_o$ (row 4.4), $6f_i - 11f_o$ (row 4.7)

Labels at bottom: $6f_i - 11f_o$, $3f_i - 6f_o$, $3f_i - 4f_o$, $6f_i - 9f_o$

TABLE 4.1b 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT. $\rho f_i / f_o$ RATIOS 4.0 TO 5.0

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Ratio fo/fi %	O (D.C.)	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	
								(Frequency of component) / fo									
5.0 60.0											100						
5.1 58.8		1	1	1	2	2	3	3	3	5	100	23	6	4	3	2	
5.2 57.7	1		3		3		4		6		100		23		6		
5.3 56.6		3	2	1	3	2	1	5	1	1	100	2	1	23	1	1	$6f_i-9f_o$
5.4 55.5			6				7				100				24		
5.5 54.5	2					5					100					24	
5.6 53.5	1		4		6		2		3		100		2		2		$3f_i-4f_o$
5.7 52.6		1	2	6	4	2	1	1	1	2	100	3	1	1	1	1	$6f_i-13f_o$
5.8 51.7			8				5				100				4		
5.9 50.8		7	1	2	1	1	2	1	4	1	100	1	3	1	1	2	
6.0 50.0	6										100						

$3f_i-6f_o$ $6f_i-13f_o$ $6f_i-11f_o$

TABLE 4.1c   3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT. f_i/f_o RATIOS 5.0 TO 6.0

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Ratio f_o/f_i	%	0 (D.C.)	(Frequency of component) / fo														
			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
6.0	50.0	6										100					
6.1	49.2		6	1	2	1	1	2	1	3		100	1	4	1	2	1
6.2	48.4			8				4				100				5	
6.3	47.6		1	2	7	3		1	1	1	3	100	2	1	1	1	1
6.4	46.8			3		8		2		3		100		3		2	
6.5	46.2	2					7					100					4
6.6	45.4			4				8				100				5	
6.7	44.8		2	2		3	1	1	8	1	1	100	1	1	4	1	1
6.8	44.2	1		2		2		3		7		100		4		3	
6.9	43.4		1	1	1	1	1	1	2	3	8	100	4	2	2	1	1
7.0	42.8											100					

$6f_i - 11f_o$
$3f_i - 8f_o$
$6f_i - 15f_o$

$6f_i - 15f_o$ $6f_i - 13f_o$
$3f_i - 8f_o$ $3f_i - 6f_o$

TABLE 4.1d 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING: MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT.
pf_i/f_o RATIOS 6.0 TO 7.0

Ratio f_o/f_i	%	0 (D.c.)	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
				(Frequency of component) / f_o														
7.0	42.8											100					
7.1	42.2		1		1		1	1	2	2	4	100	7	3	2	2	1
7.2	41.6			1		1		2		3		100		8		4	
7.3	41.0		2	2		2		1	4	1	1	100	1		8		
7.4	40.6			3			5					100				8	
7.5	40.0	2					5					100					9
7.6	39.4			3		5		2		2		100		2		2	
7.7	39.0		1	1	4	3	1				2	100	2				1
7.8	38.4			5				3				100				3	
7.9	38.0		5		2	1	1	2		2	1	100	1	2	1	1	1
8.0	37.4	3										100					

$6f_i-13f_o$
$3f_i-6f_o$
$3f_i-8f_o$
$6f_i-15f_o$

TABLE 4.1e 3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING : MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT. pf_i/f_o RATIOS 7.0 TO 8.0

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Ratio f_o/f_i	f_o/f_i %	0 (D.C.)	(Frequency of component) / f_o														
			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
8.0	37.4	3										100					
8.1	37.0		4		2			1		2		100		2		1	
8.2	36.6			5				3				100				3	
8.3	36.2			1	5	2	1				2	100	1				1
8.4	35.8			2		5		1		2		100		2		2	
8.5	35.2	1					5					100					3
8.6	34.8			3				5				100				3	
8.7	34.4		1	1		2			5			100			3		
8.8	34.0			1		2		2		5		100		3		2	
8.9	33.6		1					1	2	2	5	100	3	2	1		
9.0	33.3											100					

$6f_i-15f_o$ $3f_i-10f_o$

$3f_i-10f_o$ $3f_i-8f_o$

TABLE 4.1f   3-PULSE INHIBITED CYCLOCONVERTER WITH NATURAL SAMPLING : MAGNITUDES OF THE LOW-FREQUENCY COMPONENTS AS PERCENTAGES OF THE WANTED COMPONENT. $p f_i/f_o$ RATIOS 8.0 TO 9.0

is expressed as a proportion of the wanted output frequency (f_o). 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_i/f_o 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_i/f_o Ratio values which are all integer values. For example, when the pf_i/f_o 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 $0.9f_o$ 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 $3f_i$ harmonic family gives sub-harmonic components for a 2.0 range of Ratio Values, whilst the $6f_i$ harmonic family gives sub-harmonic components over a range of only 1.0. Thus the $3f_i$ harmonic family causes sub-harmonic distortion over a wider range of output frequencies than the $6f_i$ 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)

a) $3f_i$ HARMONICS

$3f_i - 2f_o$
$3f_i - 4f_o$
$3f_i - 6f_o$
$3f_i - 8f_o$
$3f_i - 10f_o$

b) $6f_i$ HARMONICS

$6f_i - 5f_o$ $6f_i - 7f_o$ $6f_i - 9f_o$ $6f_i - 11f_o$ $6f_i - 13f_o$ $6f_i - 15f_o$ $6f_i - 17f_o$ $6f_i - 19f_o$
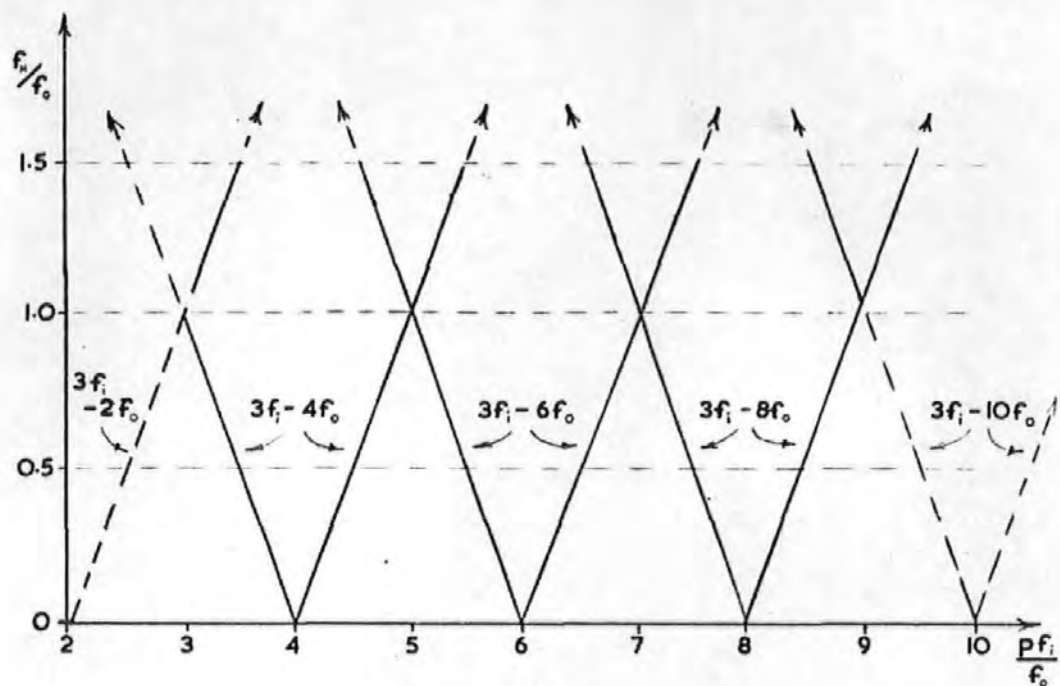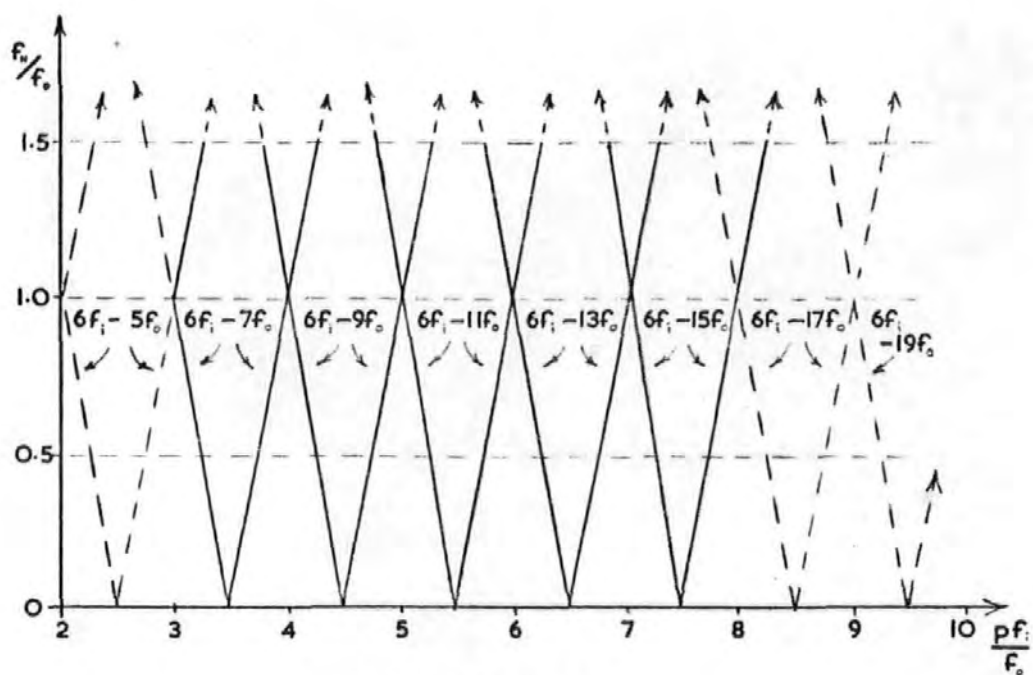
FIG 4·2   VARIATION OF THE FREQUENCIES OF THE SIGNIFICANT GENERAL HARMONICS WITH THE $\frac{pf_i}{f_o}$ RATIO

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.with the pf_i/f_o 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 pf_i/f_o 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_o.

Further data from the computer analysis showed that the $(3f_i-6f_o)$ and the $(6f_i-9f_o)$ harmonics in a 3-phase output each form co-phasal sets, whilst the other harmonics form 3-phase sets. The $(3f_i-6f_o)$ and $(6f_i-9f_o)$ 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 $(3f_i-4f_o)$ 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

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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 pf_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
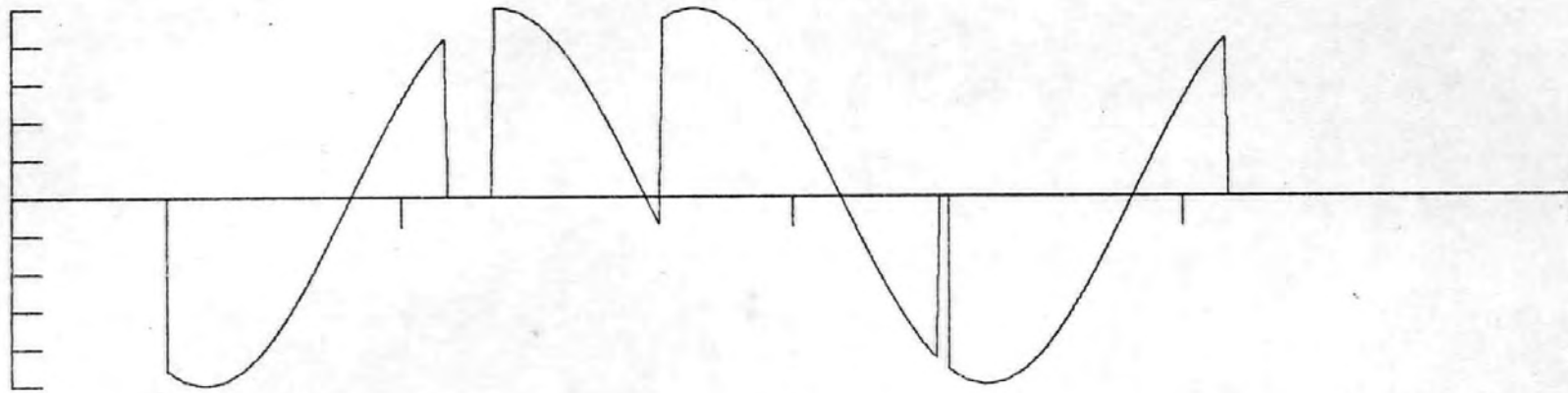
OUTPUT VOLTAGE WAVEFORM, NATURAL SAMPLING
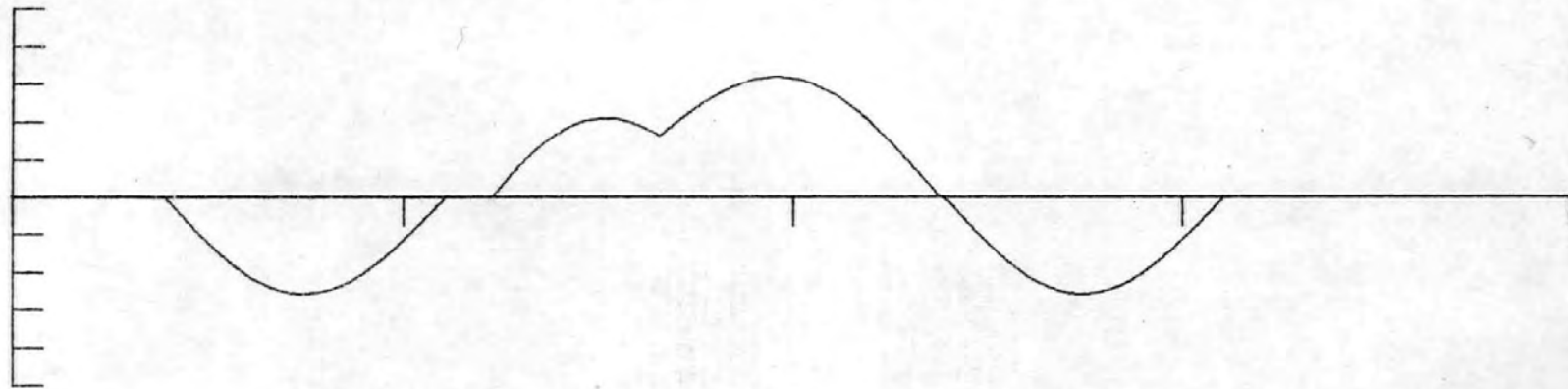
LOAD CURRENT WAVEFORM, NATURAL SAMPLING

FIG 4·3 COMPUTER GRAPH PLOT FOR $\frac{Pf}{F_0}$ RATIO= 4·0
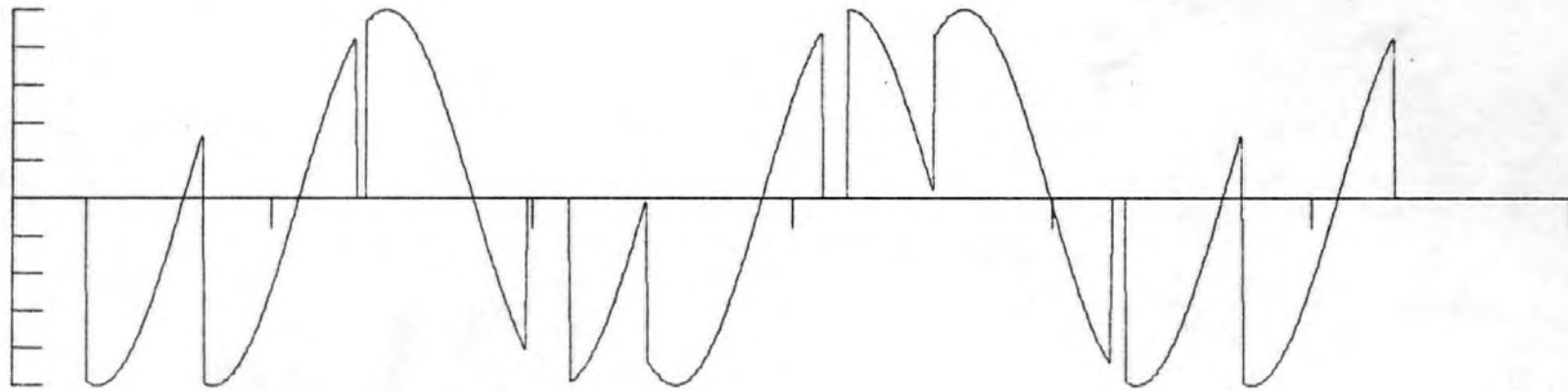
OUTPUT VOLTAGE WAVEFORM.NATURAL SAMPLING
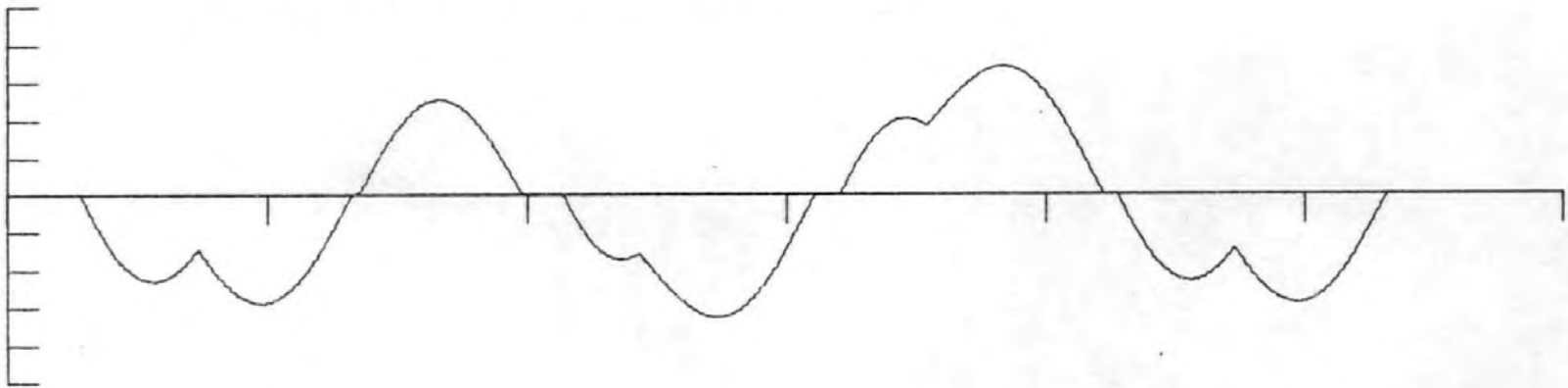
LOAD CURRENT WAVEFORM.NATURAL SAMPLING

FIG 4.4 COMPUTER GRAPH PLOT FOR $\frac{p.f.}{f_o}$ RATIO= 4.5
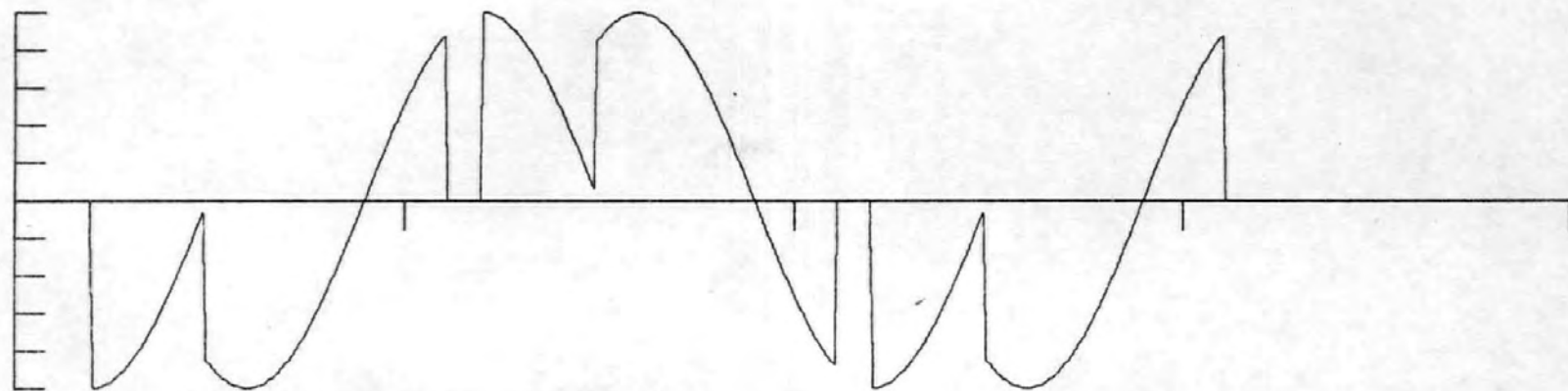
OUTPUT VOLTAGE WAVEFORM. NATURAL SAMPLING
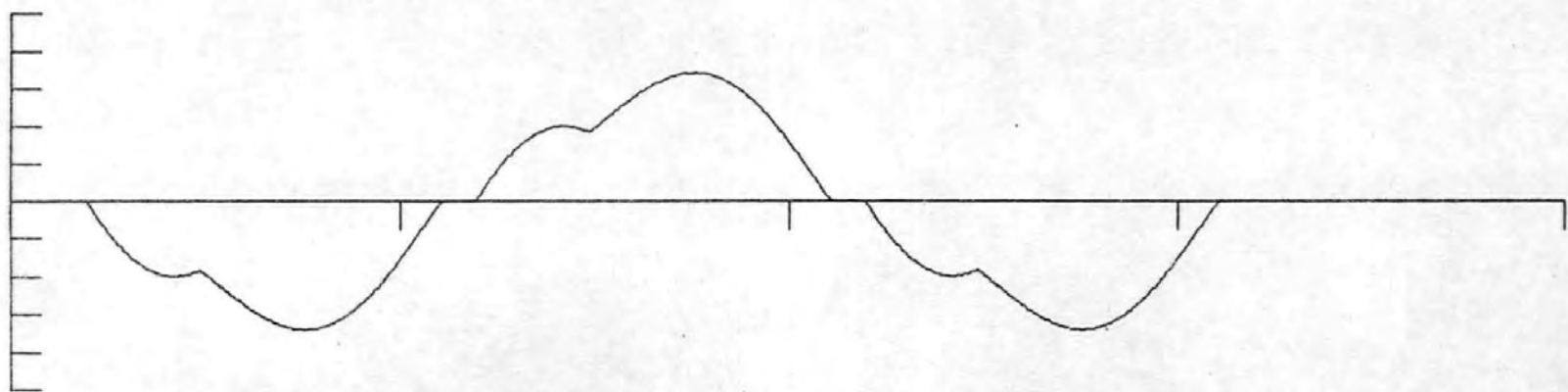
LOAD CURRENT WAVEFORM. NATURAL SAMPLING

FIG 4.5 COMPUTER GRAPH PLOT FOR $\frac{p f}{f_o}$ RATIO = 5.0
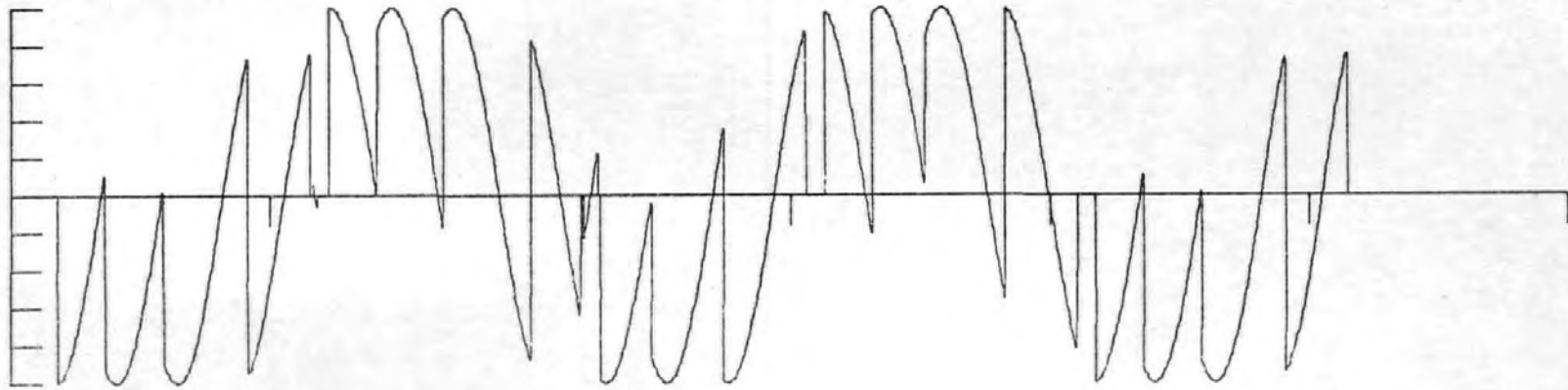
OUTPUT VOLTAGE WAVEFORM. NATURAL SAMPLING
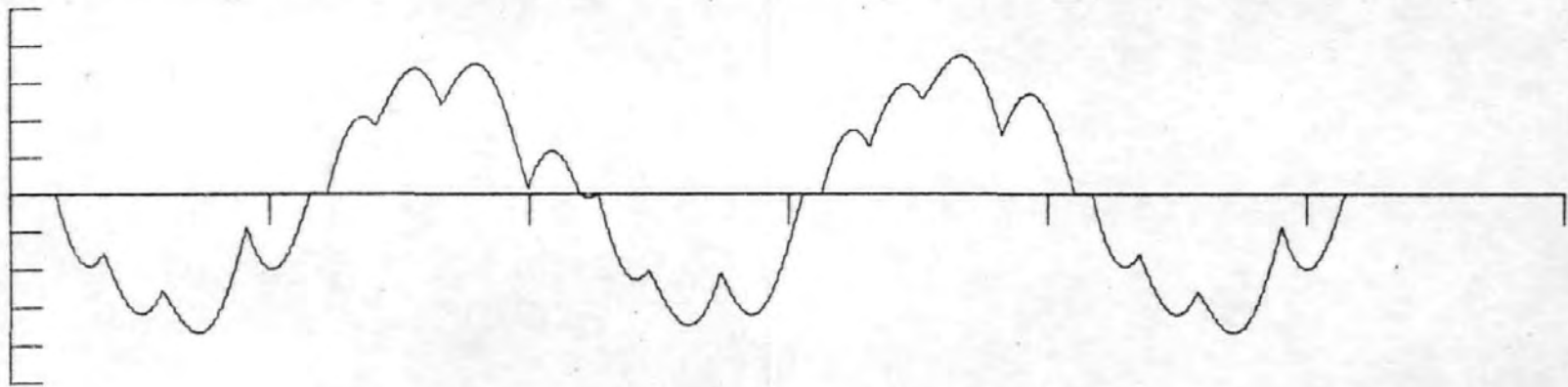
LOAD CURRENT WAVEFORM. NATURAL SAMPLING

FIG 4·6 COMPUTER GRAPH PLOT FOR $\frac{pf_i}{F_o}$ RATIO = 8·5

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output component for each graph plot,
appropriate pf_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 pf_i/f_o 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 pf_i/f_o 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) <u>Application of Modified Control Techniques</u>
<u>to the Inhibited Cycloconverter when</u>
<u>supplying Non-sinusoidal Load Current</u>

(4.4.1) <u>General</u>

It has been shown in this chapter that for any specified pf_1/f_o Ratio, the frequency of a sub-harmonic component is only a particular value of a general harmonic. For any other pf_1/f_o 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

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 pf_i/f_o 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

Ratio f_o/f_i	%	0 (D.c)	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
								(Frequency of component) / fo									
3.8	79.1			16				11				100				10	
4.8	62.4	2		3		4		6		16		100		8		5	
5.8	51.7			9				6				100				5	
6.8	44.7	1		2		3		4		8		100		5		3	
7.8	38.4			6				4				100				4	
8.8	34.0	1		2		2		3		6		100		4		2	

TABLE 4.2   3-PULSE INHIBITED CYCLOCONVERTER : MAGNITUDES OF THE LOW FREQUENCY COMPONENTS FOR REGULAR SAMPLING
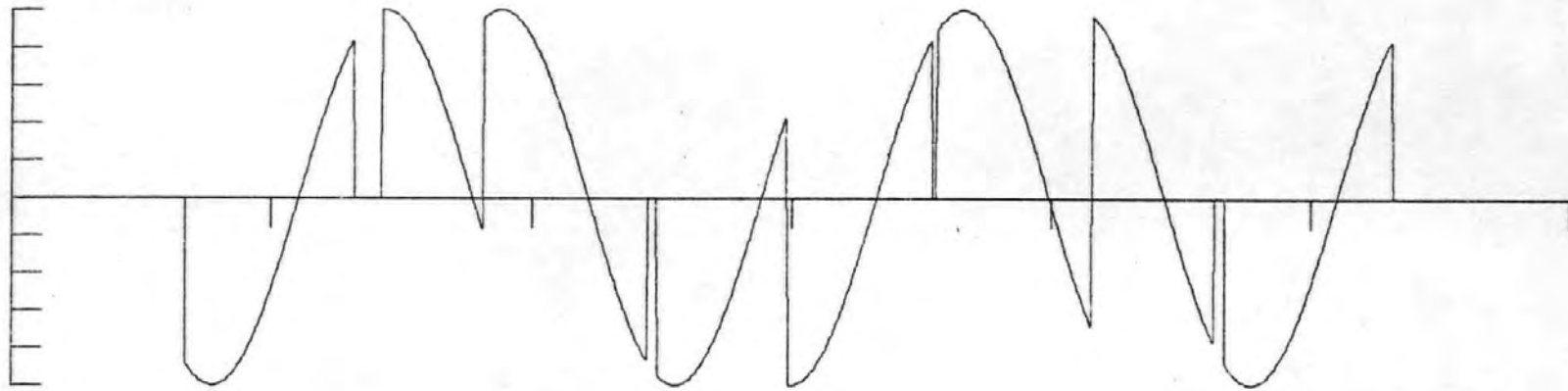
OUTPUT VOLTAGE WAVEFORM. REGULAR SAMPLING
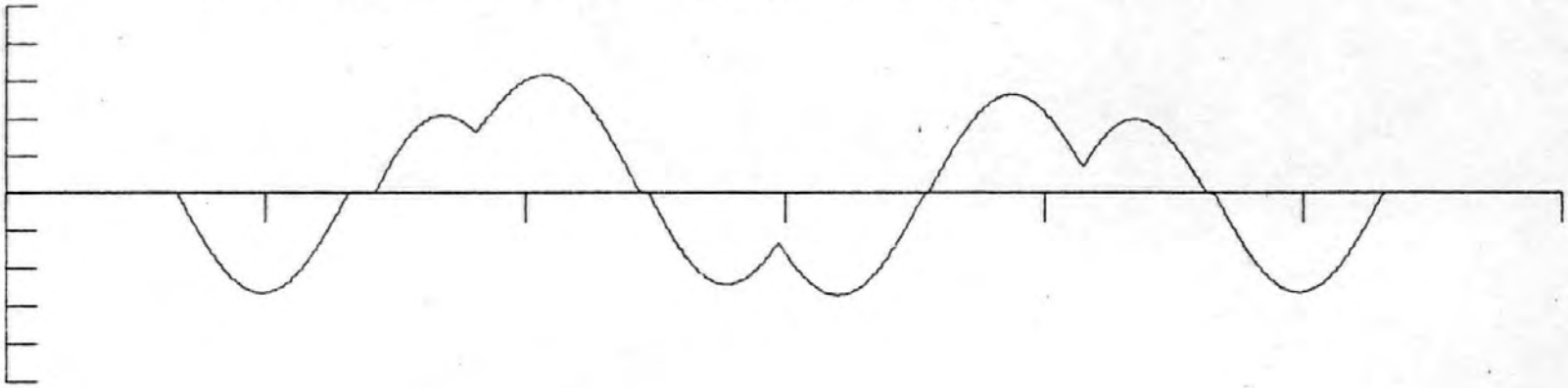
LOAD CURRENT WAVEFORM. REGULAR SAMPLING

FIG 4·7 COMPUTER GRAPH PLOT FOR $\frac{p.f.}{f_o}$ 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_i/f_o 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.

Ratio f_o/f_i	%	0 (D.c)	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
						(Frequency of component) / f_o											
3.8	79.1			20				14				100				11	
4.8	62.4	2		5		6		8		20		100		12		7	
5.8	51.7			11				6				100				6	
6.8	44.2	1		3		2		5		11		100		6		5	
7.8	38.4			7				5				100				4	
8.8	34.0	1		2		2		3		7		100		5		3	

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 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_i/f_o 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 pf_1/f_o 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_o/f_i	%	0 (D.c)	(Frequency of component) / f_o

Ratio f_o/f_i	%	0 (D.c)	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				8	
4.8	62.4	1		3		3		6		9		100		11		3	
5.8	51.7			11				5				100				5	
6.8	44.2			2		2		3		8		100		3		3	
7.8	38.4			4				3				100				3	
8.8	34.0			1		2		2		5		100		3		2	

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
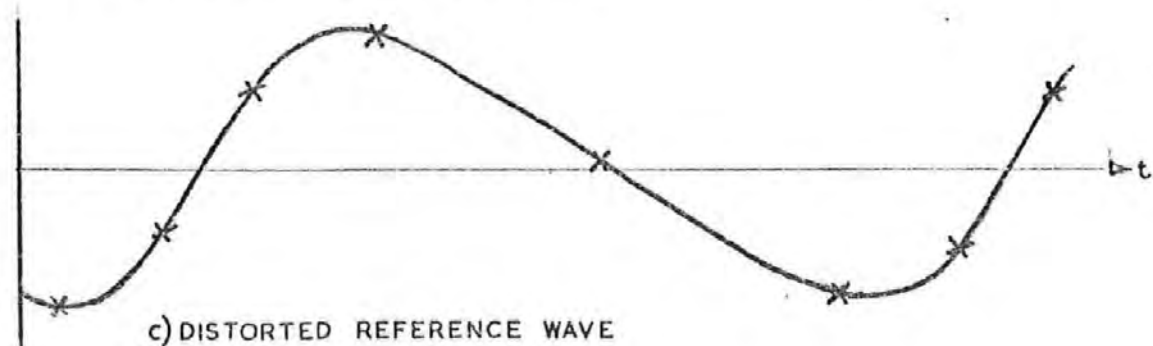distortion.

a) COSINUSOIDAL CONTROL WITH REGULAR SAMPLING

b) SAW-TOOTH REFERENCE WAVE

c) DISTORTED REFERENCE WAVE
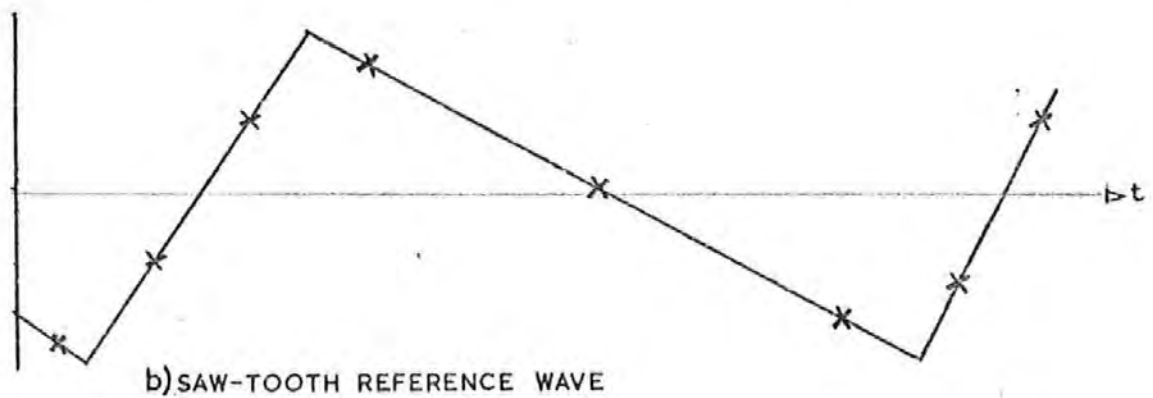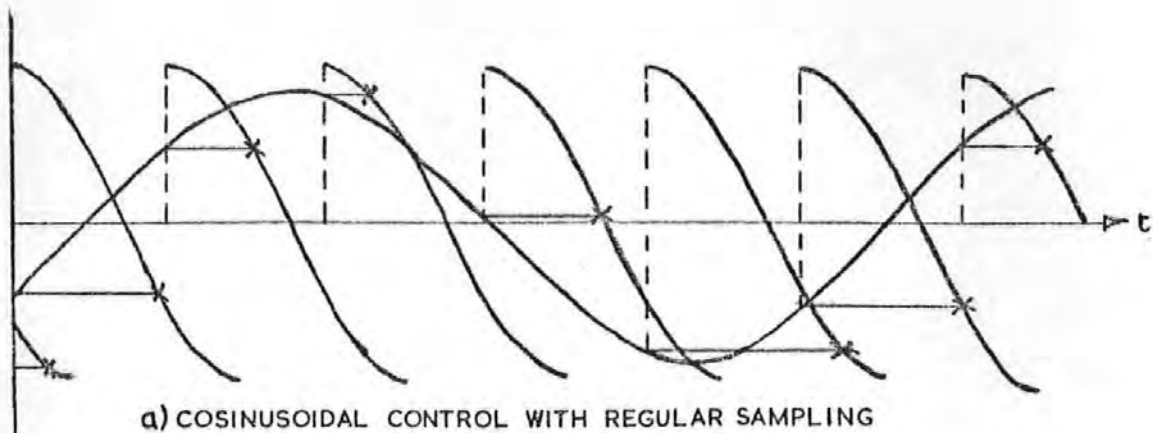
FIG 4·8 EQUIVALENCE OF REGULAR SAMPLING OF A
SINUSOIDAL REFERENCE WAVE AND NATURAL
SAMPLING OF A NON-SINUSOIDAL REFERENCE WAVE

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 pf_i/f_o 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 $1.2f_o$ 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. These components are both specific values of the $(3f_i - 6f_o)$ general harmonic.

POSITIVE
GROUP

$\alpha = 0$ $\alpha = 0$

SWITCHING INSTANTS
FOR SINUSOIDAL REFERENCE

SWITCHING INSTANTS
FOR DISTORTED REFERENCE

NEGATIVE
GROUP

$\alpha = 0$ $\alpha = 0$

SWITCHING INSTANTS
FOR SINUSOIDAL REFERENCE

SWITCHING INSTANTS
FOR DISTORTED REFERENCE
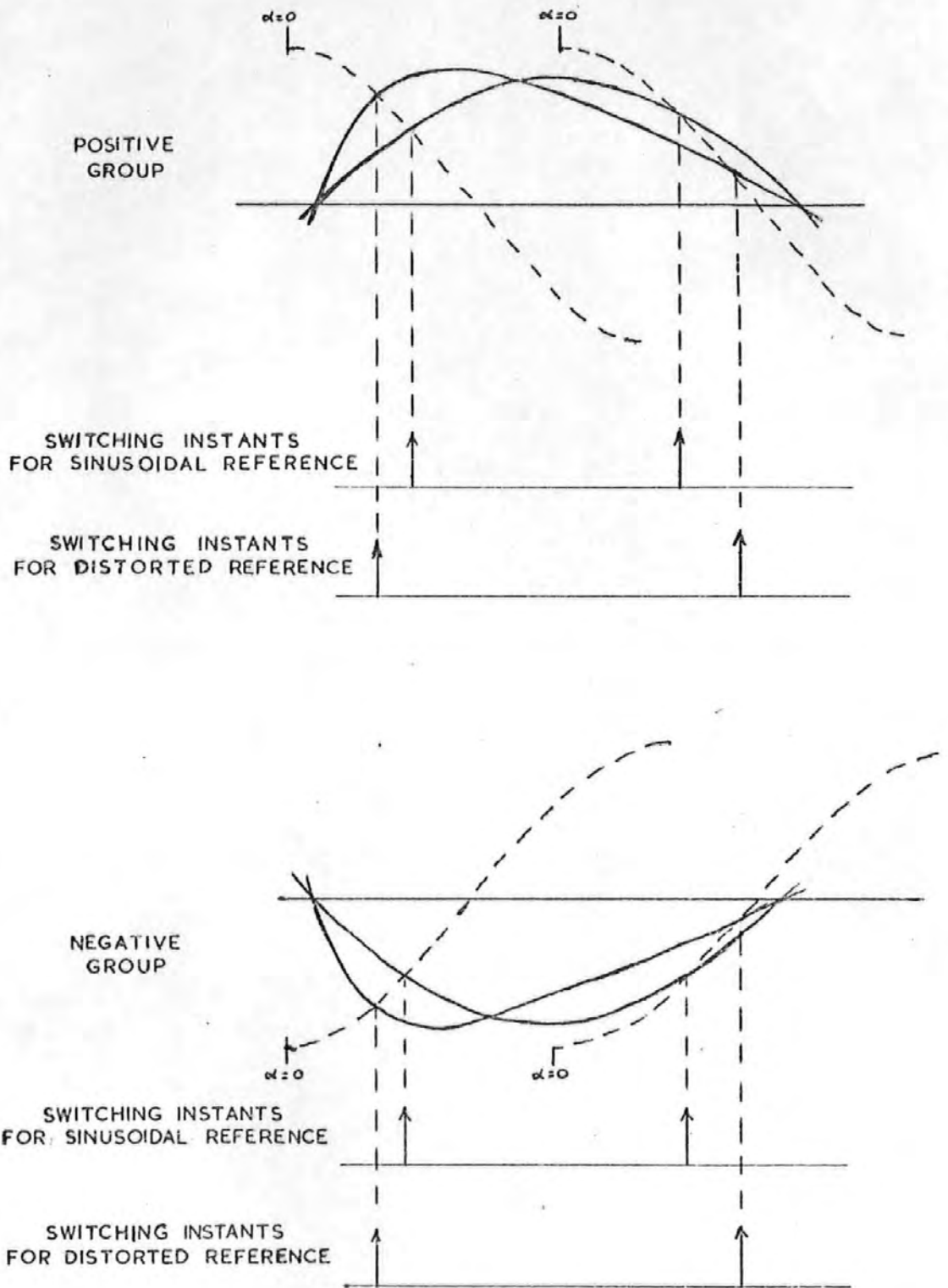
FIG 4·9 EFFECT OF DISTORTED REFERENCE
WAVE ON THE SWITCHING INSTANTS

Ratio f_o/f_i	%	0 (D.c)	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
						(Frequency of component) $/ f_o$											
3.8	79.1			8			4					100				5	
4.8	62.4			2		1	6		9			100		13		3	
5.8	51.7			12			2					100				5	
6.8	44.2			1		1	3		8			100		2		3	
7.8	38.4			3			3					100				2	
8.8	34.0	1		1		2	2		5			100		3		1	

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 $1.2f_0$ component at a pf_i/f_0 Ratio of 4.8 and the $0.2f_0$ component at a pf_i/f_0 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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Ratio f_o/f_i	%	O (D.c)	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			8				4				100				5	
4.8	62.4	1		1		3		4		10		100		10		6	
5.8	51.7			6				5				100				2	
6.8	44.2			2		2		2		3		100		5		1	
7.8	38.4			4				2				100				2	
8.8	34.0			1		1		2		5		100		3		2	

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 $(3f_i - 4f_o)$ HARMONIC
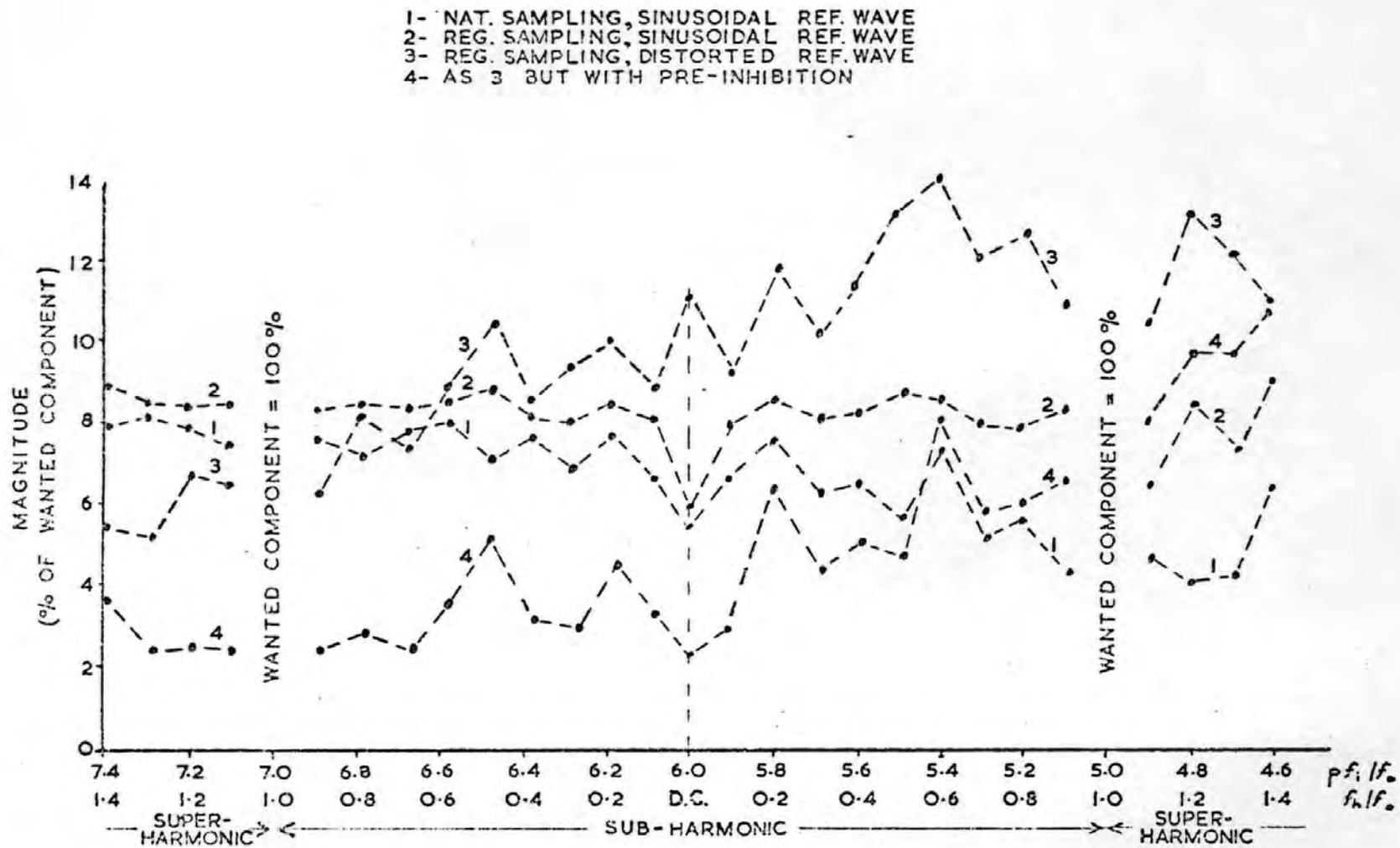
FIG 4·11   MAGNITUDE OF THE $(3f_i - 6f_o)$ HARMONIC

FIG 4·12   MAGNITUDE OF THE $(3f_i - 8f_o)$ HARMONIC
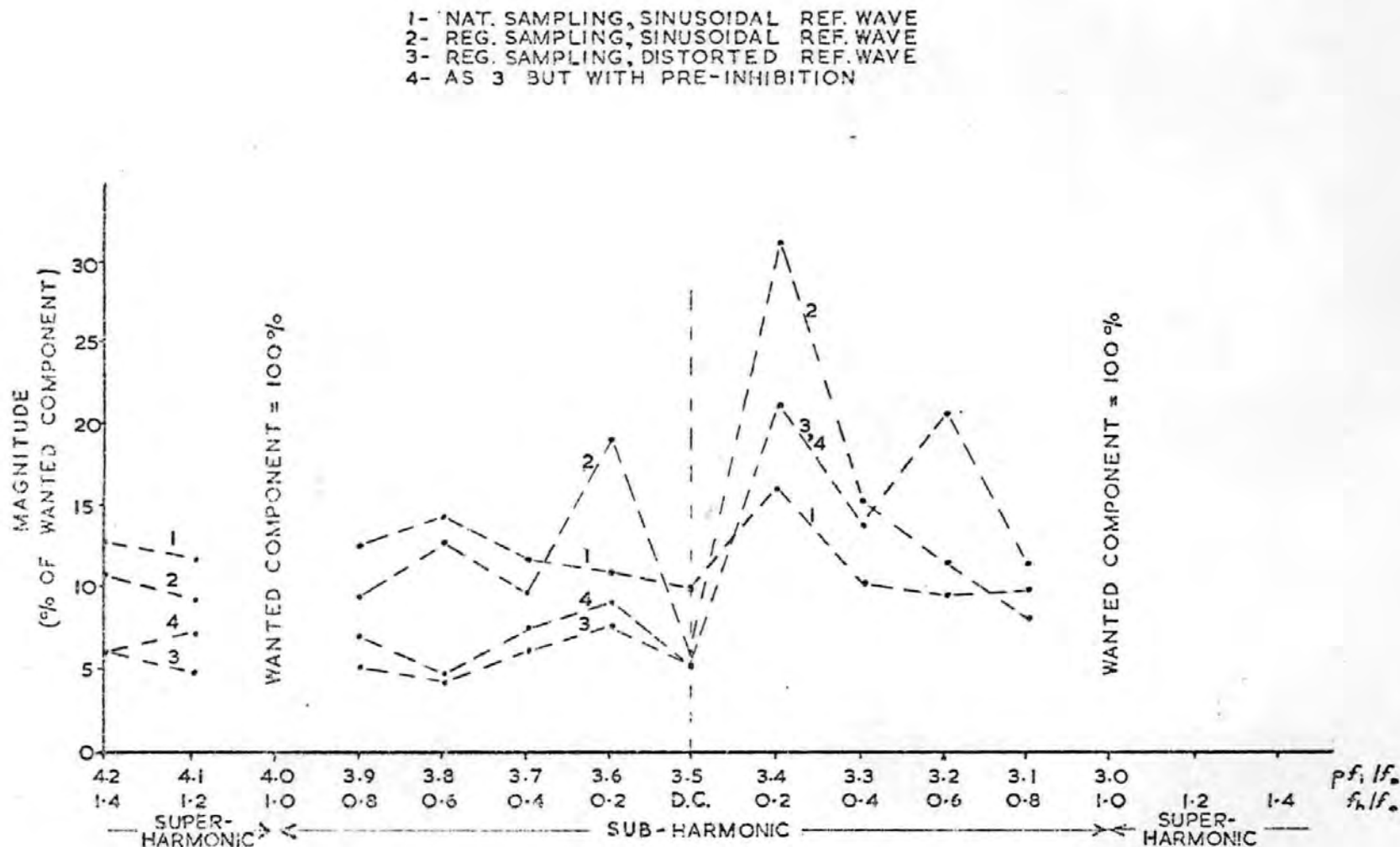
FIG 4·13 MAGNITUDE OF THE $(6f_i - 7f_o)$ HARMONIC

FIG 4·14   MAGNITUDE OF THE $(6f_i - 9f_o)$ HARMONIC

FIG 4·15 MAGNITUDE OF THE $(6f_i - 11f_0)$ HARMONIC

FIG 4·16   MAGNITUDE OF THE $(6f_i - 13f_o)$ 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 $(6f_i - 9f_o)$ harmonic is attenuated from a mean level of 6% of the wanted component to 3%; the $(6f_i - 11f_o)$ is attenuated from 4% to 2% (with pre-inhibition applied) and the $(6f_i - 13f_o)$ harmonic is attenuated from $3\frac{1}{2}\%$ 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 $(3f_i - 6f_o)$, $(3f_i - 8f_o)$ and the $(6f_i - 7f_o)$ 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_i/f_o 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
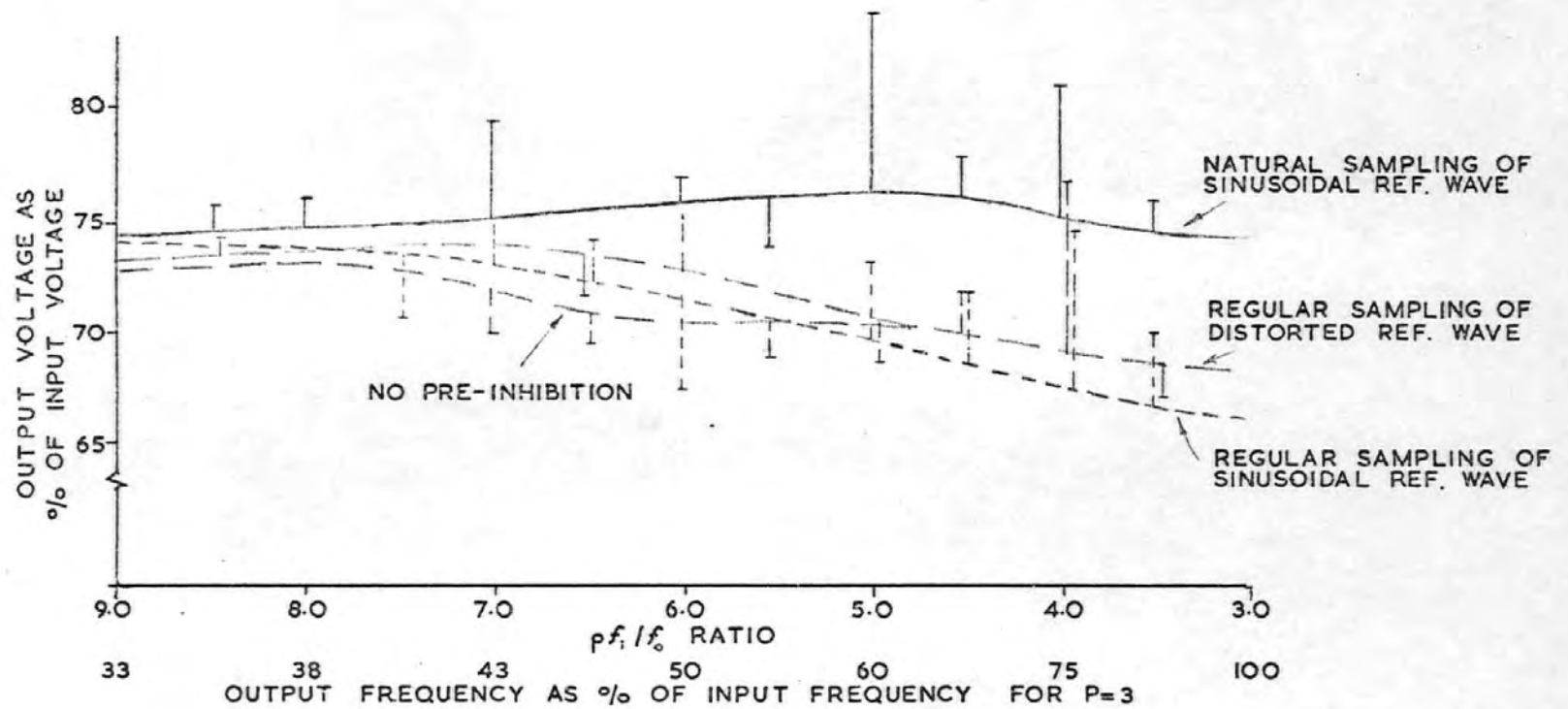
FIG 4·17 VARIATION IN OUTPUT VOLTAGE WITH OUTPUT FREQUENCY

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 pf_i/f_o Ratio is decreased. The drop in output voltage is about 6% - 7% of the input voltage.

At specific values of the pf_i/f_o 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 pf_i/f_o 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:

$$mpf_i - qf_o = f_o \qquad \text{where m and q are}$$

$$\text{or} \quad \frac{pf_i}{f_o} = \frac{1 + q}{m} \qquad \text{integer numbers}$$

For the $3f_i$ family of harmonics, $m = 1$; for the $6f_i$ 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 pf_i/f_o Ratio is an integer value.

In figure 4.17 it is also seen that discontinuities occur when the pf_i/f_o 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 $(12f_i - 17f_o)$ harmonic for a 3-pulse cycloconverter for which

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

Such harmonics as the $(12f_i - 17f_o)$ 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_i/f_o Ratio, but their magnitudes are generally low enough to have insignificant effect on the output voltage.

CHAPTER 5

EXPERIMENTAL WORK

(5.1) Introduction

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 $10k\Omega$ 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
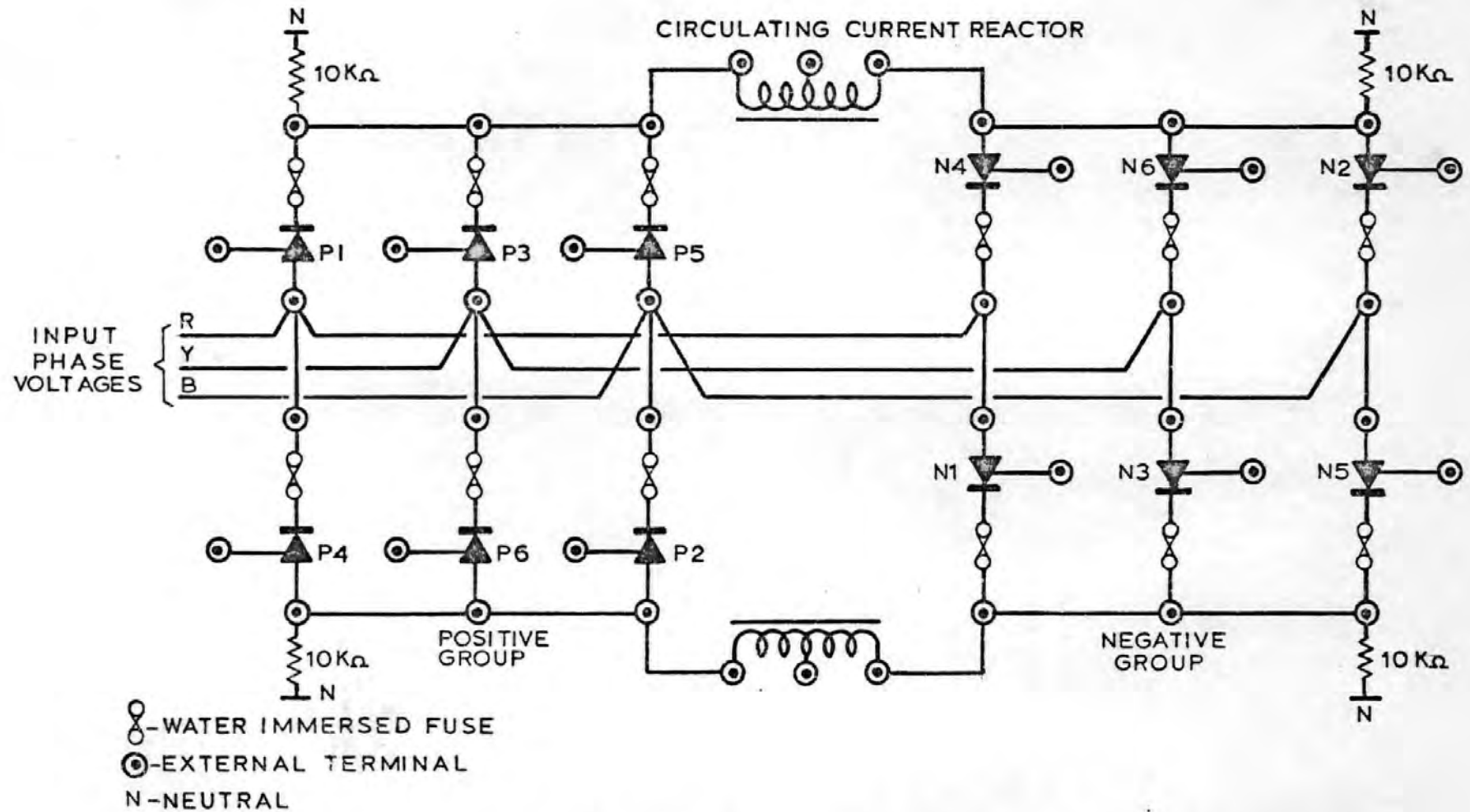
FIG. 5·1 EXTERNAL CONNECTIONS OF ONE OUTPUT PHASE OF
THE EXPERIMENTAL CYCLOCONVERTER
(SHOWN CONNECTED AS A 6-PULSE CYCLOCONVERTER)

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.

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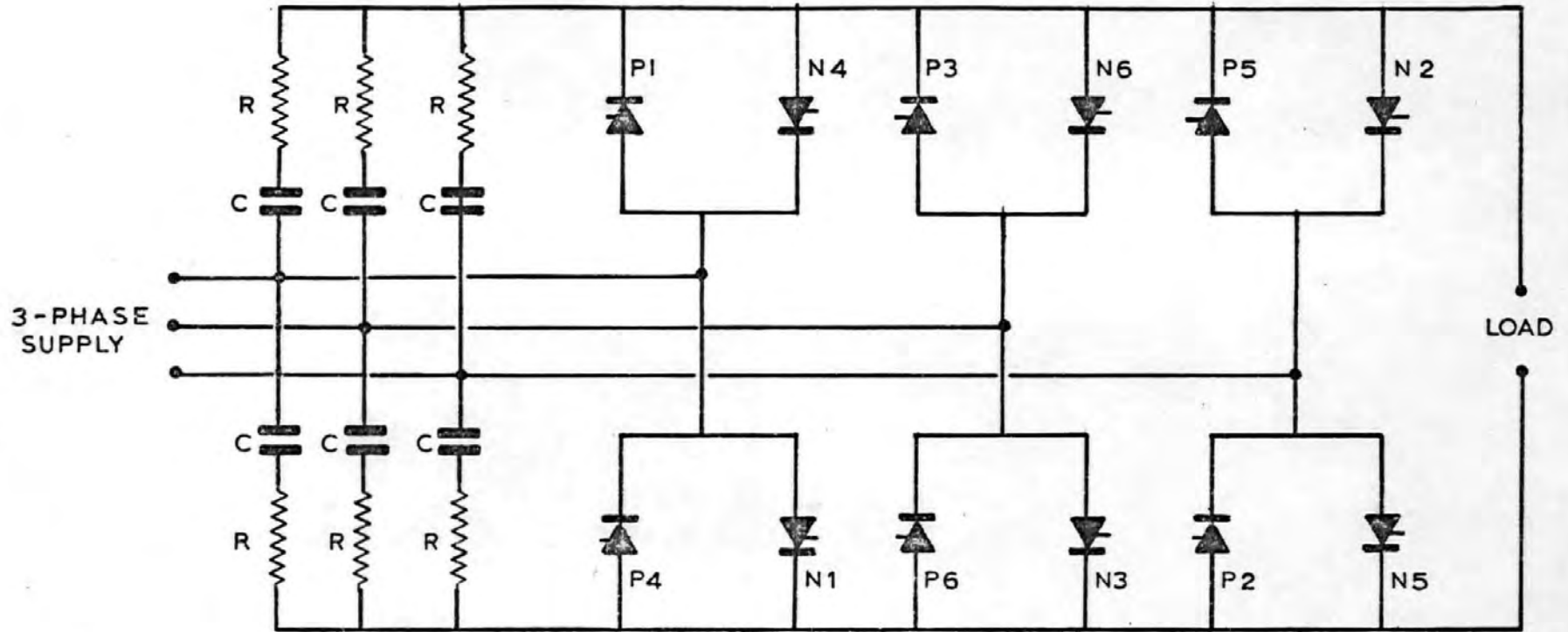
FIG. 5·2   R-C SNUBBER CIRCUITS FOR THE
INHIBITED CYCLOCONVERTER

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 240/12-0-12 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.

415 v 3-PHASE MAINS SUPPLY

R

270 Ω
100 Ω
270 Ω 100 Ω

PULSE-FORMING CIRCUITS FOR THYRISTORS P2, N5

PULSE-FORMING CIRCUITS FOR THYRISTORS P5, N2

Y

270 Ω
100 Ω
270 Ω 100 Ω

PULSE-FORMING CIRCUITS FOR THYRISTORS P4, N1

PULSE-FORMING CIRCUITS FOR THYRISTORS P1, N4

B

270 Ω
100 Ω
270 Ω 100 Ω

PULSE-FORMING CIRCUITS FOR THYRISTORS P6, N3

PULSE-FORMING CIRCUITS FOR THYRISTORS P3, N6

240 / 12-0 -12 v

N

OV   REFERENCE VOLTAGE

FIG. 5·3   INPUT STAGE OF FIRING CONTROL CIRCUIT FOR 6-PULSE CYCLOCONVERTER WITH SINGLE PHASE OUTPUT

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a
6 - PHASE
OUTPUT
FROM MAINS
TRANSFORMER

b
FOR
COMMUTATIONS
INVOLVING
THYRISTORS
P1, P3, P5

c
FOR
COMMUTATIONS
INVOLVING
THYRISTORS
P4, P6, P2

d
FOR
COMMUTATIONS
INVOLVING
THYRISTORS
N4, N6, N2

e
FOR
COMMUTATIONS
INVOLVING
THYRISTORS
N1, N3, N5
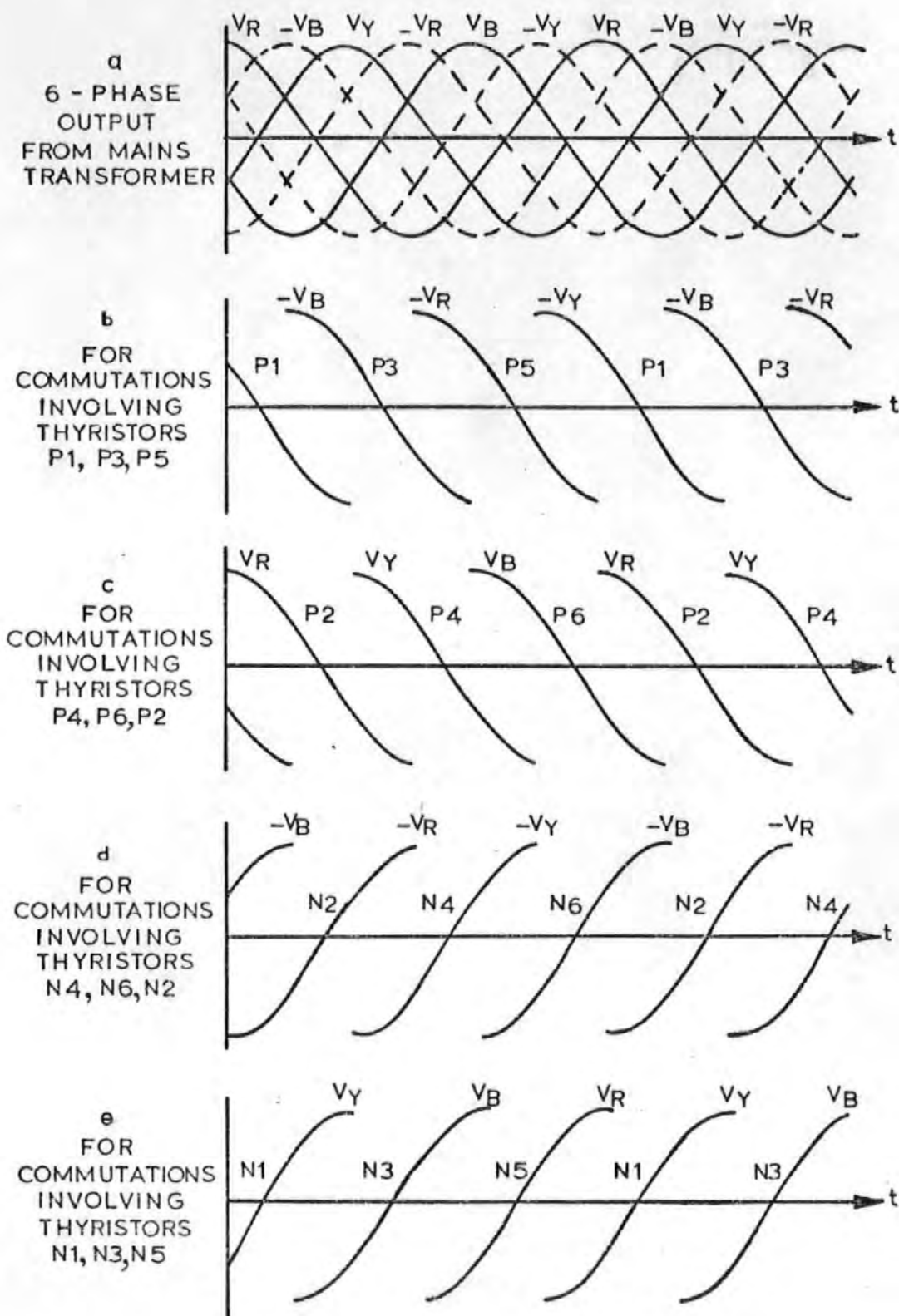
FIG 5·4   DERIVATION OF THE
COSINE TIMING WAVES
FROM THE SUPPLY VOLTAGES

(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° resulting in commutation failure and probably a short circuit. To prevent the firing angle reaching 180°, 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
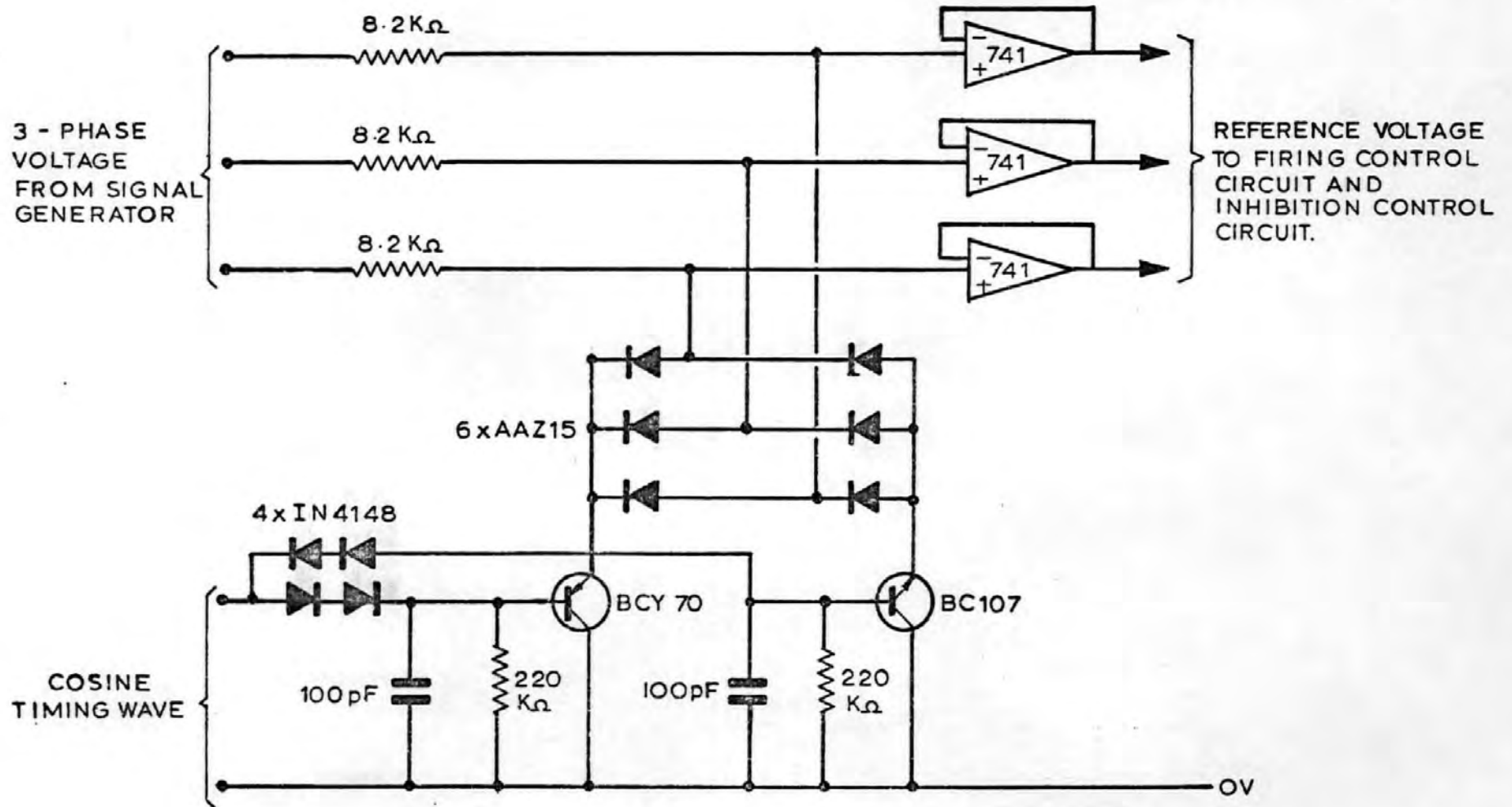
FIG. 5·5 END-STOP CONTROL OF FIRING ANGLE

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(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 wave-forms 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

10 k Ω

10 k Ω

1 M Ω

+12 V

741

−12 V

1 M Ω

0V

BAX 13

MONOSTABLE CIRCUIT

39 k

39 k

0.01 μF

0V

BISTABLE CIRCUIT

TO OUTPUT STAGE (−VE GROUP)

INTERCONNECTION WITH OTHER CONTROL CHANNELS.

NOR - GATES : 4 NOR 60, 2x2 INPUT + 2x3 INPUT, 24 v

INVERTER

AS ABOVE

TO OUTPUT STAGE (+VE GROUP)

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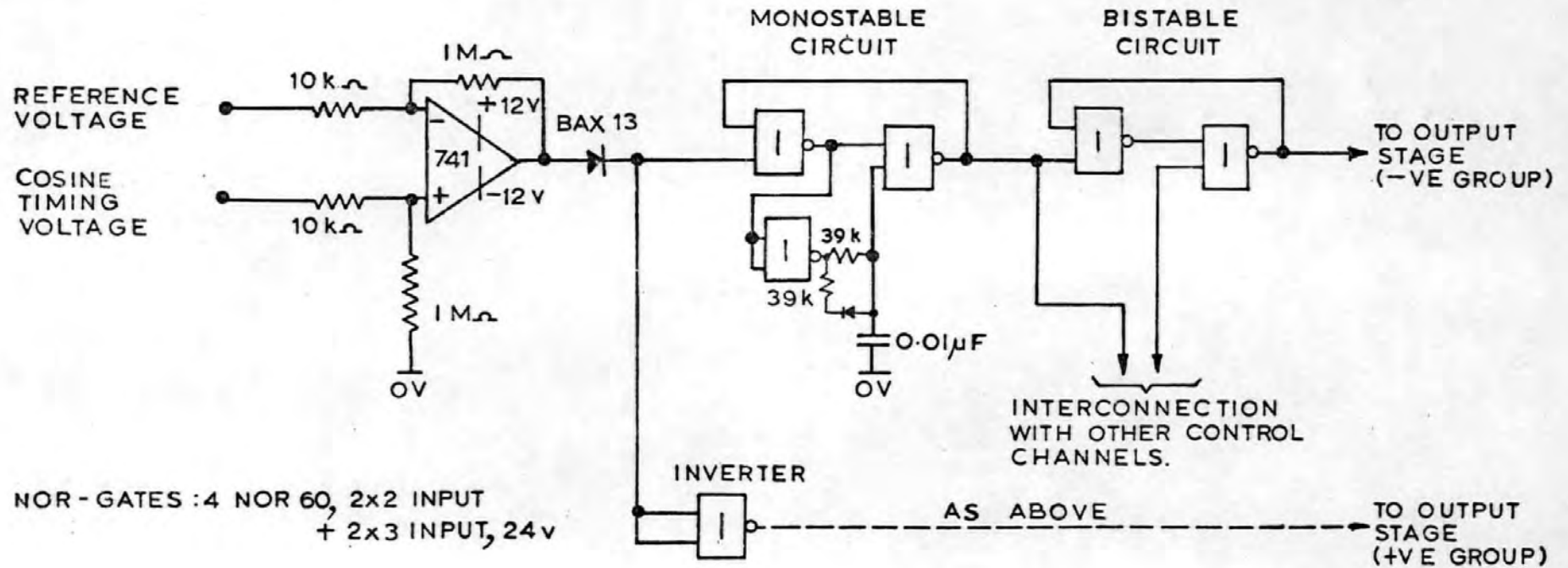
FIG. 5.6   PULSE - FORMING CIRCUIT FOR ONE THYRISTOR IN EACH GROUP
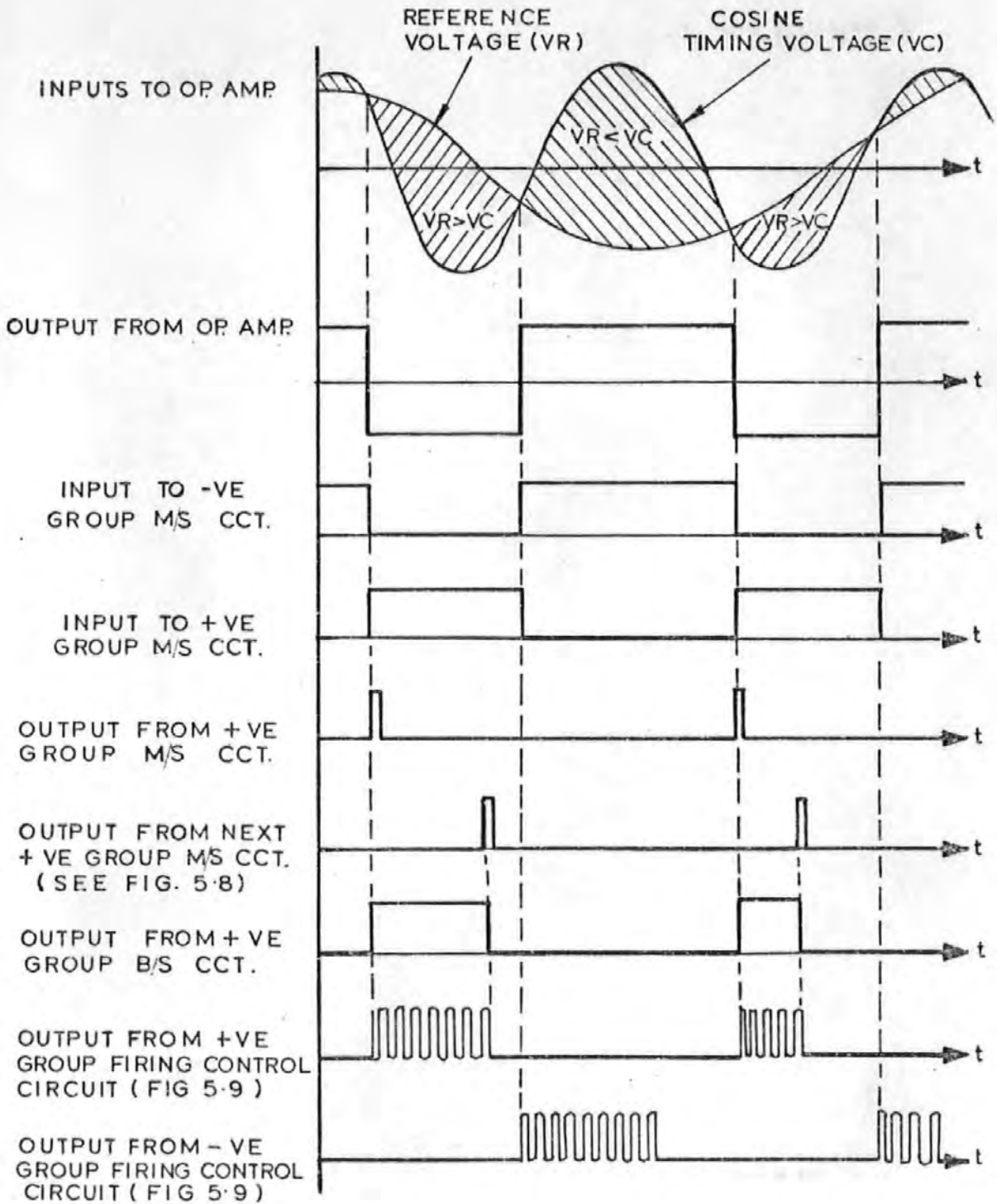
FIG. 5.7   WAVEFORMS FOR FIG 5.6

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 $0.01\mu F$ 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.
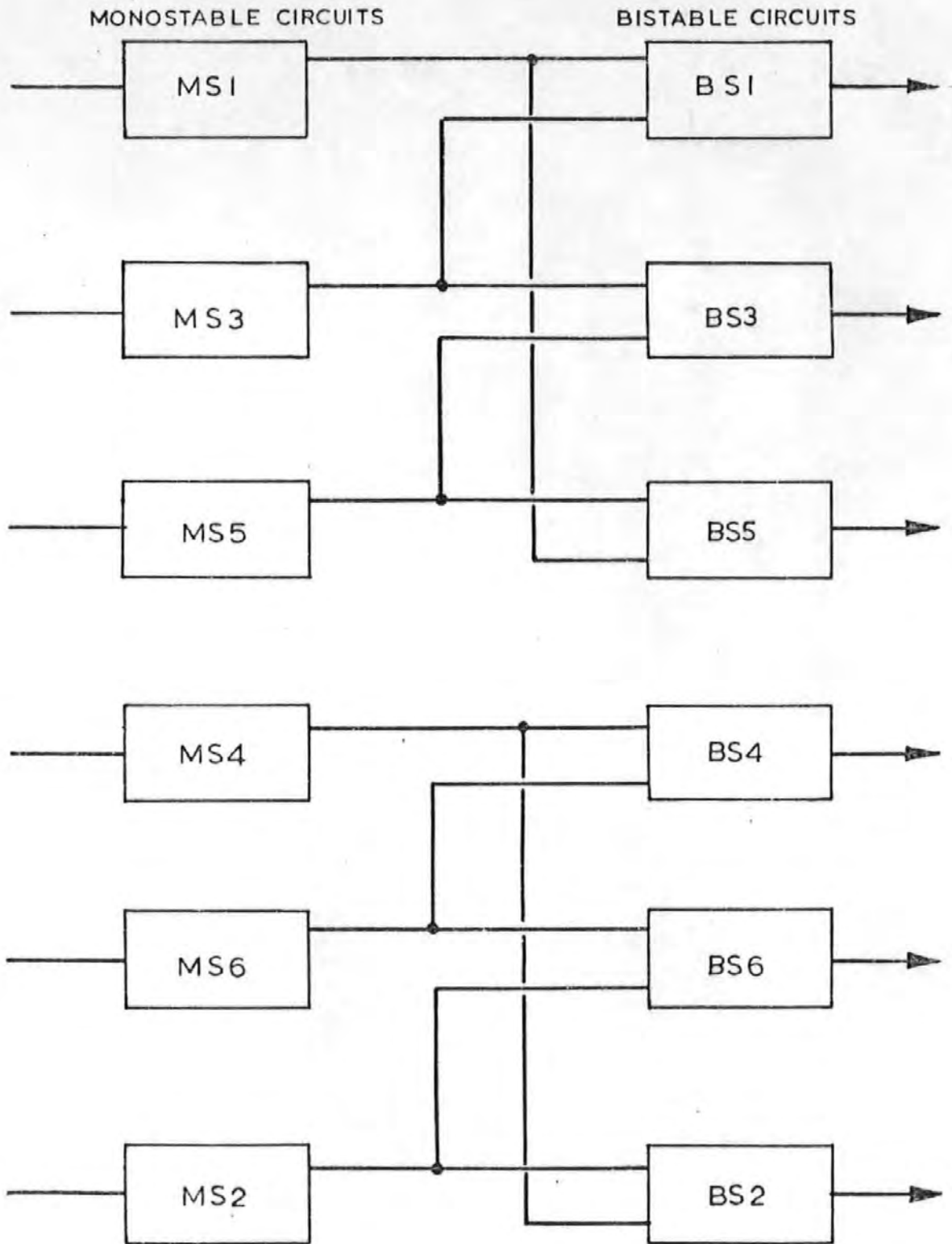
FIG 5·8   DIAGRAM SHOWING INTERCONNECTIONS BETWEEN
THE MONOSTABLE AND BISTABLE CIRCUITS OF THE
FIRING CONTROL CIRCUITS FOR EITHER THE
POSITIVE GROUP OR NEGATIVE GROUP.

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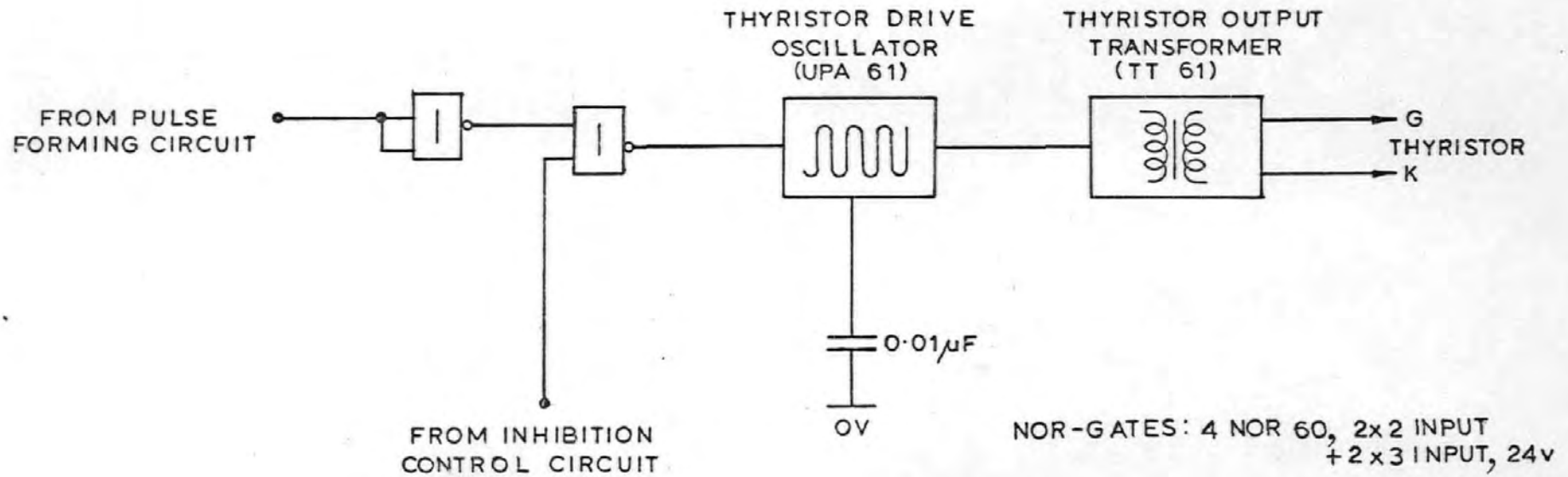
FIG 5.9   OUTPUT STAGE OF FIRING CONTROL CIRCUIT
FOR ONE THYRISTOR

(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
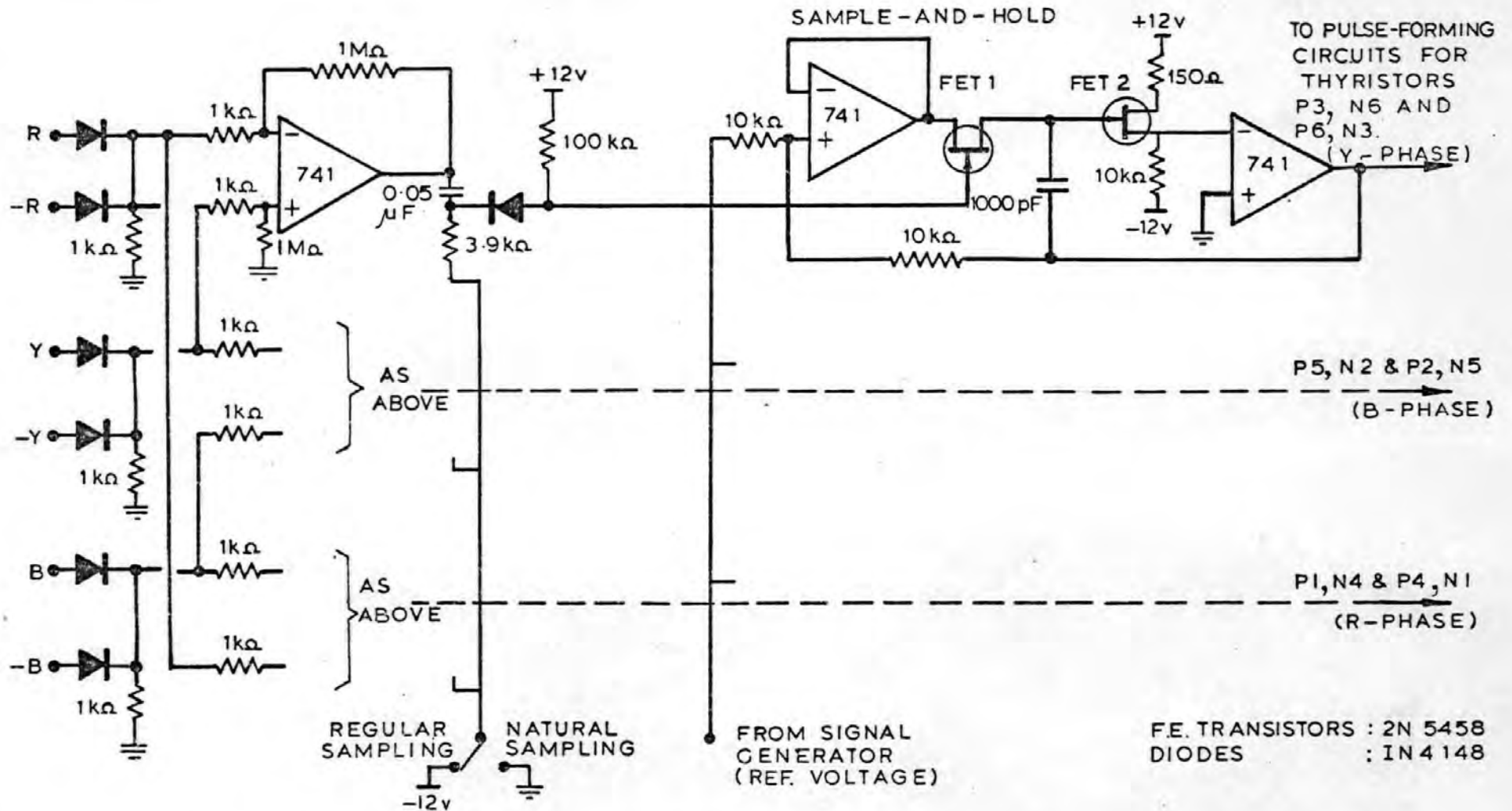
FIG. 5·10 CIRCUIT FOR SAMPLING THE REFERENCE VOLTAGE
(FOR ONE OUTPUT PHASE OF THE CYCLOCONVERTER)
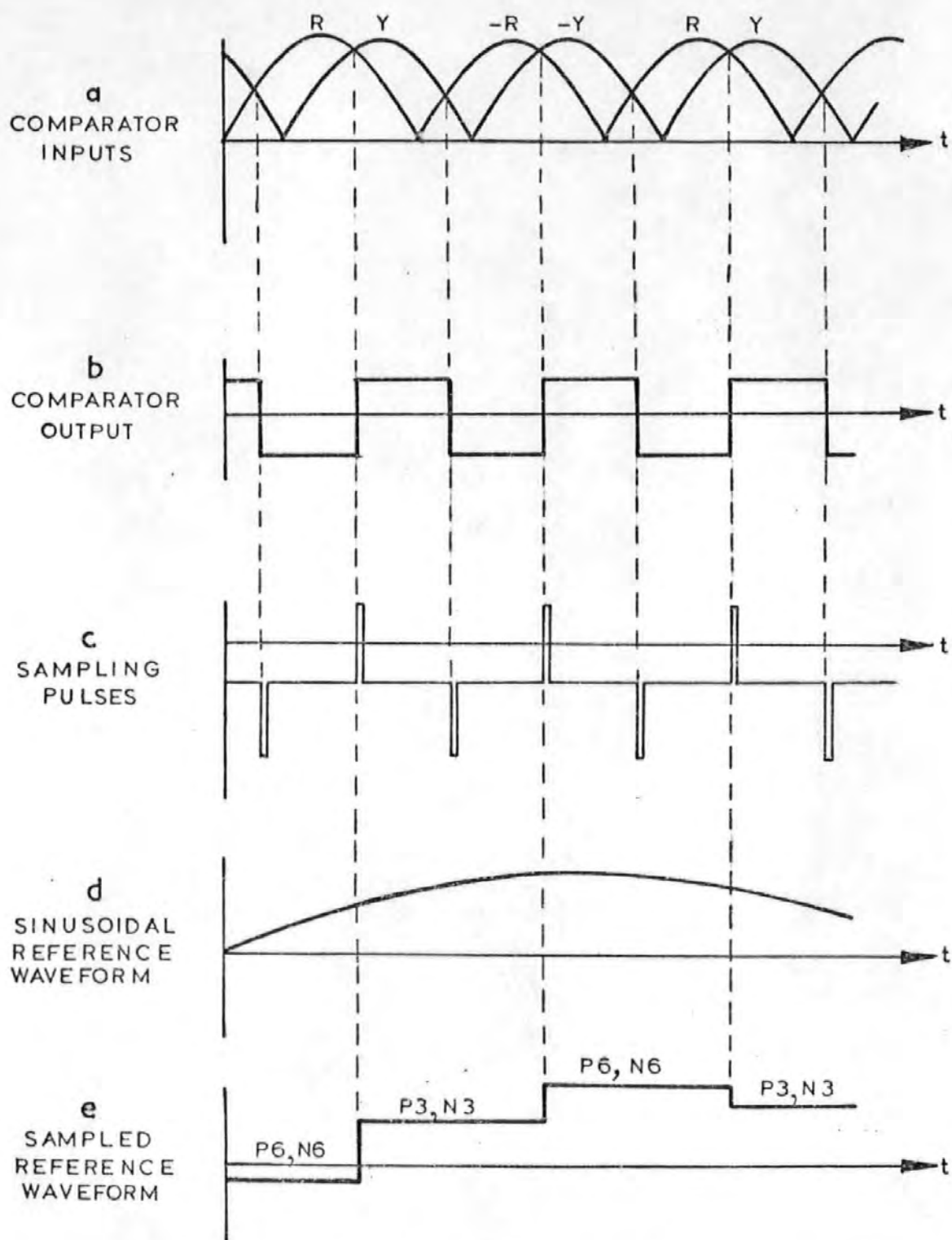
FIG. 5·II   WAVEFORMS FOR ONE CHANNEL
OF CIRCUIT SHOWN IN FIG. 5·10
WHEN SWITCHED FOR REGULAR SAMPLING

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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
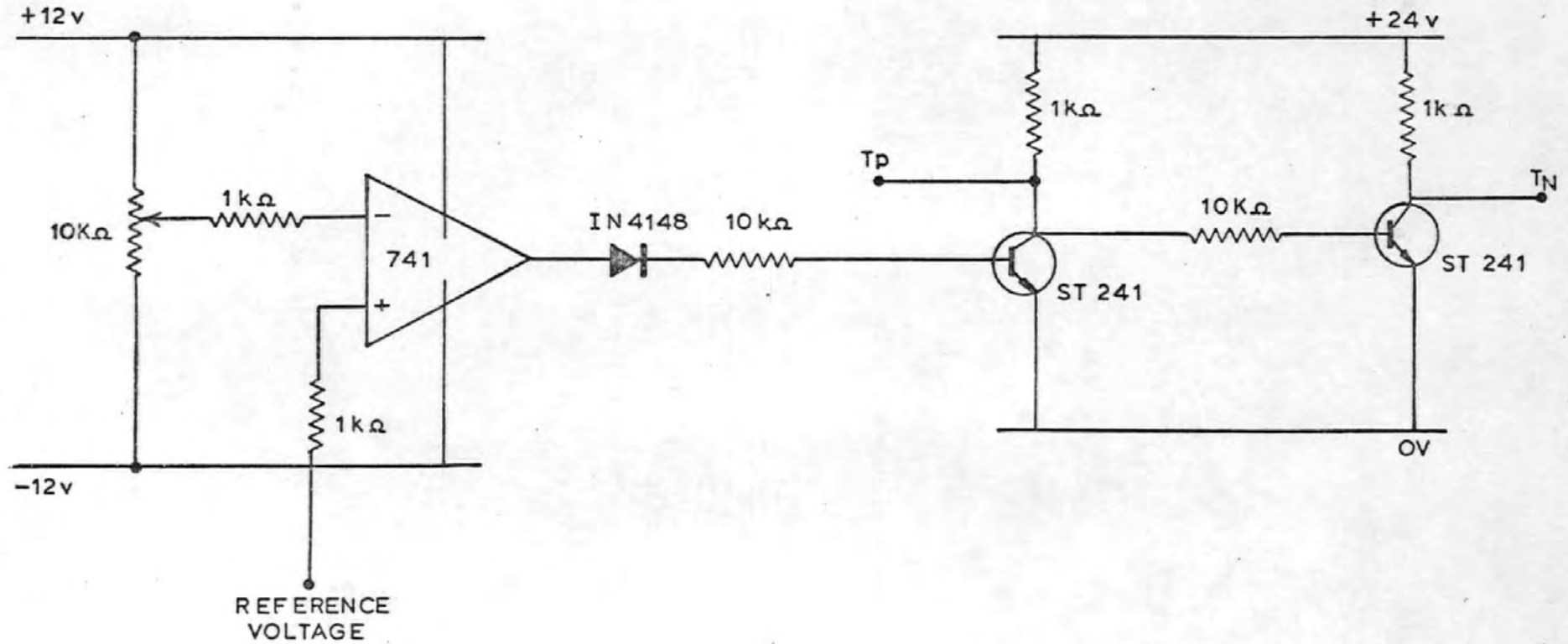
FIG. 5·12   INHIBITION CONTROL USING REFERENCE VOLTAGE

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REFERENCE
VOLTAGE

OUTPUT
FROM 741

+12v

0

−12v

OUTPUT T$_N$

24v

0

OUTPUT T$_P$

24v

0
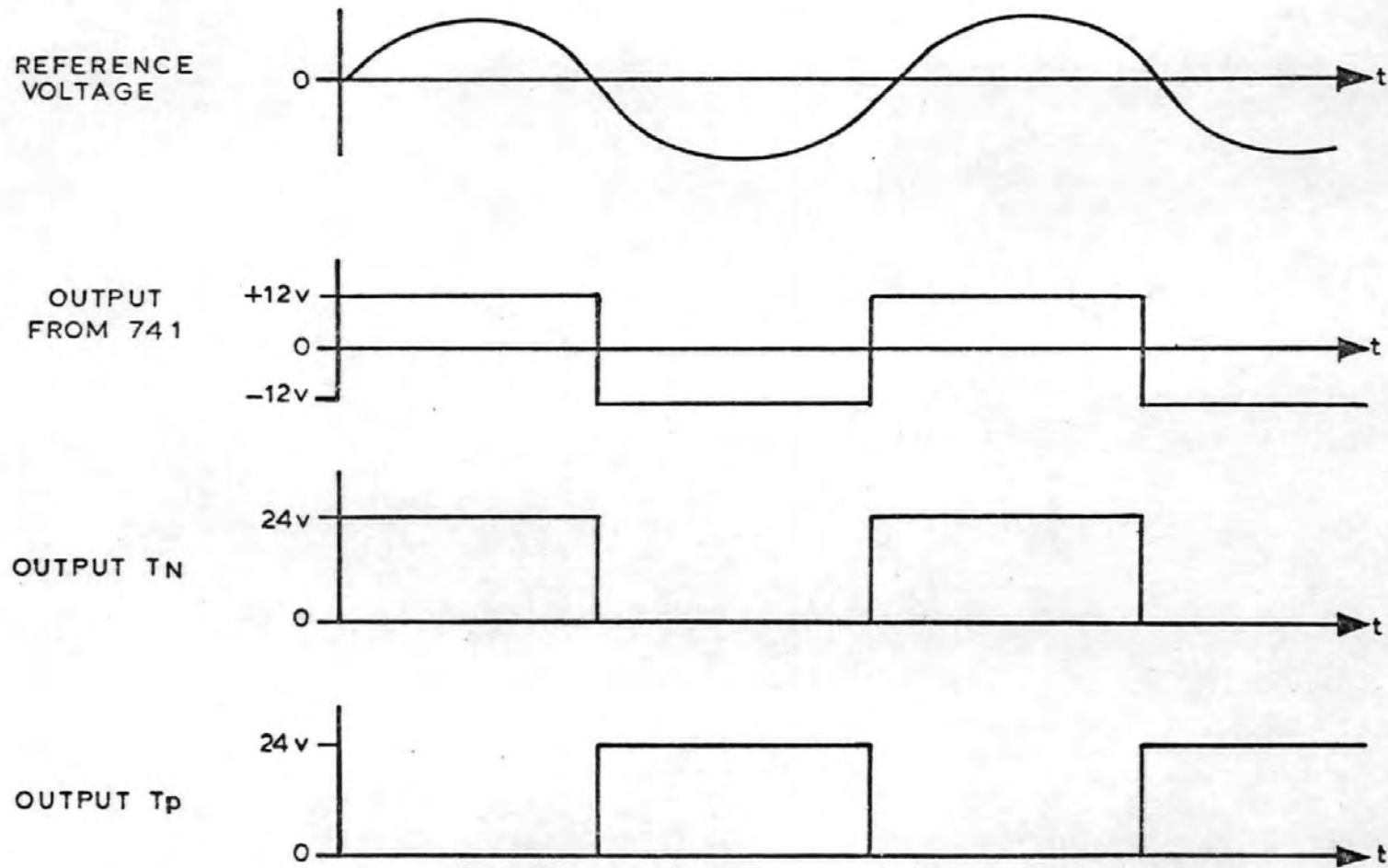
FIG 5·13   WAVEFORMS   FOR FIG 5·12

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.[8], Arikan[20]). However, d.c. current transformers require auxiliary supplies of frequency of about 2 kHz which can cause a delay of up to about $\frac{1}{4}$ 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[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-

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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
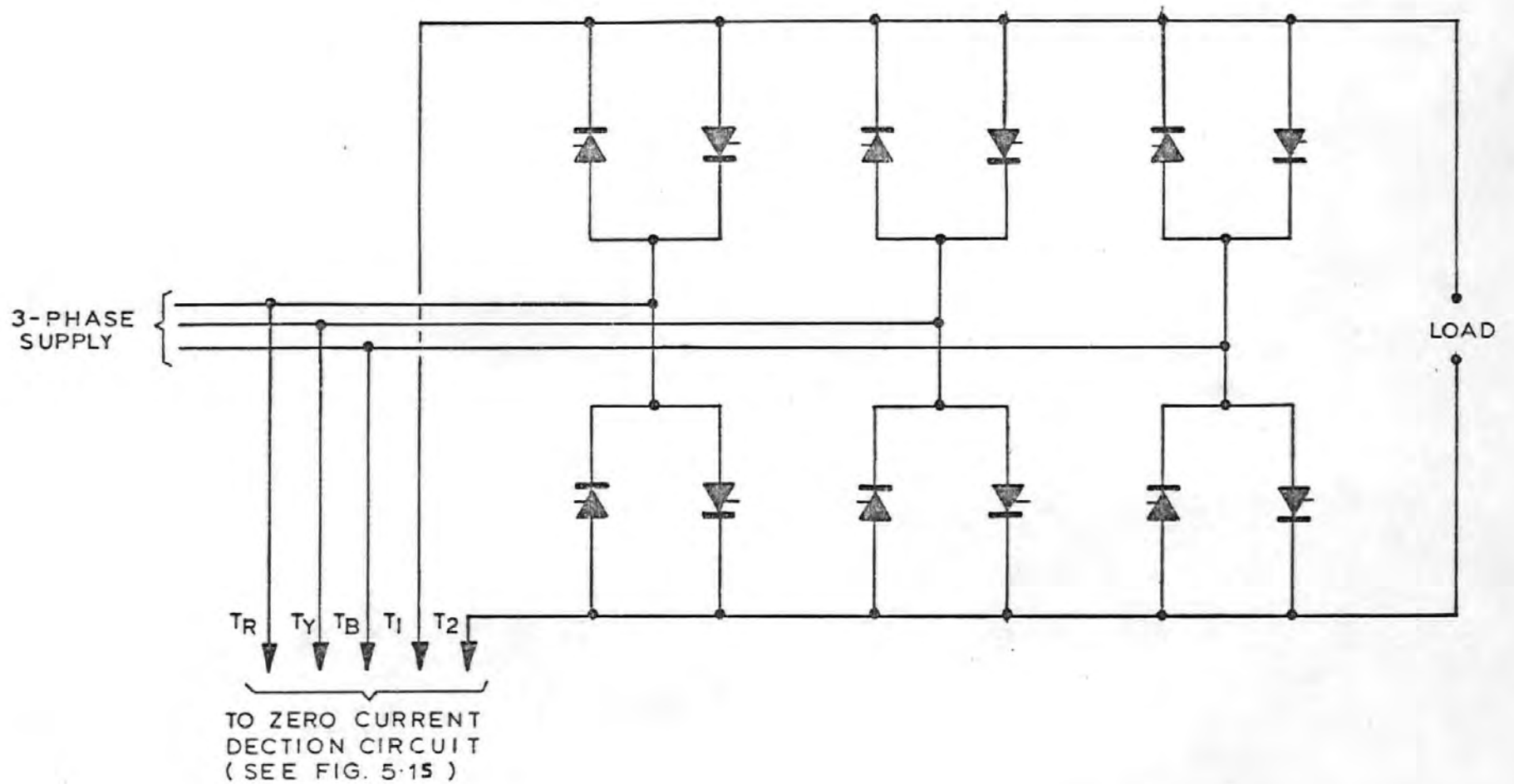
FIG. 5·14 POWER CIRCUIT CONNECTIONS FOR ZERO CURRENT
DETECTION

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 1 kΩ pre-set 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.
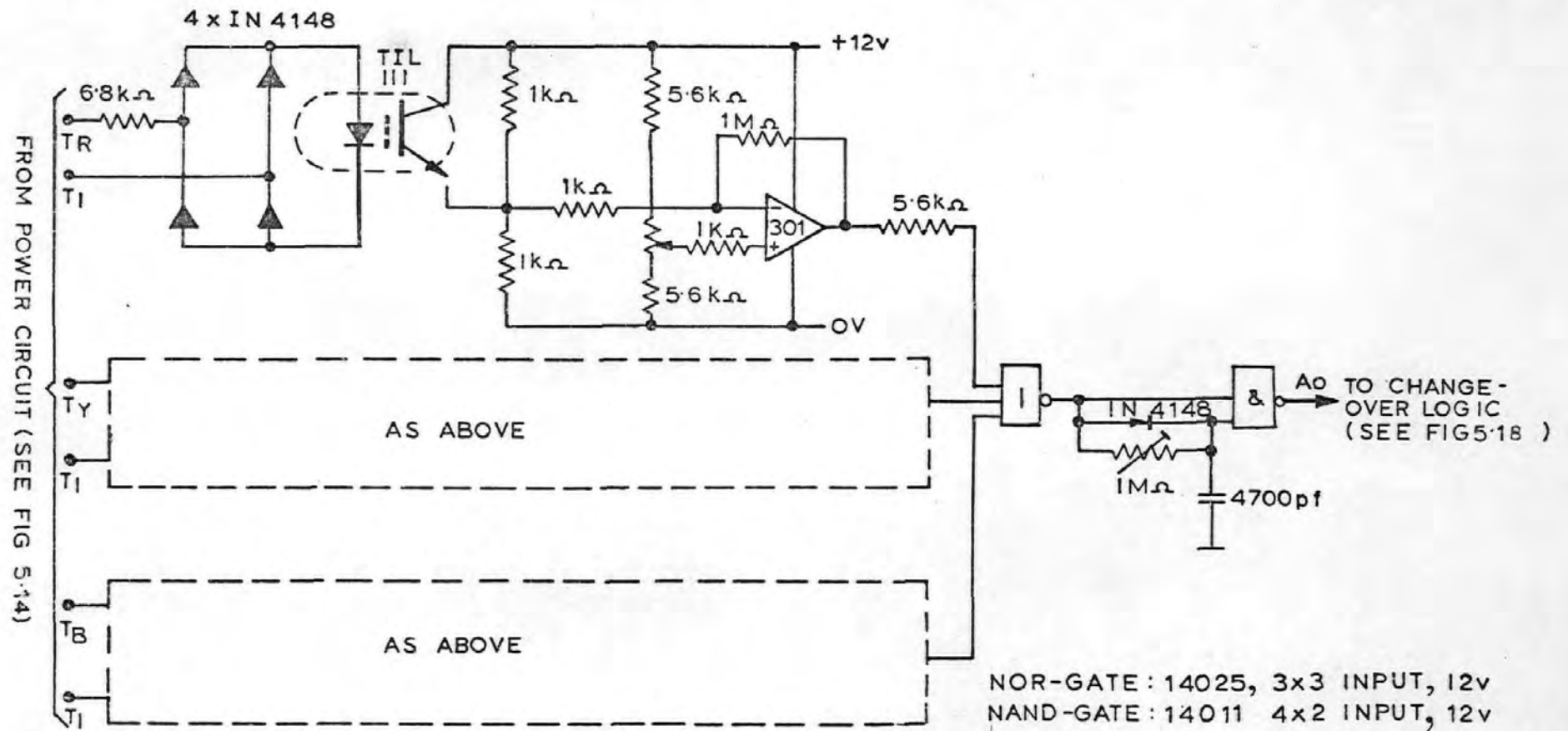
FIG. 5·15   ZERO CURRENT DETECTION CIRCUIT FOR INHIBITION CONTROL

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 Ⓐ as shown in figure 5.17.

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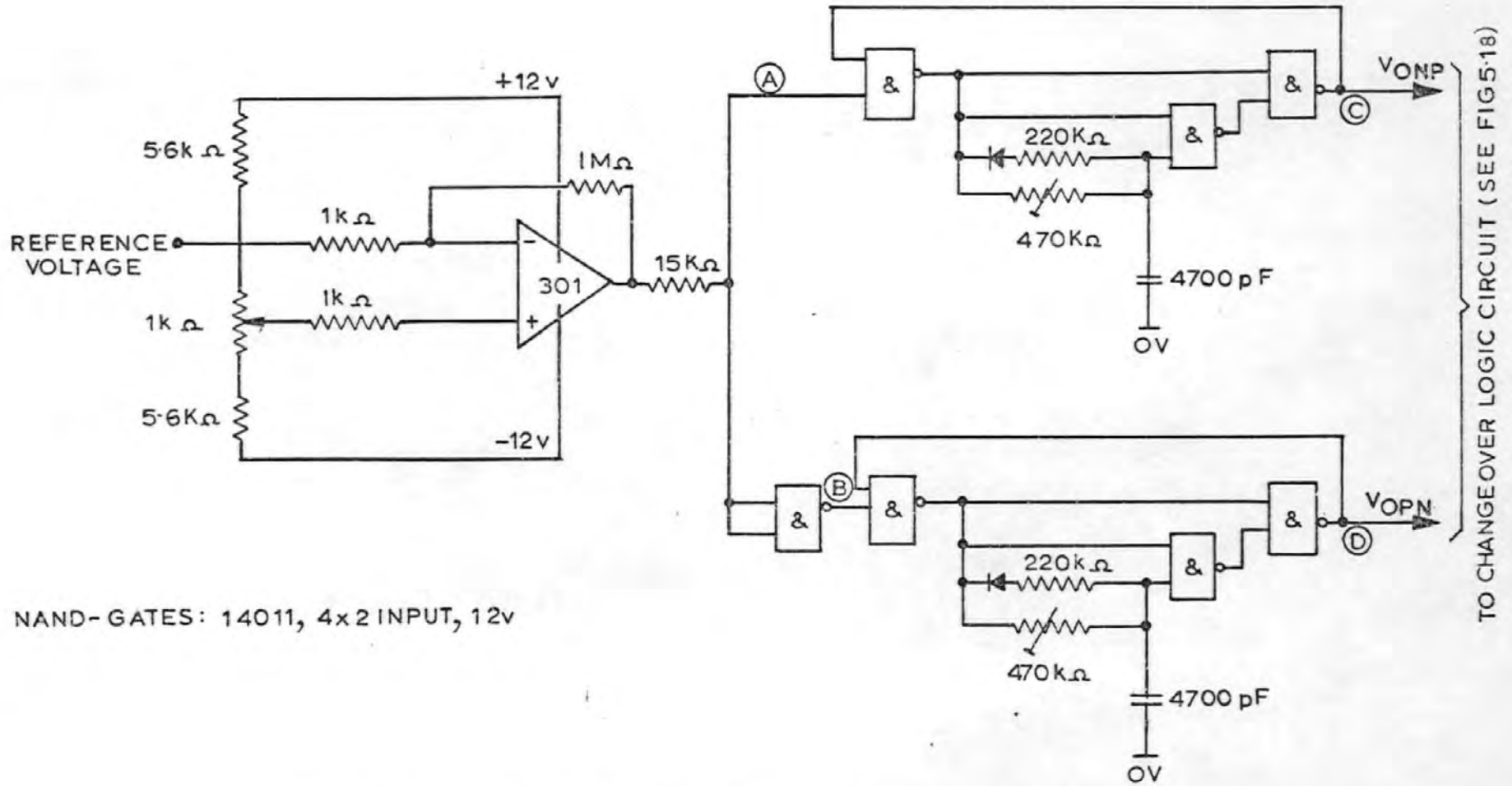
FIG. 5·16 ZERO REFERENCE VOLTAGE DETECTION
CIRCUIT FOR INHIBITION CONTROL

REFERENCE
VOLTAGE

t

A

t

B

t
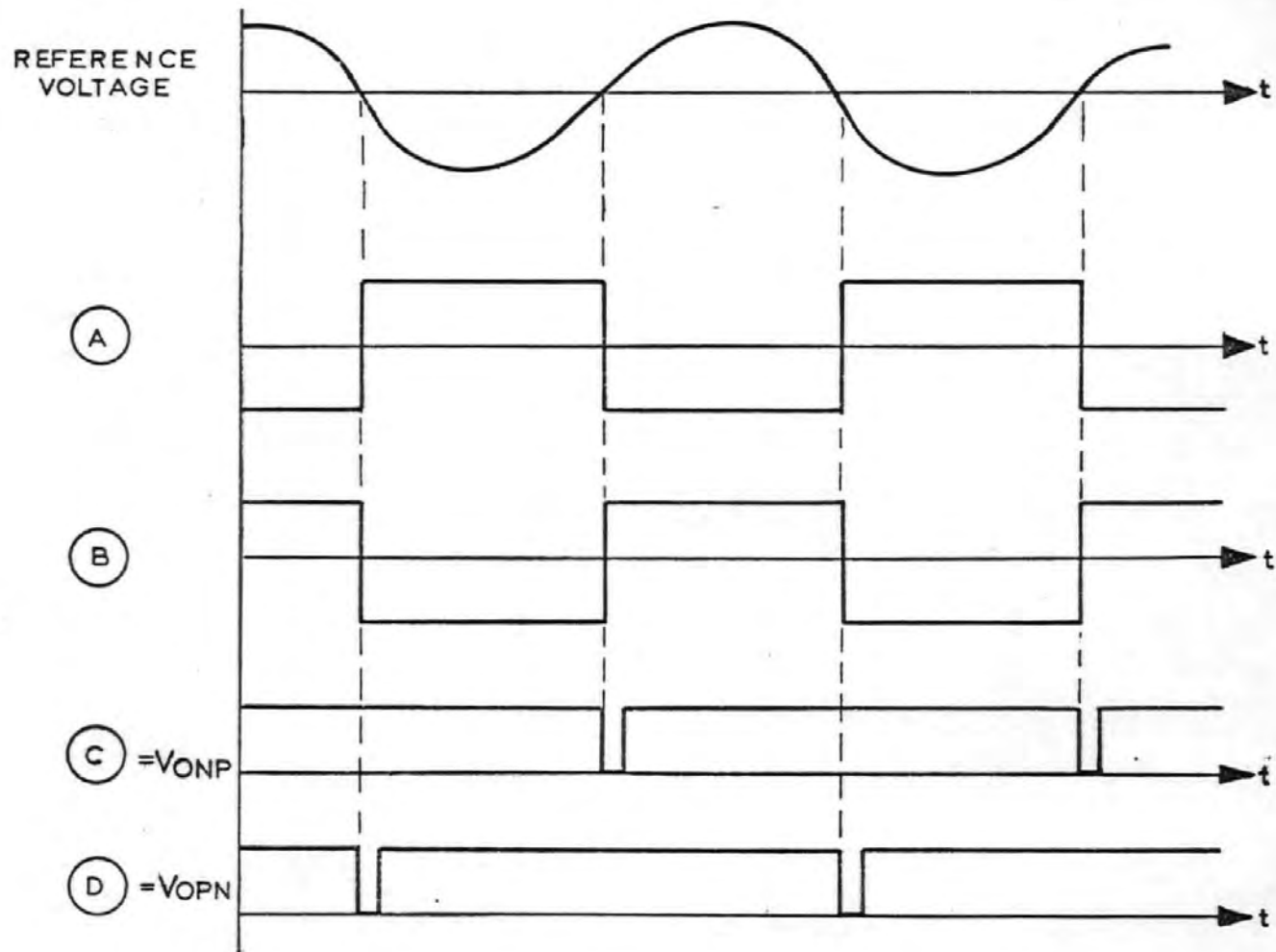
LETTERS REFER
TO FIG. 5·16

C =V$_{ONP}$

t

D =V$_{OPN}$

t

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

180
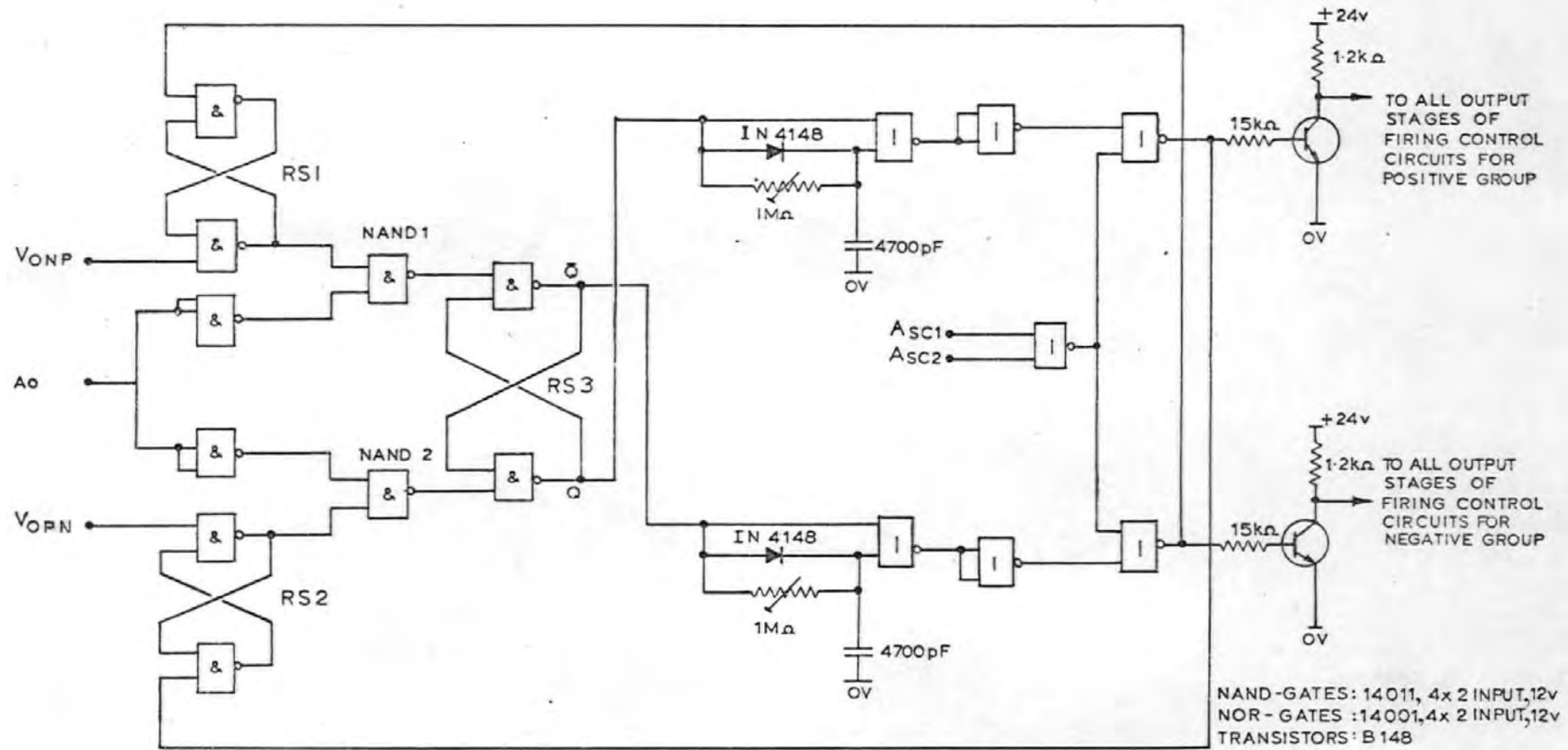
FIG 5·18 CHANGEOVER LOGIC CIRCUIT FOR INHIBITION CONTROL

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_o 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.
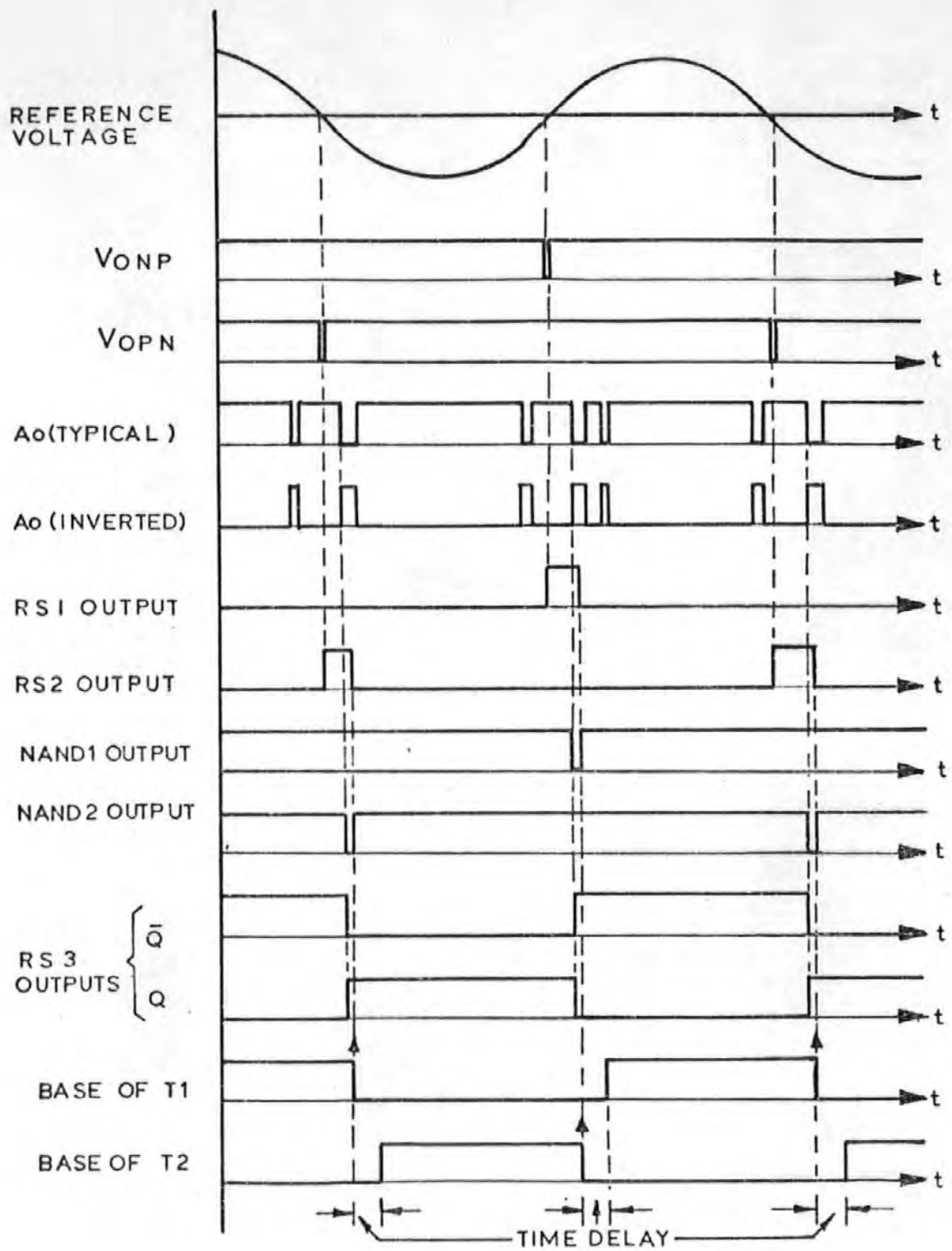
FIG 5·19   WAVEFORMS FOR FIG. 5·18

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.

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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 B148 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 K134A 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 pf_i/f_o 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_o sub-harmonic component is attenuated by employing regular in place of natural sampling when the cycloconverter is operated at a pf_i/f_o Ratio of 4.5. However, at a pf_i/f_o Ratio of 5.5, the magnitude of the 0.5 f_o 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.

SUB-HARMONIC COMPONENT OF FREQUENCY $0.5f_o$

SUPER-HARMONIC COMPONENT OF FREQUENCY $1.5f_o$
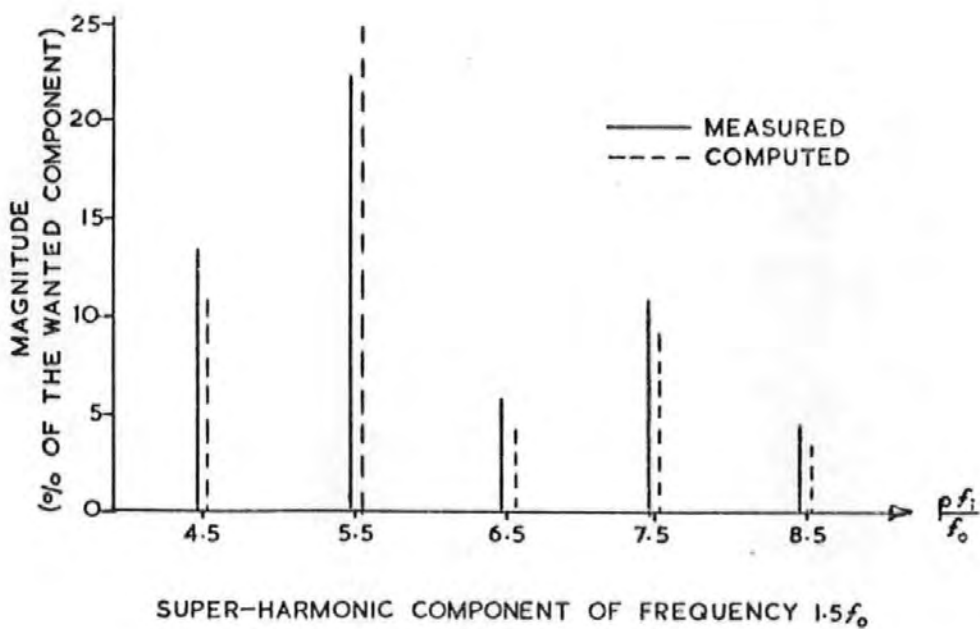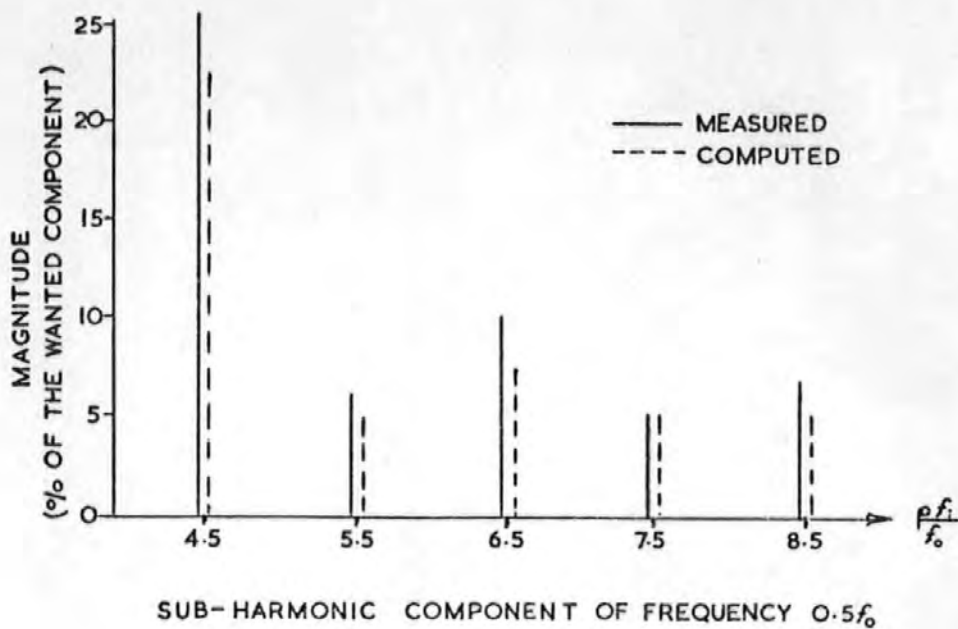
FIG 5·20   COMPARISON OF EXPERIMENTAL
AND COMPUTER RESULTS:
NATURAL SAMPLING

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SUB-HARMONIC COMPONENT OF FREQUENCY 0.5f_o

SUPER-HARMONIC COMPONENT OF FREQUENCY 1.5f_o
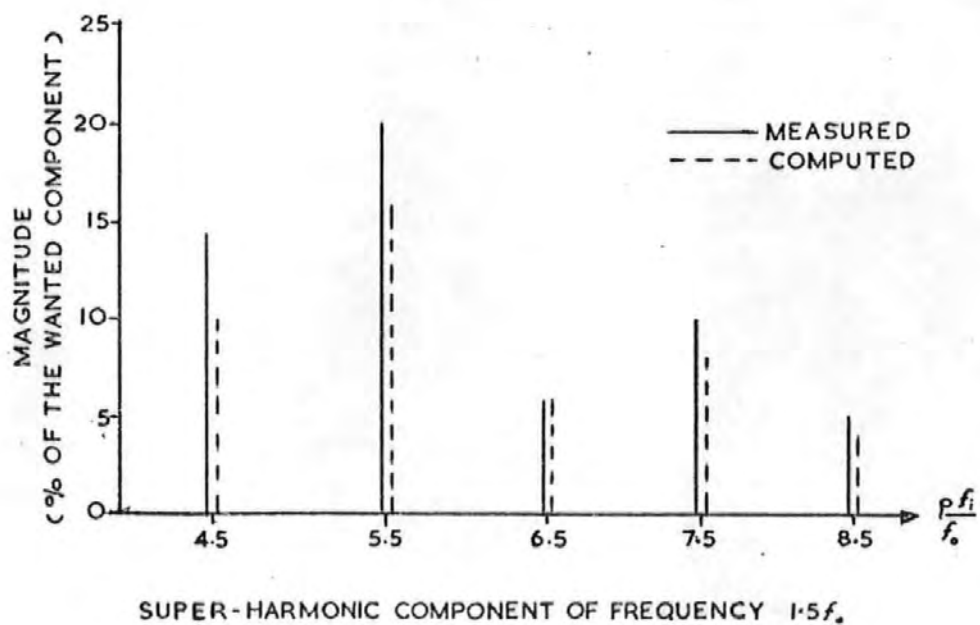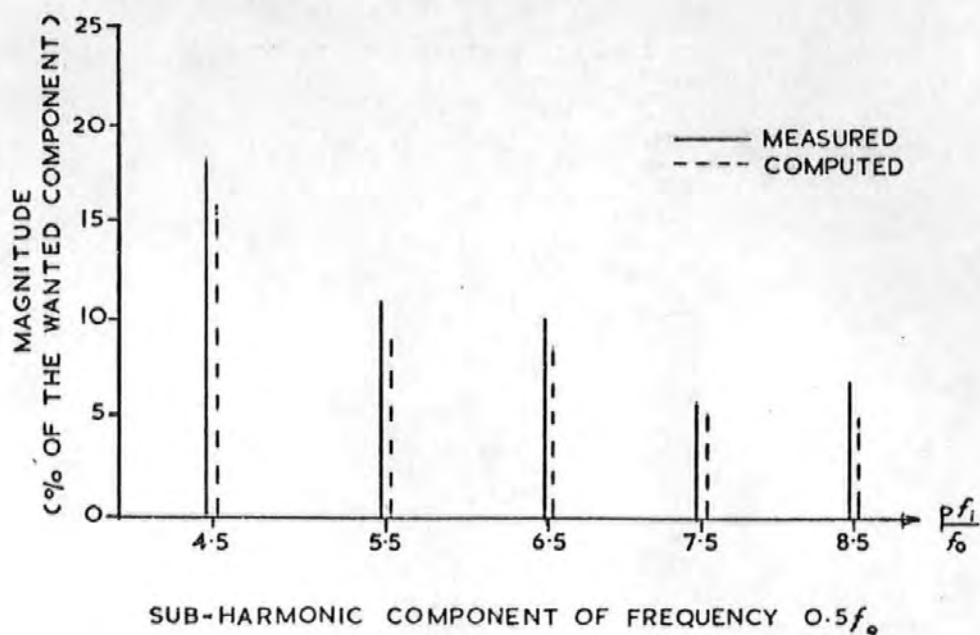
FIG 5·21   COMPARISON OF EXPERIMENTAL
AND COMPUTER RESULTS:
REGULAR SAMPLING

(5.5.3.2) Operation at Integer Values of pf_i/f_o Ratio
Output frequency = $16\frac{2}{3}$ Hz and 50 Hz

The waveforms for output frequencies of $16\frac{2}{3}$ Hz
and 50 Hz are shown in figures 5.22 and 5.23 respectively.
These are equivalent to pf_i/f_o 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 pf_i/f_o Ratio
Output Frequency = 17.6 Hz

The waveforms for an output frequency of 17.6 Hz,
corresponding to a pf_i/f_o 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.
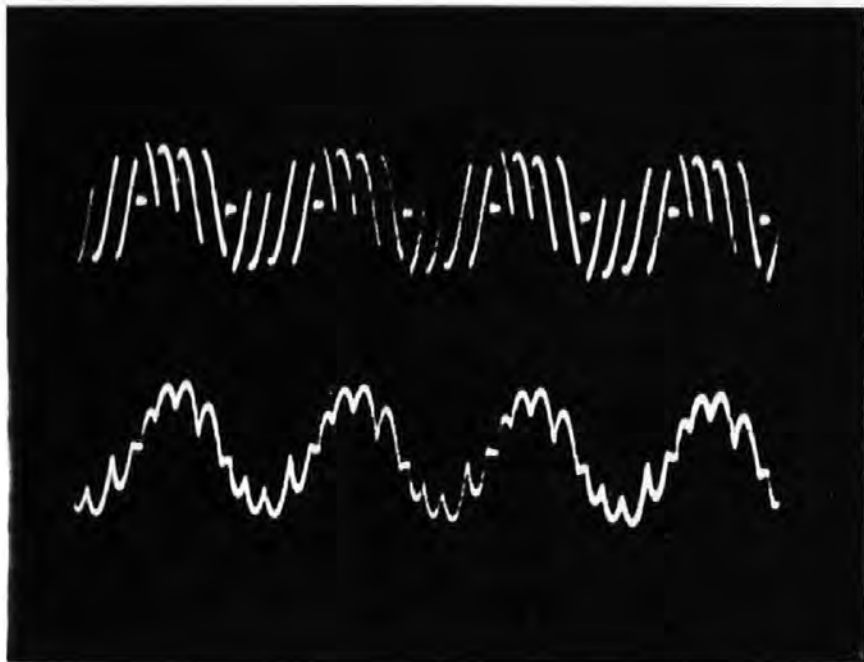
FIG 5·22 $f_o = 16\frac{2}{3}$ HZ

FIG 5·23 $f_o = 50·0$ HZ
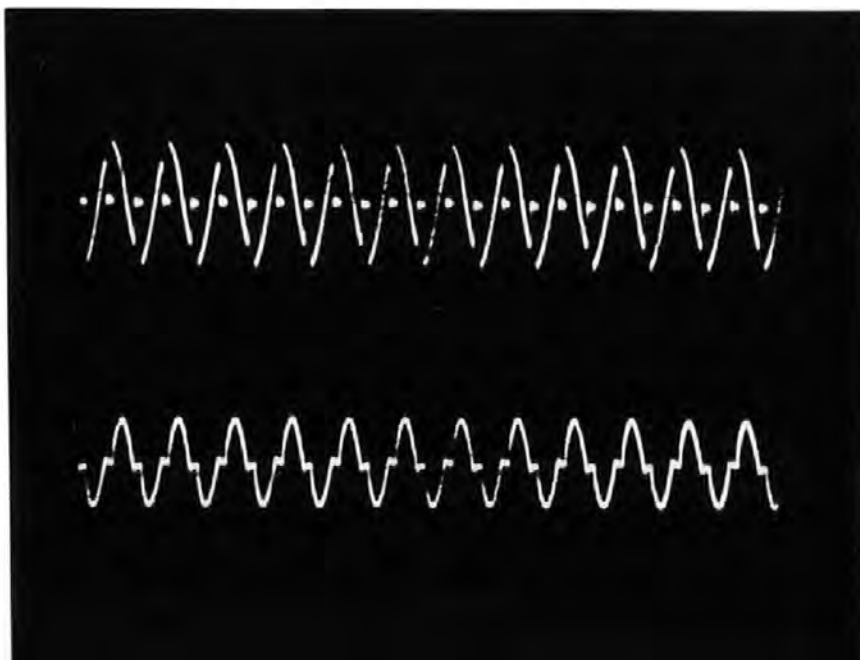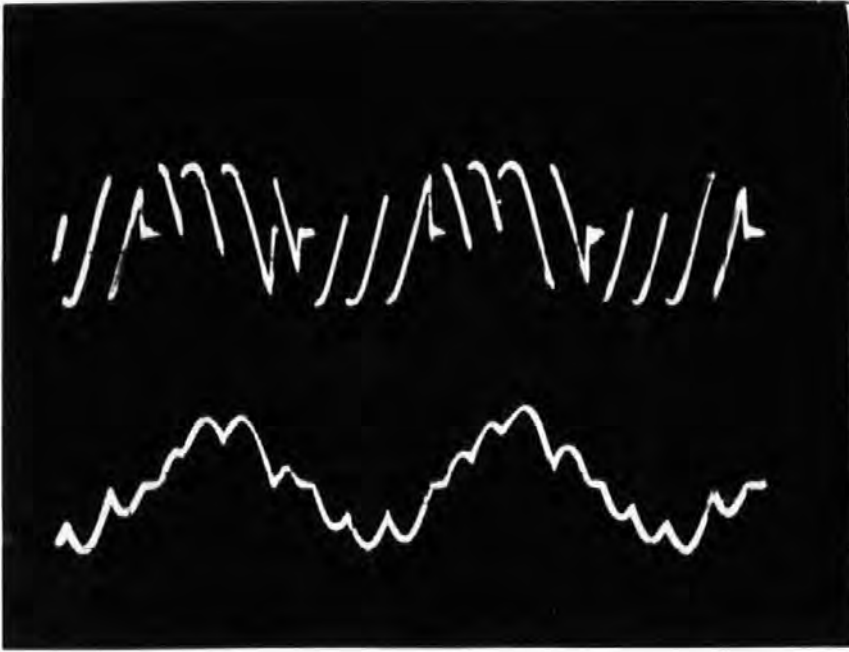
FIG 5·24 f_o = 17·6 HZ

FIG 5·25 f_o = 38·0 HZ
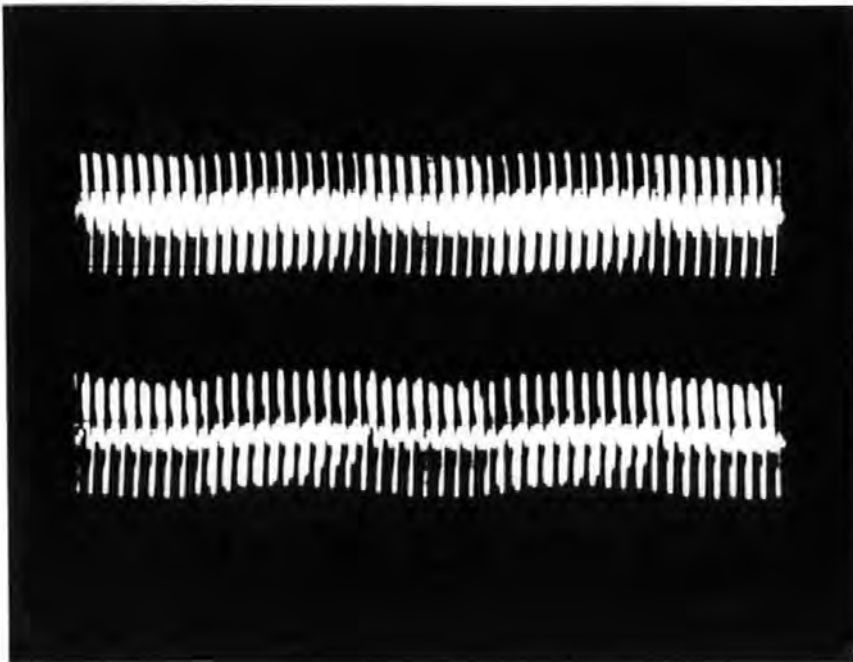
(5.5.3.4) Effect of a Sub-harmonic Component having a
Frequency considerably less than the Wanted
Output Frequency

The computed results in table 4.1a in chapter
4 showed that for operation of the cycloconverter at a
pf_i/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 intrapolation between these results, there is a high
magnitude sub-harmonic component of frequency less than
10% of the output frequency when the pf_i/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_i/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 pf_i/f_o 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 pf_i/f_o Ratio of 4.0), the waveforms are as shown in
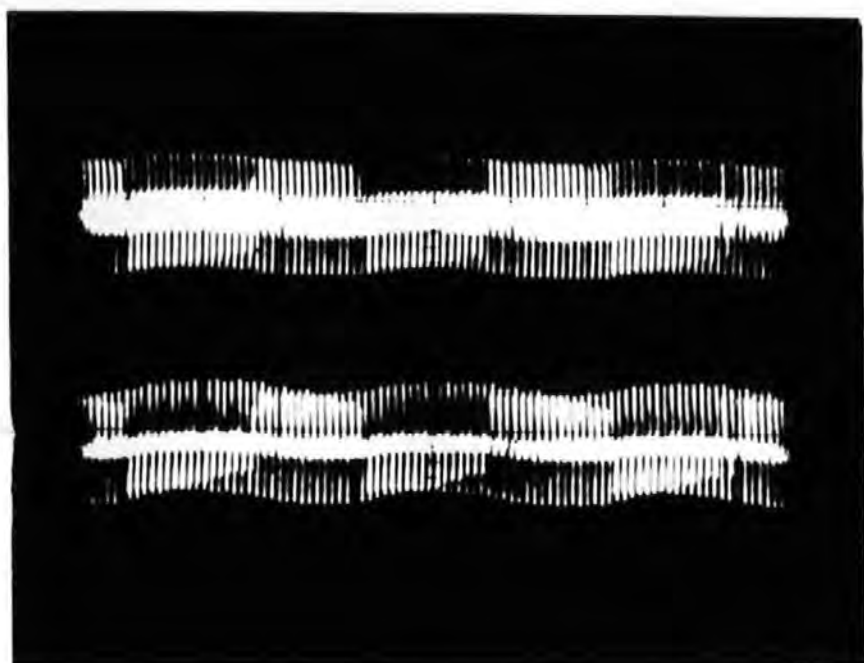
FIG 5·26 $f_o = 37.75$ HZ

FIG 5·27 $f_o = 37.5$ HZ
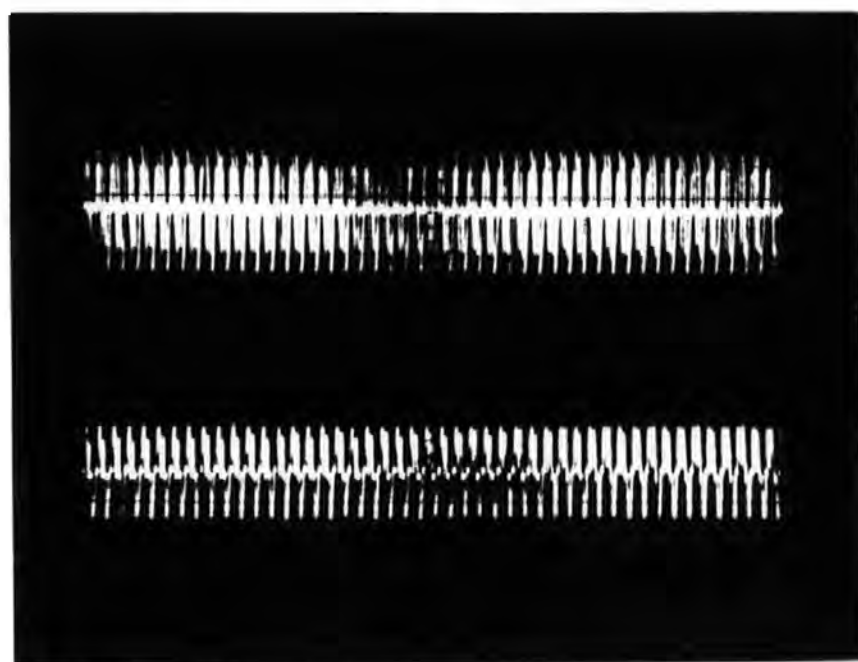
figure 5.27. The variation in nett area enclosed by the.

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 (pf_i/f_o 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 pf_i/f_o 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 pf_i/f_o 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 (pf_i/f_o Ratio =5.0)
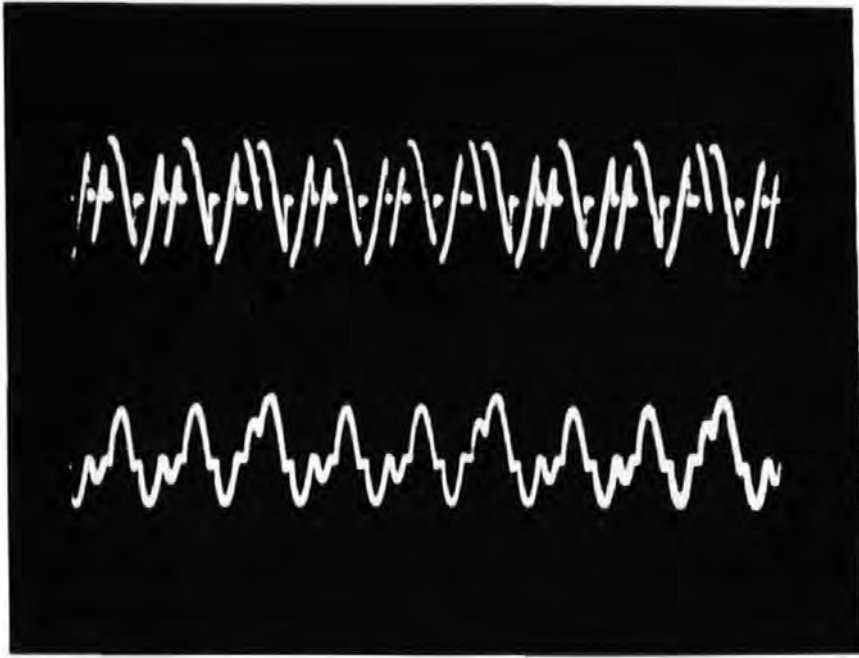
FIG 5.28 $f_0 = 35.0$ HZ

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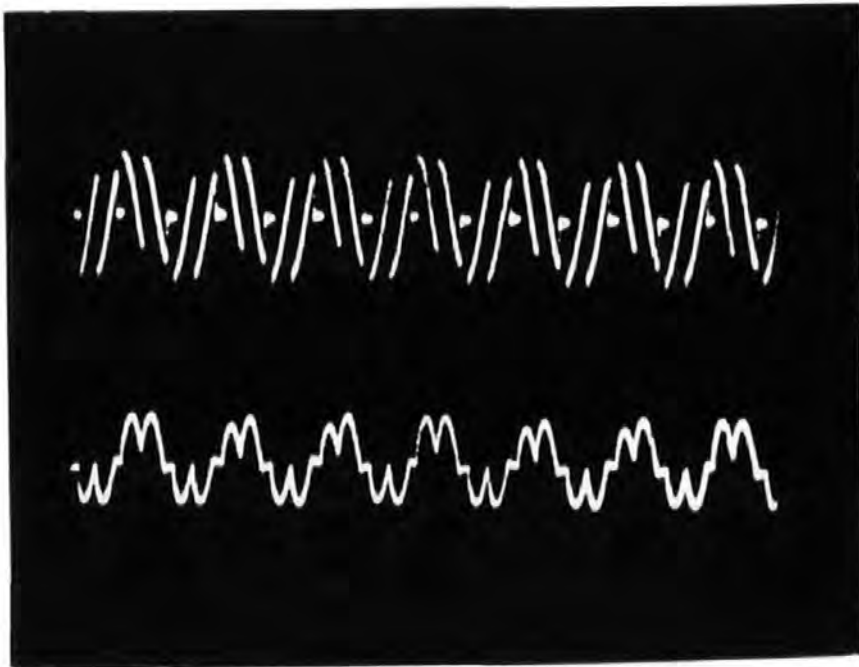
to 31.0 Hz (pf_i/f_o Ratio = 4.84) are shown in figures 5.29 - 5.31 and are discussed in detail below. A small decrease in output frequency gave similar waveforms.

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 Hz - 4 Hz = 27 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
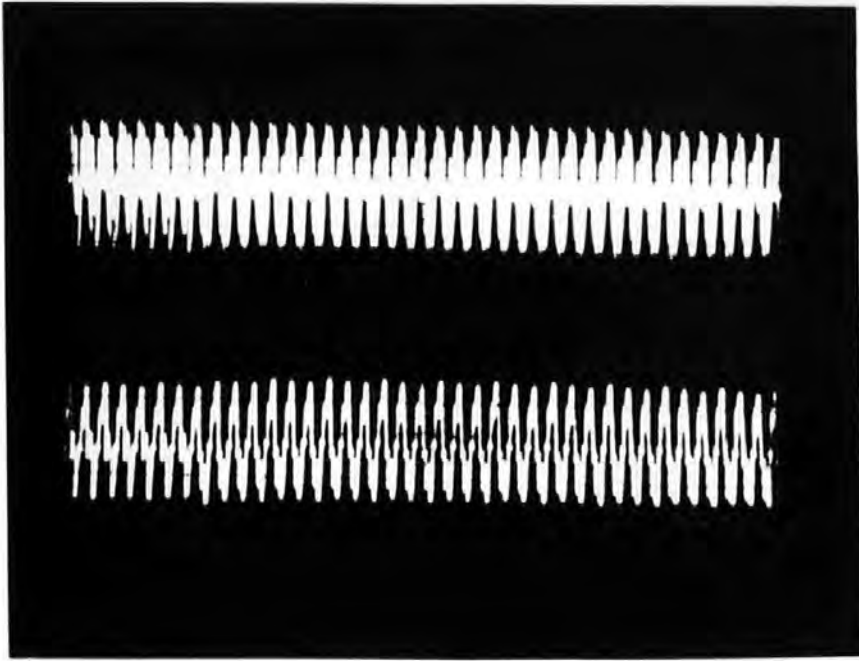
FIG 5·30 f_o = 30·0 HZ

FIG 5·31 f_o = 31·0 HZ
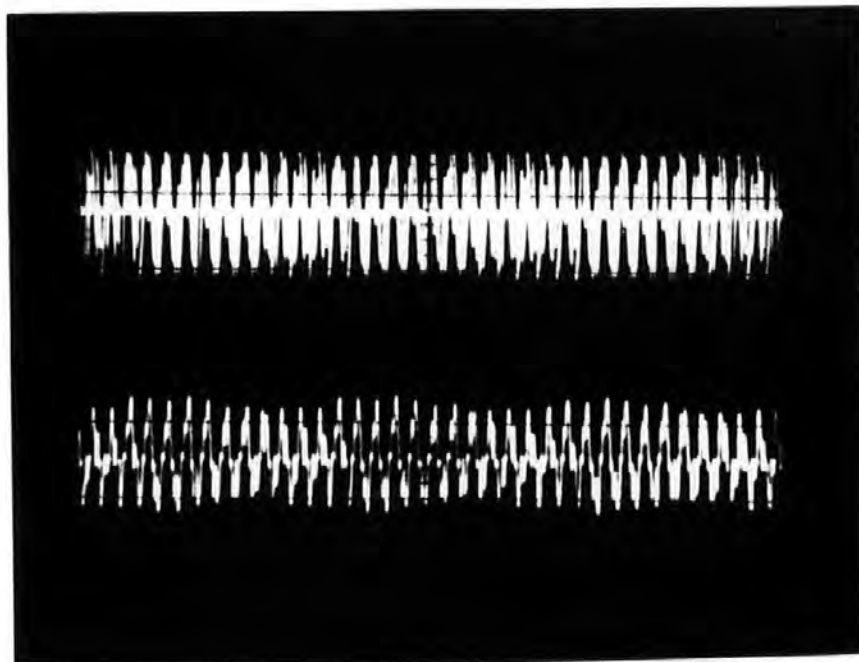
the output frequency is then present.

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 pf_i/f_o Ratio, and that the general harmonic is a sub-harmonic component over a limited range of the pf_i/f_o Ratio values; for even integer values of the pf_i/f_o 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 pf_i/f_o 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.

ACKNOWLEDGEMENTS

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APPENDIX 1

COMPUTER PROGRAM FOR THE

CIRCULATING-CURRENT CYCLOCONVERTER

```
C ----------------------------------------------------
C
C         THYRISTOR CYCLOCONVERTER WITH SINGLE PHASE OUTPUT
C             CIRCULATING CURRENT MODE
C             COSINUSOIDAL CONTROL (NATURAL AND REGULAR SAMPLING)
C             RESISTIVE LOAD
C
C          REQUIRED INPUT DATA
C          NP=PULSE NUMBER OF INPUT (=2 OR 3 IN THIS PROGRAM)
C          FI=INPUT FREQUENCY (HZ)
C          FO=OUTPUT FREQUENCY (HZ)
C          FF=FREQUENCY OF FUNDAMENTAL(LOWEST FREQUENCY IN OUTPUT)
C              FO/FF MUST EQUAL AN INTEGER
C              NP*FI/FF MUST EQUAL AN INTEGER
C          NH=HIGHEST ORDER OF HARMONIC IN THE OUTPUT
C          DM=MODULATION DEPTH
C
C          ADDITIONAL SYMBOLS USED IN THIS PROGRAM
C          N=ORDER OF HARMONIC,REFERRED TO FF
C          I=COUNTER FOR SAMPLING
C          IC=COUNTER FOR COMPUTATION FOR NATURAL SAMPLING FOLLOWED
C                    BY REGULAR SAMPLING
C          KP=COMMUTATION POINT (=1 AT FIRST COMMUTATION AFTER T=0)
C          NKP=TOTAL NUMBER OF COMMUTATION POINTS IN THE COMPUTATION
C          TIME=INSTANT AT WHICH COMPUTATION IS TO BE CARRIED OUT
C              TIME=0.0 SECS AT START OF SINUSOIDAL VRONTROL VOLTAGE
C          TINC=TIME OF START OF VC
C          T(KP)=TIME OF COMMUTATION KP
C          DT=INCREMENTAL TIME INTERVAL FOR COMPUTATION
C          VR=REFERENCE VOLTAGE
C          VC=COSINUSOIDAL CONTROL VOLTAGE
C          L=2 FOR POSITIVE GROUP
C           =1 FOR NEGATIVE GROUP
C          A(N)=FOURIER COEFFICIENT FOR COSINE TERMS
C          B(N)=FOURIER COEFFICIENT FOR SINE TERMS
C          C(N)=FOURIER COEFFICIENT FOR COSINE TERMS OF POSITIVE GROUP
C          D(N)=DITTO FOR SINE TERMS
C          E(N)=FOURIER COEFFICIENT FOR COSINE TERMS OF NEGATIVE GROUP
C          F(N)=DITTO FOR SINE TERMS
C          H(N)=MAGNITUDE(PEAK,P.U.) OF THE NTH HARMONIC
C          WD=DIFFERENCE IN ANGULAR FREQUENCY BETWEEN FI AND HARMONIC
C          WS=SUM OF DITTO
C             1P.U.=PEAK OF INPUT VOLTAGE
C
          DIMENSION T(100),A(100),B(100),C(100),D(100),E(100),F(100)
          DIMENSION HN(100),HP(100),H(100),ANUP(100),ANGN(100),ANGC(100)
          READ(5,199)NZ
      199 FORMAT(I5)
          DO 90 JK=1,NZ
          READ(5,200)FI,FO,FF,NH,NP,DM
      200 FORMAT(3F10.0,2I10,F10.0)
C
          DO 91 IC=1,2
          WRITE(6,201)
      201 FORMAT('1',5X,'INPUT DATA')
```

```
      WRITE(6,205)NP
  205 FORMAT('0','NUMBER OF PULSES=',I3)
      WRITE(6,202)FI
  202 FORMAT(' ','INPUT FREQUENCY=',F10.5,'HZ')
      WRITE(6,203)FO
  203 FORMAT(' ','OUTPUT FREQUENCY=',F10.5,'HZ')
      WRITE(6,204)FF
  204 FORMAT(' ','FUNDAMENTAL FREQUENCY=',F10.5,'HZ')
      WRITE(6,207)NH
  207 FORMAT(' ','HIGHEST ORDER OF HARMONIC=',I3)
      WRITE(6,206)DM
  206 FORMAT(' ','MODULATION DEPTH=',F10.5,'P.U.')
      IF(IC-1)96,96,97
C     NATURAL SAMPLING
   96 WRITE(6,209)
  209 FORMAT('0',5X,'OUTPUT DATA FOR CIRCULATING CURRENT CYCLO, NATURAL
     2 SAMPLING')
      GO TO 98
C     REGULAR SAMPLING
   97 WRITE(6,208)
  208 FORMAT('0',5X,'OUTPUT DATA FOR CIRCULATING CURRENT CYCLO, REGULAR
     2 SAMPLING')
   98 WRITE(6,210)
  210 FORMAT('0',5X,'KP',8X,'TIME(SECS)')
      WRITE(6,215)
  215 FORMAT(' ','NEGATIVE GROUP')
C
      PI=22.0/7.0
      DT=1.0/300.0/FI
      NKP=IFIX(NP*FI/FF+0.01)
      NKP1=NKP+1
C
C     COMMUTATION POINTS
      DO 5 L=1,2
      DO 10 KP=1,NKP1
      TINC=(KP-L+1-(2-L)*NP/2.0)/FLOAT(NP)/FI
      DO 20 I=1,151
      TIME=(I-1)*DT+TINC
      IF(IC-1)93,93,94
C     NATURAL SAMPLING
   93 VR=DM*SIN(2.0*PI*FO*TIME)
      GO TO 95
C     REGULAR SAMPLING
   94 VR=DM*SIN(2.0*PI*FO*TINC)
   95 VC=(2*L-3)*COS(2.0*PI*FI*(I-1)*DT)
C     IS THIS FOR POSITIVE OR NEGATIVE GROUP
      IF(L-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
```

A1/3

```
      WRITE(6,300)KP,TIME
  300 FORMAT(' ',4X,I3,6X,F9.4)
   10 CONTINUE
C
C     FOURIER COEFFICIENTS
      DO 40 N=1,NH
      A(N)=0.0
      B(N)=0.0
C
      DO 50 KP=2,NKP1
      R=((NP-2)/4.0-KP+L)*2.0*PI/NP
      WD=2.0*PI*(FI-N*FF)
      WS=2.0*PI*(FI+N*FF)
C     IS N=FI/FF
      Y=N-FI/FF
      ABSY=ABS(Y)
      IF(ABSY-0.01)48,48,47
   47 A(N)=A(N)+(COS(WS*T(KP-1)+R)-COS(WS*T(KP)+R))*FF/WS
     1+(COS(WD*T(KP-1)+R)-COS(WD*T(KP)+R))*FF/WD
      B(N)=B(N)+(SIN(WD*T(KP)+R)-SIN(WD*T(KP-1)+R))*FF/WD
     2-(SIN(WS*T(KP)+R)-SIN(WS*T(KP-1)+R))*FF/WS
      GO TO 50
   48 A(N)=A(N)+(T(KP)-T(KP-1))*(SIN(R))*FF/1.0-(COS(4.0*PI*FI*T(KP)+R
     3-COS(4.0*PI*FI*T(KP-1)+R))*FF/4.0/PI/FI
      B(N)=B(N)+(T(KP)-T(KP-1))*(COS(R))*FF/1.0-(SIN(4.0*PI*FI*T(KP)+R
     4-SIN(4.0*PI*FI*T(KP-1)+R))*FF/4.0/PI/FI
   50 CONTINUE
C
C     IS THIS FOR POSITIVE OR NEGATIVE GROUP
      IF(L-1)51,51,52
C     FOURIER COEFFICIENTS FOR NEGATIVE GROUP
   51 E(N)=A(N)
      F(N)=B(N)
C     HARMONICS DUE TO NEGATIVE GROUP
      HN(N)=SQRT(E(N)*E(N)+F(N)*F(N))
      ANGN(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)*C(N)+D(N)*D(N))
      ANGP(N)=180.0/PI*ATAN2(C(N),D(N))
   40 CONTINUE
C
      WRITE(6,315)
  315 FORMAT(' ','POSITIVE GROUP')
    5 CONTINUE
C
      DO 60 N=1,NH
C     FOURIER COEFFICIENTS OF COMPLETE CYCLOCONVERTER
      A(N)=(C(N)+E(N))/2.0
      B(N)=(D(N)+F(N))/2.0
C     HARMONICS DUE TO COMPLETE CYCLOCONVERTER
```

A1/4

```
      H(N)=SQRT(A(N)**2+B(N)**2)
      ANGC(N)=180.0/PI*ATAN2(A(N),B(N))
   60 CONTINUE
      IF(IC-1)80,80,81
C     NATURAL SAMPLING
   80 WRITE(6,340)
  340 FORMAT('1','CIRCULATING CURRENT CYCLO, NATURAL SAMPLING')
      GO TO 82
C     REGULAR SAMPLING
   81 WRITE(6,341)
  341 FORMAT('1','CIRCULATING CURRENT CYCLO, REGULAR SAMPLING')
C     OUTPUT FREQUENCY EXPRESSED AS HARMONIC OF THE FUNDAMENTAL
   82 NO=IFIX(FO/FF+0.01)
C
C     OUTPUT VOLTAGE (AT OUTPUT FREQUENCY)
      VN=HN(NO)*100.0
      VP=HP(NO)*100.0
      VO=H(NO)*100.0
      WRITE(6,347)
  347 FORMAT('0','VOLTAGES AT OUTPUT FREQUENCY')
      WRITE(6,348)VN
  348 FORMAT(' ','VN=',F6.2,1X,'0/0 OF INPUT VOLTAGE')
      WRITE(6,349)VP
  349 FORMAT(' ','VP=',F6.2,1X,'0/0 OF INPUT VOLTAGE')
      WRITE(6,350)VO
  350 FORMAT(' ','VO=',F6.2,1X,'0/0 OF INPUT VOLTAGE')
      WRITE(6,360)
  360 FORMAT('0','HARMONICS IN OUTPUT(0/0 OF VOLTAGE',3X,'PHASE ANGLE (
     1EGS) OF HARMS',3X,'PHASE ANGLE(DEGS) OF HARMS')
      WRITE(6,361)
  361 FORMAT(' ',7X,'AT OUTPUT FREQUENCY)',18X,'WRT B(N)',5X,'FROM T=0,(
     1 REFERRED TO')
      WRITE(6,370)
  370 FORMAT(' ',2X,'FREQ',2X,'POSITIVE',1X,'NEGATIVE',1X,'COMPLETE',3X
     3,'POSITIVE',1X,'NEGATIVE',1X,'COMPLETE',9X,'INPUT FREQ')
      WRITE(6,380)
  380 FORMAT(' ',3X,'HZ',4X,'GROUP',4X,'GROUP',4X,'CYCLO',6X,'GROUP',4X
     4'GROUP',4X,'CYCLO')
C
C     EXPRESS HARMONICS AS 0/0 OF OUTPUT VOLTAGE
      DO 70 N=1,NH
      HN(N)=HN(N)*100.0/VN*100.0
      HP(N)=HP(N)*100.0/VP*100.0
      H(N)=H(N)*100.0/VO*100.0
      FREQ=N*FF
C     PHASE ANGLES REFERRED TO INPUT FREQUENCY
      ANGPFI=FI/FREQ*ANGP(N)
      ANGNFI=FI/FREQ*ANGN(N)
      ANGCFI=FI/FREQ*ANGC(N)
      WRITE(6,400)FREQ,HP(N),HN(N),H(N),ANGP(N),ANGN(N),ANGC(N),ANGPFI,
     5ANGNFI,ANGCFI
  400 FORMAT(' ',4(F6.2,3X),5X,3(F7.2,2X),5X,3(F7.2,2X))
   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

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

$$\text{where } A_o = f_f \int_{t_1}^{t_2} F(t) \, dt \quad (A2.2)$$

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

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

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

Adaption of the Basic Principle of the Cycloconverter

(a)   Fourier Coefficient, A_o

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

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

where f_i = input frequency

and R depends on which input phase is being

A2/1

utilised at any particular instant to fabricate the output voltage waveform.

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

$$A_o = f_f \int_{t_1}^{t_2} \sin(2\pi f_i t + R) dt$$

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

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.

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

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

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

$$= F(t) \cos(2\pi N f_f t)$$

$$= \sin(2\pi f_i t + R) \cos(2\pi N f_f t) \qquad \text{(from (A2.5))}$$

$$= \frac{1}{2} \sin(W_D t + R) + \frac{1}{2} \sin(W_S t + R) \qquad (A2.7)$$

where $W_D = 2\pi(f_i - N f_f)$

$W_S = 2\pi(f_i + N f_f)$

$\therefore$ from (A2.7) and (A2.3)

$$A_N = f_f \int_{t_1}^{t_2} \left[\sin(W_D t + R) + \sin(W_S t + R) \right] dt$$

$$= -f_f \left[\frac{1}{W_D} \cos(W_D t + R) + \frac{1}{W_S} \cos(W_S t + R) \right]_{t_1}^{t_2}$$

This expression can be re-written as:

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

$$+ \sum_{K=1}^{N_K} \frac{f_f}{W_S} \left[\cos(W_S\, t(K-1) + R) - \cos(W_S\, t(K) + R) \right] \tag{A2.8}$$

In the special case when $f_i = N f_f$ giving $W_D = 0$, the above expression cannot apply.

$\therefore$ in this case (A2.7) becomes:

$$\frac{1}{2} \sin(R) + \frac{1}{2} \sin(4\pi f_i t + R) \tag{A2.9}$$

$$\text{since } W_S = 2\pi(f_i + N f_f) = 4\pi f_i$$

$\therefore$ from (A2.9) and (A2.3), and using $t(K)$ and $t(K-1)$,

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

$$+ \sum_{K=1}^{N_K} \frac{f_f}{4\pi f_i} \left[\cos(4\pi f_i\, t(K-1) + R) - \cos(4\pi f_i\, t(K) + R) \right] \tag{A2.10}$$

(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 $N \neq \dfrac{f_i}{f_f}$,

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

$$- \sum_{K=1}^{N_K} \frac{f_f}{W_S} \left[\sin(W_S t(K) + R) - \sin(W_S\, t(K-1) + R) \right] \quad \text{(A2.11)}$$

For $N = \dfrac{f_i}{f_f}$,

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

$$- \sum_{K=1}^{N_K} \frac{f_f}{4\, f_i} \left[\sin(4\pi f_i\, t(K) + R) - \sin(4\pi f_i\, t(K-1) + R) \right]$$
$$\text{(A2.12)}$$

(d)   Magnitudes and Phase Angles of the Harmonics (H_N, Θ_N)

$$H_N = (A_N^2 + B_N^2)^{\frac{1}{2}} \qquad\qquad \text{(A2.13)}$$

$$\Theta_N = \tan^{-1} \frac{B_N}{A_N} \qquad\qquad \text{(A2.14)}$$

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

ϕ = phase angle defining which input phase voltage is conducting

w = frequency of the input voltage to the cycloconverter

also $v = iR + L \dfrac{di}{dt}$

where R = resistance of the load

L = inductance of the load

i = instantaneous load current

$\therefore K \sin (wt + \phi) = iR + L \dfrac{di}{dt}$

Taking Laplace Transforms:

$$\frac{K(w \cos \phi + s \sin \phi)}{s^2 + w^2} = (R + Ls)\, I - Li(0)$$

$\therefore I = \dfrac{K(w\cos\phi + s \sin \phi) + (s^2 + w^2)(Li(0))}{(s^2 + w^2)(R + Ls)}$ \hfill (A3.1)

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

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

$$= (As + B)(R + Ls) + C(s^2 + w^2)$$

$$= ARs + BLs + BR + ALs^2 + Cs^2 + Cw^2$$

equating s^0, s^1 and s^2:

$$Kw \cos \emptyset + w^2 Li(0) = BR + Cw^2 \qquad (A3.3)$$

$$K \sin \emptyset = AR + BL \qquad (A3.4)$$

$$Li(0) = AL + C \qquad (A3.5)$$

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

$$Kw \cos \emptyset + w^2 Li(0) = BR + (Li(0) - AL)w^2$$

$$\therefore Kw \cos \emptyset = BR - ALw^2 \qquad (A3.6)$$

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

$$Kw \, R \cos \emptyset + w^2 LK \sin \emptyset = BR^2 + BL^2 w^2$$

$$\therefore B = \frac{K(wR \cos \emptyset + w^2 L \sin \emptyset)}{R^2 + w^2 L^2} \qquad (A3.7)$$

$\therefore$ from (A3.4) and (A3.7):

$$A = \frac{K}{R}\left(\sin \emptyset - \frac{wLR \cos \emptyset + w^2 L^2 \sin \emptyset}{R^2 + w^2 L^2}\emptyset\right)$$

$$= \frac{K}{(R^2 + w^2 L^2)} \, (R \sin \emptyset - wL \cos \emptyset) \qquad (A3.8)$$

$\therefore$ from (A3.5) and (A3.8):

$$C = Li(0) - \frac{K}{R^2 + w^2 L^2} \, (RL \sin \emptyset - wL^2 \cos \emptyset) \qquad (A3.9)$$

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

$$I = \frac{K}{(R^2 + w^2 L^2)} \left[\frac{R(w \cos \emptyset + s \sin \emptyset)}{s^2 + w^2} \right.$$

$$+ \frac{wL(w \sin \emptyset - s \cos \emptyset)}{s^2 + w^2}$$

$$\left. + \frac{wL \cos \emptyset - R \sin \emptyset + (R^2 + w^2 L^2) \, i(0)}{s + R/L} \right]$$

transforming back to time function gives:

$$i = \frac{K}{(R^2 + w^2L^2)} \left[R \sin (wt + \emptyset) - wL \cos (wt + \emptyset) \right.$$

$$\left. + \exp(- \frac{R}{L}t).(wL \cos \emptyset - R \sin \emptyset + (R^2 + w^2L^2)i(0)) \right]$$

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

$$i = \frac{K}{R(1 + w^2t_c^2)} \left[\sin (wt + \emptyset) - wt_c \cos(wt + \emptyset) \right.$$

$$\left. + \exp(- \frac{t}{t_c}).(wt_c \cos \emptyset - \sin \emptyset) \right]$$

$$+ \exp(- \frac{t}{t_c}).(i(0))$$

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

APPENDIX 4

COMPUTER PROGRAM FOR THE
INHIBITED CYCLOCONVERTER

```
C       THYRISTOR CYCLOCONVERTER WITH SINGLE PHASE OUTPUT
C           CIRCULATING CURRENT-FREE MODE
C           COSINUSOIDAL CONTROL
C           SINUSOIDAL REFERENCE WAVE
C           NATURAL SAMPLING
C           R-L LOAD
C           GROUP CHANGEOVER AT COMPUTATION POINT NEAREST
C                   INSTANT OF ZERO OUTPUT CURRENT AFTER ZERO VR
C
C       REQUIRED INPUT DATA
C       NP=PULSE NUMBER OF INPUT (=2, 3 OR 6 IN THIS PROGRAM)
C       FI=INPUT FREQUENCY (HZ)
C       FO=OUTPUT FREQUENCY (HZ)
C       FF=FREQUENCY OF FUNDAMENTAL(LOWEST FREQUENCY IN OUTPUT)
C           FO/FF MUST EQUAL AN INTEGER (=FA)
C               NP*FI/FF MUST EQUAL AN INTEGER
C       RAT= NP*FI/FO
C       NH=HIGHEST ORDER OF HARMONIC IN THE OUTPUT
C       DM=MODULATION DEPTH
C
C       ADDITIONAL SYMBOLS USED IN THIS PROGRAM
C       ND=NUMBER OF DATA CARDS (EXCLUDING CARD STATING ND)
C       N=ORDER OF HARMONIC,REFERRED TO FF
C       I=COUNTER FOR SAMPLING
C       KP=COMMUTATION POINT (=1 AT FIRST COMMUTATION AFTER T=0)
C       NKP=NO. OF KP IN CC MODE
C       KC=COMMUTATION POINT FOR CHANGEOVER IN INHIBITED CYCLOCONVERTER
C       KC=1 IS FOR CHANGEOVER FROM POSITIVE TO NEGATIVE GROUP
C       FOURIER ANALYSIS IS FROM KC=2 TO AVOID START ERRORS
C       KCL=LAST KC
C       KPP(KC)=POSITIVE GROUP KP NEXT TO KC
C       KPN(KC)=NEGATIVE GROUP KP NEXT TO KC
C       KPS='DUMMY'KP FOR START OF FOURIER ANALYSIS FOR ONE GROUP
C       KPF='DUMMY'KP FOR FINISH OF DITTO
C       TIME=INSTANT AT WHICH COMPUTATION IS TO BE CARRIED OUT
C           TIME=0.0 SECS AT START OF SINUSOIDAL REFERENCE VOLTAGE
C       TIMC=TIME OF START OF VC
C       DT=INCREMENTAL TIME INTERVAL FOR COMPUTATION
C       TP(KP)=TIME OF COMMUTATION OF POSITIVE GROUP
C       TN(KP)=DITTO OF NEGATIVE GROUP
C       TC(KC)=TIME OF START OF CONDUCTION AT
C               CHANGEOVER COMMUTATION POINT KC
C       VR=REFERENCE VOLTAGE
C       AL IS PROPORTIONAL TO THE LOAD CURRENT
C       ANGC(NO)=COMPUTED(OUTPUT) PHASE SHIFT OF AO FROM VR
C       VC=COSINUSOIDAL CONTROL VOLTAGE
C       L=2 FOR POSITIVE GROUP
C        =1 FOR NEGATIVE GROUP
C       A(N)=FOURIER COEFFICIENT FOR SINE TERMS
C       B(N)=FOURIER COEFFICIENT FOR COSINE TERMS
C       H(N)=MAGNITUDE(PEAK,P.U.) OF THE NTH HARMONIC
C       WD=DIFFERENCE IN ANGULAR FREQUENCY BETWEEN FI AND HARMONIC
C       WS=SUM OF DITTO
C           1P.U.=PEAK OF INPUT VOLTAGE
C
        DIMENSION TP(500),TN(500),TC(120),KPP(120),KPN(120),T(500)
        DIMENSION H(300),ANGC(300),ALOAD(5000),TLOAD(5000)
        READ(5,199)ND
  199   FORMAT(I5)
        DO 90 JK=1,ND
        READ(5,200)FI,RAT,FA,JH,NP,DM,TIMCO
  200   FORMAT(3F10.0,2I10,F10.0,F10.0)
        FO=NP*FI/RAT
        FF=FO/FA
C
        WRITE(6,218)
  218   FORMAT('1',22('START'))
        WRITE(6,201)
  201   FORMAT('0',5X,'INPUT DATA')
        WRITE(6,205)NP
  205   FORMAT('0','NUMBER OF PULSES=',I3)
        WRITE(6,202)FI
  202   FORMAT(' ','INPUT FREQUENCY=',F10.5,'HZ')
        WRITE(6,213)RAT
  213   FORMAT(' ','RATIO=',F10.5)
```

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```
       WRITE(6,203)FO
   203 FORMAT(' ','OUTPUT FREQUENCY=',F10.5,'HZ')
       WRITE(6,204)FF
   204 FORMAT(' ','FUNDAMENTAL FREQUENCY=',F10.5,'HZ')
       WRITE(6,207)NH
   207 FORMAT(' ','HIGHEST ORDER OF HARMONIC=',I3)
       WRITE(6,206)DM
   206 FORMAT(' ','MODULATION DEPTH=',F10.5,'P.U.')
       WRITE(6,212)TIMCO
   212 FORMAT(' ','TIME CONSTANT=',F10.7,'SECS')
       WRITE(6,214)
   214 FORMAT('0', 5X,'*** COSINUSOIDAL CONTROL, SINUSOIDAL REFERENCE, NA
      1TURAL SAMPLING ***')
C
       DO 91 IC=1,3
C
C         CONSTANTS
       PI=22.0/7.0
       DT=1.0/FI/300.0
       NKP=IFIX(NP*FI/FF+0.01)+IFIX(3*NP*FI/FO+0.01)
       NKP1=NKP+1
       NO=IFIX(FO/FF+0.01)
       KCL=2+2*NO
       WI=2.0*PI*FI
       WO=2.0*PI*FO
       NIA=IFIX(KCL/2.0/FO/DT/4.0)
       IF(KCL .GE. 110) GO TO 86
       IF(NKP1 .GE. 500) GO TO 86
       IF(NH .GE. 300) GO TO 86
       IF(NIA .GE. 5000) GO TO 86
       GO TO 85
    86 WRITE(6,215)
   215 FORMAT(' ',5('DIMENSION EXCEEDED',5X))
       GO TO 90
    85 CONTINUE
C
       WRITE(6,219)
   219 FORMAT('1',22('NEXT'))
       IF(IC-2)96,93,94
    96 WRITE(6,208)
   208 FORMAT('0',5X,'OUTPUT DATA FOR ',8('VR/-0',3X))
       GO TO 221
    93 WRITE(6,209)
   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','COMMUTATION POINTS FOR UNINHIBITED CYCLOCONVERTER')
       WRITE(6,210)
   210 FORMAT(' ',2X,'GROUP',8X,'KP',6X,'TIME(SECS)',5X,'MODULATION(SECS)
      1')
C
C
C      COMMUTATION POINTS FOR UNINHIBITED CYCLOCONVERTER
       DO 5 L=1,2
       DO 10 KP=1,NKP1
       IF(NP-3)77,78,79
    77 TIMC=(KP-1)/2.0/FI
       GO TO 92
    78 TIMC=(2*KP-4+L)/6.0/FI
       GO TO 92
    79 TIMC=(KP-3+L)/6.0/FI
    92 DO 20 I=1,151
       TIME=(I-1)*DT+TIMC
       VC=(2*L-3)*COS(2.0*PI*(I-1)/300.0)
       VR=DM*SIN(2.0*PI*FO*TIME-(IC-1)*2.0*PI/3.0)
C      IS THIS FOR POSITIVE OR NEGATIVE GROUP
       IF(L-1)15,15,16
    16 IF(VR-VC)20,21,21
    15 IF(VC-VR)20,22,22
    20 CONTINUE
       WRITE(6,230)
   230 FORMAT(' ',1X,'NO COMMUTATION')
C
    21 TP(KP)=TIME
```

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```
        TIMOD=TIME-TIMC
        WRITE(6,300)KP,TIME,TIMOD
  300   FORMAT(' ','POSITIVE',6X,I3,8X,F7.4,8X,F7.4)
        GO TO 10
   22   TN(KP)=TIME
        TIMOD=TIME-TIMC
        WRITE(6,310)KP,TIME,TIMOD
  310   FORMAT(' ','NEGATIVE',6X,I3,8X,F7.4,8X,F7.4)
C
   10   CONTINUE
    5   CONTINUE
C
C
C                      TIMES FOR CHANGEOVER IN INHIBITED CYCLOCONVERTER
C
        L=2
        IA=0
        TIMI=0.0
        KCL1=KCL+1
C
        DO 150 KC=1,KCL1
        AO=0.0
        TIMVZ=KC/2.0/FO+(IC-1)/3.0/FO
        IF(KC .EQ. 1) GO TO 141
        TIMI=TC(KC-1)
  141   IF(L-1)154,154,151
C
C          FIND FIRST KP OF INCOMING(POSITIVE) GROUP
  151   DO 153 KP=1,NKP1
        IF(TP(KP) .GE. TIMI) GO TO 152
  153   CONTINUE
  152   IF(KC .EQ. 1) GO TO 142
        KPP(KC-1)=KP
        GO TO 142
C
C          FIND FIRST KP OF INCOMING(NEGATIVE) GROUP
  154   DO 156 KP=1,NKP1
        IF(TN(KP) .GE. TIMI) GO TO 140
  156   CONTINUE
  140   KPN(KC-1)=KP
C
  142   KP1=KP
        DO 146 KP=KP1,1000
        IF(KP-3)158,159,160
  158   R=(L-KP)*PI
        GO TO 161
  159   R=(4*(L-KP)+1)*PI/6.0
        GO TO 161
  160   R=(2*L-KP-1)*PI/3.0
  161   TIM=TIMI
  162   IF(TIMCO .LT. 0.0001) GO TO 155
        AL=1.0/(1.0+(WI*TIMCO)**2)*(SIN(WI*TIM+R)-WI*TIMCO
       1*COS(WI*TIM+R)+(EXP((TIMI-TIM)/TIMCO))*(WI*TIMCO*COS(R+WI*TIMI)-
       2SIN(R+WI*TIMI)))+AO*EXP((TIMI-TIM)/TIMCO)
        GO TO 157
  155   AL=SIN(WI*TIM+R)
  157   IF(L-1)168,168,162
  168   IF(AL+0.00001)164,170,170
  169   IF(AL .GT. 0.00001) GO TO 164
  170   AL=(2*L-3)*0.00001
        IF(TIM .GE. TIMVZ) GO TO 163
  164   TIM=TIM+DT*4.0
        IF(L-1)143,143,144
  143   IF(TIM .LT. TN(KP)) GO TO 165
        TIMI=TN(KP)
        GO TO 145
  144   IF(TIM .LT. TP(KP)) GO TO 165
        TIMI=TP(KP)
  145   AO=AL
  146   CONTINUE
C
  165   IA=IA+1
        ALOAD(IA)=AL
        TLOAD(IA)=TIM-DT*4.0
        GO TO 162
C
C          CHANGEOVER OCCURS NOW
```

```
      163 IA=IA+1
          ALOAD(IA)=AL
          TLOAD(IA)=TIM
          TIM=TIM+DT*4.0
          TC(KC)=TIM
C            RECORD LAST KP OF OUTGOING GROUP AND THEN CHANGE TO OTHER GROUP
          IF(L-1)171,171,166
      171 KPN(KC)=KP-1
          L=2
          GO TO 167
      166 KPP(KC)=KP-1
          L=1
      167 CONTINUE
C
      150 CONTINUE
C
          ILAST=IA
C
C     FOURIER ANALYSIS
          DO 67 N=1,NH
          A=0.0
          B=0.0
          DCCOMP=0.0
          L=2
          IKP1=2
C
          DO 53 KC=3,KCL
C         WHICH GROUP IS THIS
          IF(L-1)51,51,52
C         POSITIVE GROUP
       52 KPS=KPP(KC-1)-1
          KPF=KPP(KC)+1
          GO TO 57
C         NEGATIVE GROUP
       51 KPS=KPN(KC-1)-1
          KPF=KPN(KC)+1
C
       57 TS=TC(KC-1)
          TF=TC(KC)
          KPS1=KPS+1
          DO 50 KP=KPS1,KPF
C         WHICH GROUP IS THIS
          IF(L-1)59,59,58
C         POSITIVE GROUP
       58 T(KP)=TP(KP)
          T(KP-1)=TP(KP-1)
          GO TO 64
C         NEGATIVE GROUP
       59 T(KP)=TN(KP)
          T(KP-1)=TN(KP-1)
       64 T(KPS)=TS
          T(KPF)=TF
C
C     CHECK FOR CURRENT DISCONTINUITIES
          ISTART=IKP1-1
          DO 60 IA=ISTART,ILAST
          IF(TLOAD(IA) .GE. T(KP-1)) GO TO 61
       60 CONTINUE
          WRITE(6,328)
      328 FORMAT('0','ERROR?')
       61 IKP1=IA
          DO 62 IA=IKP1,ILAST
          IF(TLOAD(IA) .GT. T(KP)) GO TO 63
       62 CONTINUE
          WRITE(6,329)
      329 FORMAT('0','ERROR???')
       63 IKP2=IA-1
          IKPX=IKP1+1
          IF(IKPX .GT. IKP2) GO TO 69
          IF(L .EQ. 2) GO TO 42
          IF(ALOAD(IKPX)+0.00002)43,43,44
       42 IF(ALOAD(IKPX)-0.00002)44,43,43
       44 DO 45 IA=IKPX,IKP2
          IF(L .EQ. 2) GO TO 46
          IF(ALOAD(IA)+0.00002)49,49,45
       46 IF(ALOAD(IA)-0.00002)45,49,49
```

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```
   45 CONTINUE
      T(KP-1)=T(KP)
      GO TO 69
   49 T(KP-1)=TLOAD(IA-1)
   43 DO 65 IA=IKPX,IKP2
      IF(L .EQ. 2) GO TO 60
      IF(ALOAD(IA)+0.00002)65,65,68
   66 IF(ALOAD(IA)-0.00002)68,65,65
   65 CONTINUE
      GO TO 69
   68 T(KP)=TLOAD(IA)
C
   69 IF(NP-3)76,73,74
   76 R=(L-KP)*PI
      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 WD=2.0*PI*(FI-N*FF)
      WS=2.0*PI*(FI+N*FF)
      DCCOMP=DCCOMP+FF/2.0/PI/FI*(COS(2.0*PI*FI*T(KP-1)+R)-COS(2.0*PI*FI
     1*T(KP)+R))
C     FOURIER COEFFICIENTS
C     IS N=FI/FF
      Y=N-FI/FF
      ABSY=ABS(Y)
      IF(ABSY-0.01)48,48,47
   47 A=A+((COS(WS*T(KP-1)+R)-COS(WS*T(KP)+R))*FF/WS
     1+((COS(WD*T(KP-1)+R)-COS(WD*T(KP)+R))*FF/WD
      B=B+((SIN(WD*T(KP)+R)-SIN(WD*T(KP-1)+R))*FF/WD
     2-(SIN(WS*T(KP)+R)-SIN(WS*T(KP-1)+R))*FF/WS
      GO TO 50
   48 A=A+(T(KP)-T(KP-1))*(SIN(R))*FF/1.0-(COS(4.0*PI*FI*T(KP)+R)
     3-COS(4.0*PI*FI*T(KP-1)+R))*FF/4.0/PI/FI
      B=B+(T(KP)-T(KP-1))*(COS(R))*FF/1.0-(SIN(4.0*PI*FI*T(KP)+R)
     4-SIN(4.0*PI*FI*T(KP-1)+R))*FF/4.0/PI/FI
   50 CONTINUE
C
      IF(L-1)55,55,56
   55 L=2
      GO TO 53
   56 L=1
   53 CONTINUE
C
C     HARMONICS DUE TO COMPLETE CYCLOCONVERTER
      H(N)=SQRT(A*A+B*B)
C
C     PHASE DIFFERENCE BETWEEN H(N) AND B(N)
      ANGC(N)=ATAN2(A,B)*180.0/PI
   67 CONTINUE
C     END FOURIER COEFFICIENTS
C
C     OUTPUT VOLTAGE (AT OUTPUT FREQUENCY)
      VO=H(NO)*100.0
      WRITE(6,350)VO
  350 FORMAT(' ',1X,'VOLTAGE AT OUTPUT FREQ (VO)= ',F7.2,1X,'0/0 OF INPU
     2T VOLTAGE')
C
C     PHASE ANGLE (AT OUTPUT FREQUENCY)
      WRITE(6,352)ANGC(NO)
  352 FORMAT(' ',1X,'PHASE DIFF BETWEEN VO AND VR=',F7.2,'DEGS AT OUTPUT
     2 FREQ')
      PSFI=ANGC(NO)*FI/FO
      WRITE(6,354)PSFI
  354 FORMAT(' ',1X,'PHASE DIFF BETWEEN VO AND VR=',F7.2,'DEGS AT INPUT
     1FREQ')
C     DC COMPONENT
      DCCOMP=DCCOMP*10000.0/VO
      WRITE(6,359)DCCOMP
  359 FORMAT('0',1X,'DC COMPONENT=',F10.3,'% OF VO')
C
      WRITE(6,360)
  360 FORMAT('0',1X,'HARMONICS IN OUTPUT')
      WRITE(6,370)
  370 FORMAT(' ',4X,'FREQ(HZ)',2X,'0/0 OF VO',8X,'PHASE ANGLES')
      WRITE(6,371)
```

```
  371 FORMAT(' ',28X,'B(N)',7X,'FI',7X,'FO')
C
C       EXPRESS HARMONICS AS 0/0 OF OUTPUT VOLTAGE
C                           AND CALCULATE DISTORTION FACTOR
        SUSQ=0.0
        SUSQW1=0.0
        DO 70 N=1,NH
        H(N)=H(N)*100.0/V0*100.0
        FREQ=N*FF
        ANGFI=ANGC(N)*FI/FREQ
        ANGFO=ANGC(N)*FO/FREQ
        WRITE(6,400)FREQ,H(N),ANGC(N),ANGFI,ANGFO
  400 FORMAT(' ',5(1X,F10.3))
        SUSQ=SUSQ+H(N)*H(N)
        SUSQW1=SUSQW1+(H(N)*FO/FREQ)**2
   70 CONTINUE
C
        DF=100.0/SQRT(SUSQ)
        DFW1=100.0/SQRT(SUSQW1)
        WRITE(6,390)DF
  390 FORMAT('0','UNWEIGHTED DISTORTION FACTOR=',F10.4)
        WRITE(6,391)DFW1
  391 FORMAT(' ','WEIGHTED(FO/FH) DISTORTION FACTOR=',F10.4)
C
C
   91 CONTINUE
C
   90 CONTINUE
        STOP
        END
```

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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.

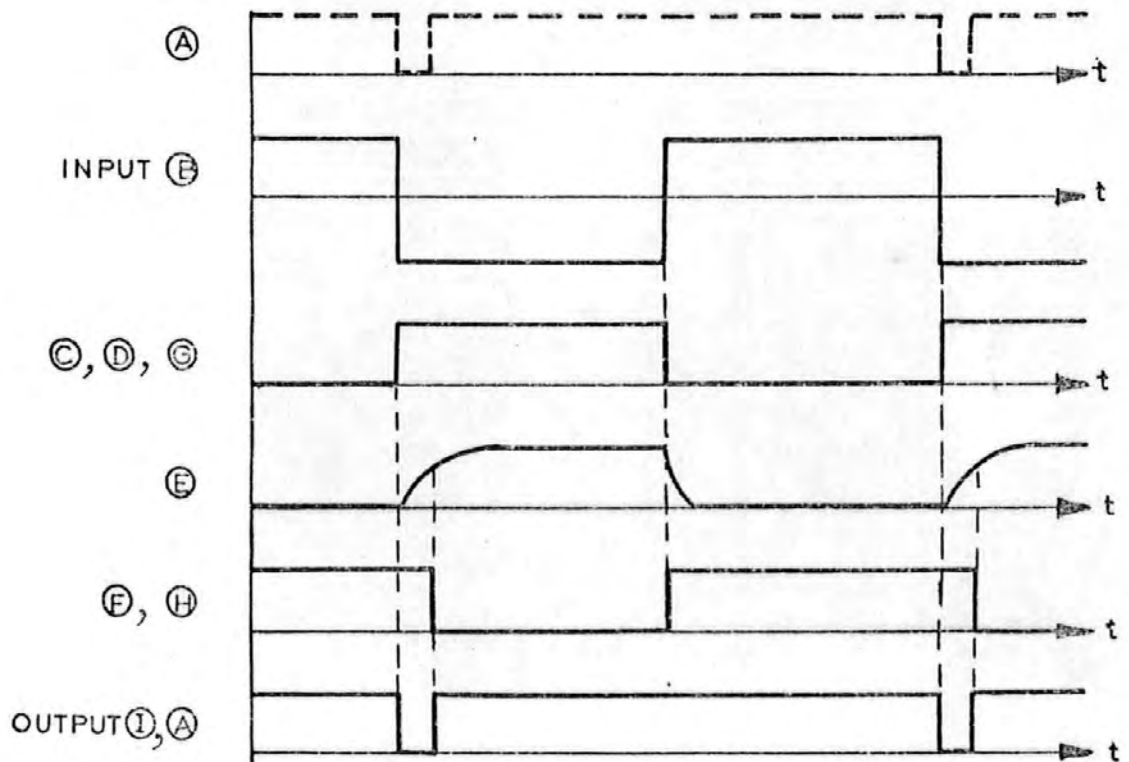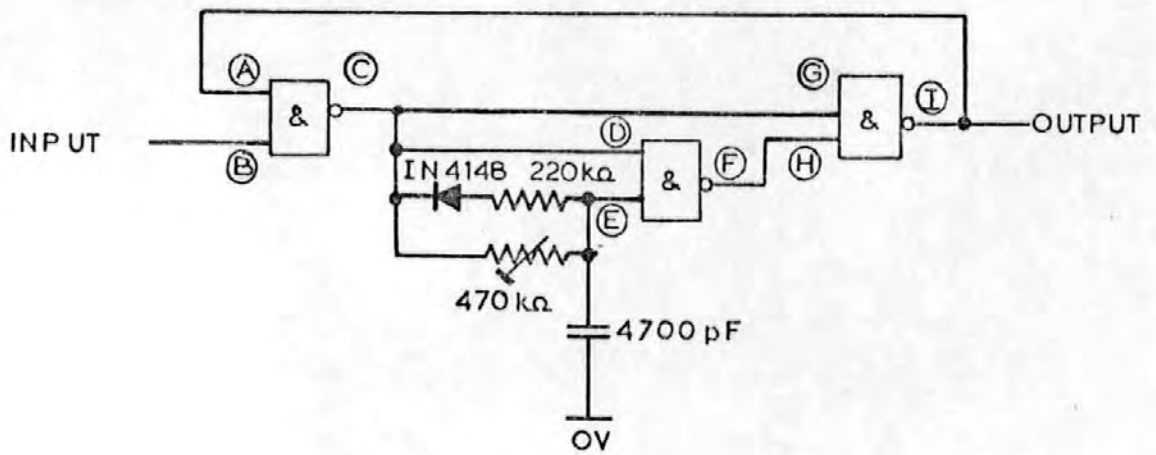The discharge of the capacitor is not utilised

FIG A6.1 MONOSTABLE CIRCUIT
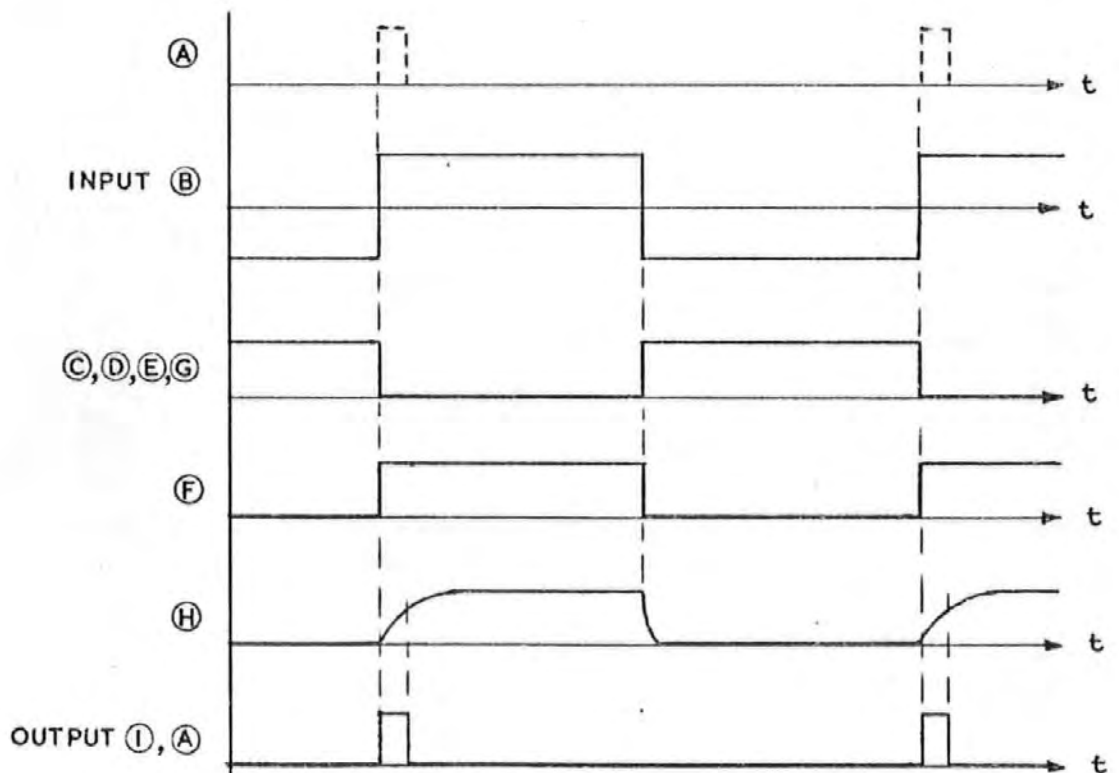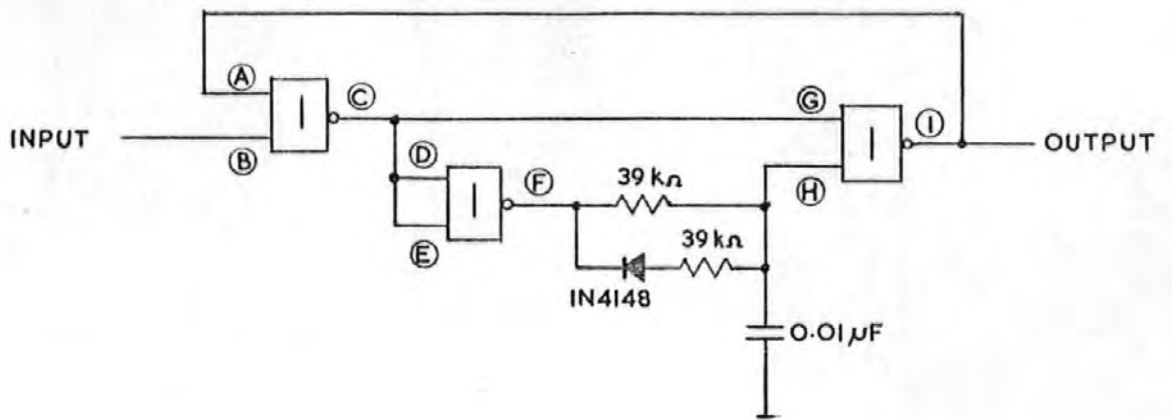DIAGRAM AND WAVEFORMS:
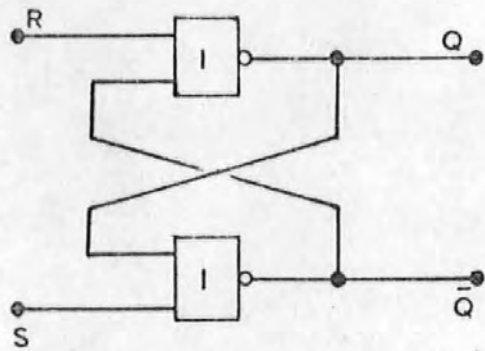NAND-GATES

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

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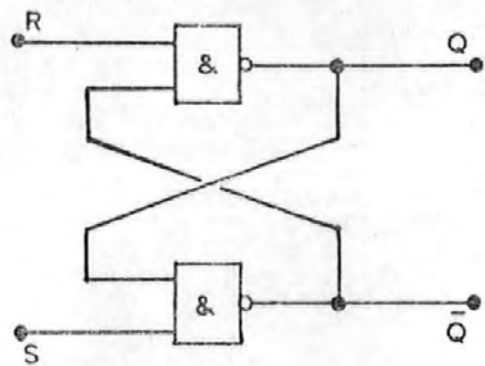resistance branch is added to accelerate it.

(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 $\overline{Q}$.

R	S	Qt	Qt+1
O	O	O	O
O	O	I	I
O	I	O	O
O	I	I	O
I	O	O	I
I	O	I	I
I	I	} INDETERMINATE	
I	I		

a) USING NOR-GATES

R	S	Qt	Qt+1
I	I	O	O
I	I	I	I
I	O	O	O
I	O	I	O
O	I	O	I
O	I	I	I
O	O	} INDETERMINATE	
O	O		

b) USING NAND-GATES

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