THE DESIGN OF A TRAFFIC DETECTOR USING THE PHASE-LOCK-LOOP CONCET

By

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A project report submitted to the Department of Electrical Engineering, Ahmadu Bello University, Zaria, Nigeria in partial fulfilment of the requirements for the award of of the degree

MASTER OF ENGINEERING

in

ELECTRONICS AND TELECOMMUNICATIONS

June 1980

APPROVED BY:-

EXTERNAL EXAMINER. . . . . .

INTERNAL EXAMINER. . . . . .

INTERNAL EXERMINER. . . . . .

DEDICATION

Dedicated to the PRINCE OF PEACE

Who alone can prevent Delay and

Destruction at the Cross-road of

Human life.

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ABSTRACT

The Design of a Traffic Detector using the phase-lock-loop concept is presented. The Detector is capable of adjusting to environmental fluctuations with no loss of accuracy. The Detector oscillator is discussed in light of the components susceptible to changes in environmental conditions and the ability of the phase-lock-loop discriminative circuit to cope with such drifts. The versatility of the Detector is demonstrated by illusting two of its many potential applications - the Automatic Traffic Data Collection and Semi-Vehicle-Actuated Traffic Control.

ACKNOWLEDGEMENT

I am greatly indebted to Engr. Dr. B.J. Olufeagba, Ph.D., M.I.E.E.E., M.N.S.E., Senior Lecturer who supervised the project. His help and untiring efforts cannot be over acknowledged.

To Mr. P.F.U. Taylor, M.Sc., I am also indebted for his assistance both morally and materially. I must thank the Head of Department, Professor D.G.O. Morris and all the Departmental staff at all levels who have assisted in one way or another to see to the successful end of the project. Mention must be made of Mr. Mike who typed the manuscript painstakingly.

Last but not the least my profound gratitude also goes to 'Biola and 'Tosin for their understanding and co-operation.

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List of Symbols

ω	-	angular frequency
f	-	frequency
L	-	Inductance
C	-	Capacitance
δ	-	Depth of penetration
η	-	wave impedance
μ	-	relative permitivity
σ	-	conductivity
KHz	-	one thousand Hertz
gm	-	transfer conductance
h_{ie}	-	Input impedance (transistor common emitter)
h_{oe}	-	Output impedance (" " ")
h_{fe}	-	Transistor current gain
V_{cc}	-	Supply Voltage
X_c	-	Impedance of a capacitor
j	-	$(-1)^{\frac{1}{2}}$
Σ	-	Summation Sin
nF	-	Nanofarad
μF	-	Microfarad
mA	-	Milliampere
$\int$	-	Integral sign
$\bar{A}$	-	compliment of A

Objective of the Thesis

In this thesis, the design and implementation of a versatile Traffic Detector which can be used for traffic management and control is presented. It is one of the projects embarked upon at Ahmadu Bello University, Department of Electrical Engineering which not only serves to demonstrate the process of technology development (6) but will also permit the proper exploitation of a family of versatile feedback Traffic Controllers developed in the Department. The available technology is used to solve the local problems in a standard fashion.

The Traffic Detector employs a phase-locked-loop concept to convert the passage of vehicles into electrical signals. It is reliable and easy to instal. It is economical and since the power consumption is low, readily maintainable. The versatility of the Detector is demonstrated by two examples from a wide spectrum of potential applications. These are in the traffic survey data collection where a hard copy is automatically produced as a histogram and the semi-vehicle-actuated traffic controller of a T-junction. The design procedure employs known theoretical and practical engineering methods and a systematic analysis of circuits and subsystems is presented.

CHAPTER ONE

INTRODUCTION

1.0 Synopsis

The need for and ways of Traffic Control are stated. The advantages and types of Traffic Detectors in traffic management are enumerated.

1.1. Maximising Utility of Automobile Mobility

One of the greatest achievements of the present day civilization is mobility which improves the social life and promotes economic growth. This achievement notwithstanding, mobility has brought with it many socio-economic problems. Road transportation which is undoubtedly a major means of information, goods and people transfer both in Urban and Rural areas has associated with it traffic congestions in Urban centres and this is costly in social and financial terms. Time is lost. Fuel is wasted and people suffer from the nuisance of start/stop of engines and exhaust pollution while the engine, transmission and brakes deterioriate .

Life is lost in accidents caused by uncontrolled Urban Traffic.

In order to derive the maximum benefit from automobile mobility, there is need for sound Urban Traffic management - "Control of Traffic to provide safe, convenient and economic movement of vehicles and pedestrians".

A considerable amount of effort is now devoted to optimizing many attributes of the road networks. These include the reduction of the journey time and accident rates.

Many strategies[1] have been employed to combat the problem arising from increased use of motor transportation in urban centres. These include the construction of new roads, flyovers and highways aimed at increasing the maximum possible traffic volume, while in some instances reduction of Traffic density is achieved by a compulsory assignment of odd and even number days, one way systems, parking regulations; reducing of the delays and conflicts at intersections by using human controllers, i.e. policeman, Traffic Wardens; and Automatic Traffic Signals. The use of Automatic Traffic Signals provides a very economical and versatile method that can reduce both the accident rate and delay at intersections.

1.2 Traffic Detectors in Networks and Intersection Control

The development[2] of Traffic Control from manually operated to fixed time signals where the right of way for each approach is predetermined has helped to ease traffic congestion but with sub-optimal efficiency especially for junctions with considerable variation in traffic volume.

In 1930, a Vehicle Actuated Traffic Signal was installed in the United States of America with a microphone being used as the Vehicle Detector. While the Vehicle Detection by such means presented an approach for making system timing responsive, the use of microphones was rapidly dropped for more reliable and less temperamental methods. In this thesis, a Vehicle Detector using a magnetic inductive loop is developed, analysed and used for a type of Vehicle Actuated control and as

a traffic survey instrument.

Vehicle-Actuated control of Traffic with its advantage of flexibility and sensitivity to Traffic Variation is now commonly used at urban centres as well as at some isolated junctions where the side-road Traffic experiences very long delays. It also helps to reduce the nuisance of turning Vehicles at these intersections.

Up to date information of the Traffic situation is necessary for proper urban traffic control. Traffic Detectors have also come to assume a very important role in Data Collection Surveys. These surveys are frequently used for planning and timing studies and the Detectors tend to provide a cheap and reasonably reliable source for 24-hour data collection. The Detector in this thesis is able to perform this function and in a simple example present the data as a histogram.

1.3 Types of Traffic Detectors

Automatic data collection involves the detection of the vehicle by vehicle detectors and counting of the same. The vehicle detectors used are usually installed on the road surface and include a variety of techniques[1] such as given below.

(i) The Pneumatic Detectors

This type of detector is quite popular in the United Kingdom and needs careful maintenance for reliable operation. It consists of a

thick-walled rubber tube of about 12 mm diameter, a diaphram and a counter. The rubber tube is fixed to the road surface perpendicular to the flow of traffic. One end is terminated with a plug having an air-release hole to avoid echo pulses while the other is terminated by a diaphram. When the vehicle wheels cross, the compression of the tube generates air waves which propagate in both directions. The diaphram is then pushed to operate a make-and-break contact which the counter records.

This type of Detector appears simple but requires regular maintenance. The tube must be checked regularly for punctures and blockages while the diaphram and switch must be cleaned and renewed regularly.

(ii) The Hydraulic Detector

This type of Detector is similar in set up to the pneumatic type with the exeption that the tube is filled with liquid to give better sensitivity. However the maintenance of this type is more difficult as the liquid must be replaced regularly.

(iii) Electromagnetic Beam Detector

In this class are the photo electric method, radar, infra red and ultrasonic beams. The electromagnetic beam is focused across the road and the passage of vehicles causes an interruption which is interpreted by the Receiver. The circuitry and installations are quite

complicated and maintenance is quite difficult.

(iv) The Phase-Locked-Loop Vehicle Detector

An Inductive Vehicle Detector requires less stringent mechanical maintenance than even the Pneumatic type which appears to be the simplest of the aforementioned types. The installation involves the cutting of a narrow slot on the road surface where the coil is buried. It is safer since no intruder, (children) can have easy access to it. This will also function well under varying weather conditions characteristic of Nigeria.

The type developed in this thesis uses a phase-locked-loop and as such it is self-adjusting and can accommodate parameter variations of the electric circuit usually associated with time and weather. The detector is represented diagramatically in fig. 1.1. It consists of a detector coil buried just below the road surface. This is a rectangular coil made of few turns of copper wire and insulated from moisture. The coil forms part of a Detector Oscillator (D.O.) whose frequency is changed when the wheels of a vehicle crosses it. The sudden change in frequency of the D.O. is converted into a d.c. signal by the phase-locked-loop discriminator. The output of the PLL is amplified by an operational amplifier. A schmitt trigger monostable circuit is included which

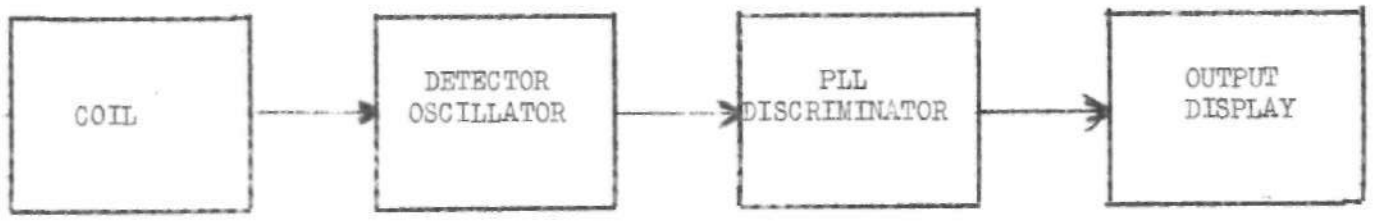


Fig. 1.1.: INDUCTIVE PLL TRAFFIC DETECTOR

converts the output of the operational amplifier into electrical pulses that can be used to drive other electrical systems.

CHAPTER TWO

PROXIMITY TRANSDUCERS

2.0 Synopsis

A brief survey of proximity transducers is presented here with particular emphasis on metal detector. The relative merits of the inductive type selected for the Traffic Detector are enumerated.

2.1 Introduction

In many industrial and everyday situations need often arises to sense the presence and the spatial behaviour of a body within a reference system. Such information is usually employed to determine control actions.

The transducers which have found wide applications include mechanical, pneumatic and electrical types as outlined in chapter one. For a sensor to be useful, its characteristics[7] such as linearity, sensitivity, repeatability and maximum error, must be known. There are many electrical sensors in use today and each exploit the changes in one of several possible electrical properties of matter that can be influenced by the proximity of another body. Such properties include

(a) Photo-electricity (b) Resistance (c) Capacitance
(d) Inductance and (e) Electrolytic

The choice of property used in measurement system depends on the type of sensor application contemplated. Each type of transducer exhibits specific characteristics which can then dictate the mode of application. The inductance property is selected here because of its suitability and installation simplicity.

2.2 The Magnetic Induction Proximity Transducer

The change in inductance of a magnetic coil due to the physical presence of a body can be exploited for the detection and measurement of physical quantities involved. Usually, the variation of inductance is used to change a parameter of an electric circuit which includes the variable inductance coil.

The variation of inductance of a coil can be achieved in a number of ways.(8)

(a) By changing the geometry of the coil, for example the overall length of the coil.

(b) By changing the effective permeability of the magnetic core material.

(c) By changing the reluctance of the magnetic path as in the variation of air gap of a magnetic circuit.

(d) By changing the coupling between two or more elements as in the LVDT. (Linear Voltage Dependent Transformer).

Many transducers have been designed using the above principles usually with an associated electric circuit to allow the display or conversion to measurable quantities. Such circuits which are used to convert inductance changes to other electrical readout quantities include the -

(i) simple a.c. potentiometer (ii) a.c. bridge

(iii) tuned oscillator (iv) transformer type sensors.

The change of inductance in this Inductive Proximity/Presence transducer is due to change in permeability of the magnetic coil and eddy current action when a conducting body comes into its proximity. A tuned Colpitts oscillator is used to interprete the change for electrical readout. The circuit was

selected because of its advantages over the rest for this particular application. This can be seen as circuit (i) above is simple but gives non-linear problem for small loads, (ii) presents balancing problems leading to circuit complexity and (iv) often needs a dummy transducer to annul the back-off output current at quiescence. Though method (a) above is simple, it provides limited flexibility and latitude for the accommodation of environmental variation.

Method (b) not only suffers from the above mentioned constraints, but it is also difficult to achieve a balance adjustment without the use of expensive components.

Method (d) would need more than a coil and consequently the influence of self-capacitance on circuit performance becomes more prominent.

The mechanism for inductance change can be illustrated as follows. The inductance of a coil is given as

$$L = \mu_o \mu_r \frac{A N^2}{l} \text{ Henrys}$$

where

μ_o = permeability of free space

= $4\pi \times 10^{-7}$ Henry/meter

μ_r = Relative permeability of the coil core

A = Cross-sectional area of the magnetic path in square meters

N = Number of turns

l = length of the magnetic circuits in meters

When a magnetic material is introduced into the core of the coil, the relative permeability changes and consequently the inductance of the coil changes. Another phenomenon which contributes to the change in inductance is the eddy current [7,8,9] induction. Consider a long straight conductor carrying an alternating current of constant amplitude. It is well known that it will set up an a.c. magnetic field concentric with its axis. This field, by Maxwells concept, has an associated a.c. electric field and consequently induces an eddy current in any conducting body in its neighbourhood. The flux due to the eddy current in the conducting body opposes the main flux. Since the inductance is given by

$$L = \frac{\partial \phi}{\partial I}$$

where

N = *number of turns*

ϕ = flux

I = current

there is an effective reduction in the flux and hence the inductance drops.

These two phenomena producing inductance change appear to counteract each other. However, the change in inductance due to permeability change is very negligible as the coil is buried inside the earth and the axle metal of the vehicle is reasonably far away from its core. However, the axle metal is close enough to the coil such that the eddy current effect plays a greater role in determining the direction of the inductance

change. Practical results confirm this as the frequency of an LC oscillator rises when a vehicle crosses. The frequency of the oscillator may be expressed as

$$f = \frac{1}{2\pi}\sqrt{\frac{1}{LC}}$$

which shows that a drop in inductance for a constant capacitance will produce a rise in frequency, f.

2.3 Tuned Oscillator Induction Detector

The change in coil inductance caused by the presence of metallic objects has been used to build a class of sensors called Metal Detectors. The variable inductor forms part of a tuned oscillator circuit and the change in frequency caused by the change in inductance can be interpreted by several techniques which include:

(a) the Beat Frequency Oscillator (BFO)[10]

(b) Induction Balance

(c) Phase logic

(d) the Phase-Locked-Loop technique, which is the method used in this project.

In the Beat Frequency Oscillator (BFO) technique, fig. 2.1 two oscillators are used - one the search coil oscillator (SCO), (usually of the LC tuned type) and the beat frequency oscillator.

The two oscillators are tuned such that there is a small frequency difference between them, say about 10 Hz. The outputs are then mixed

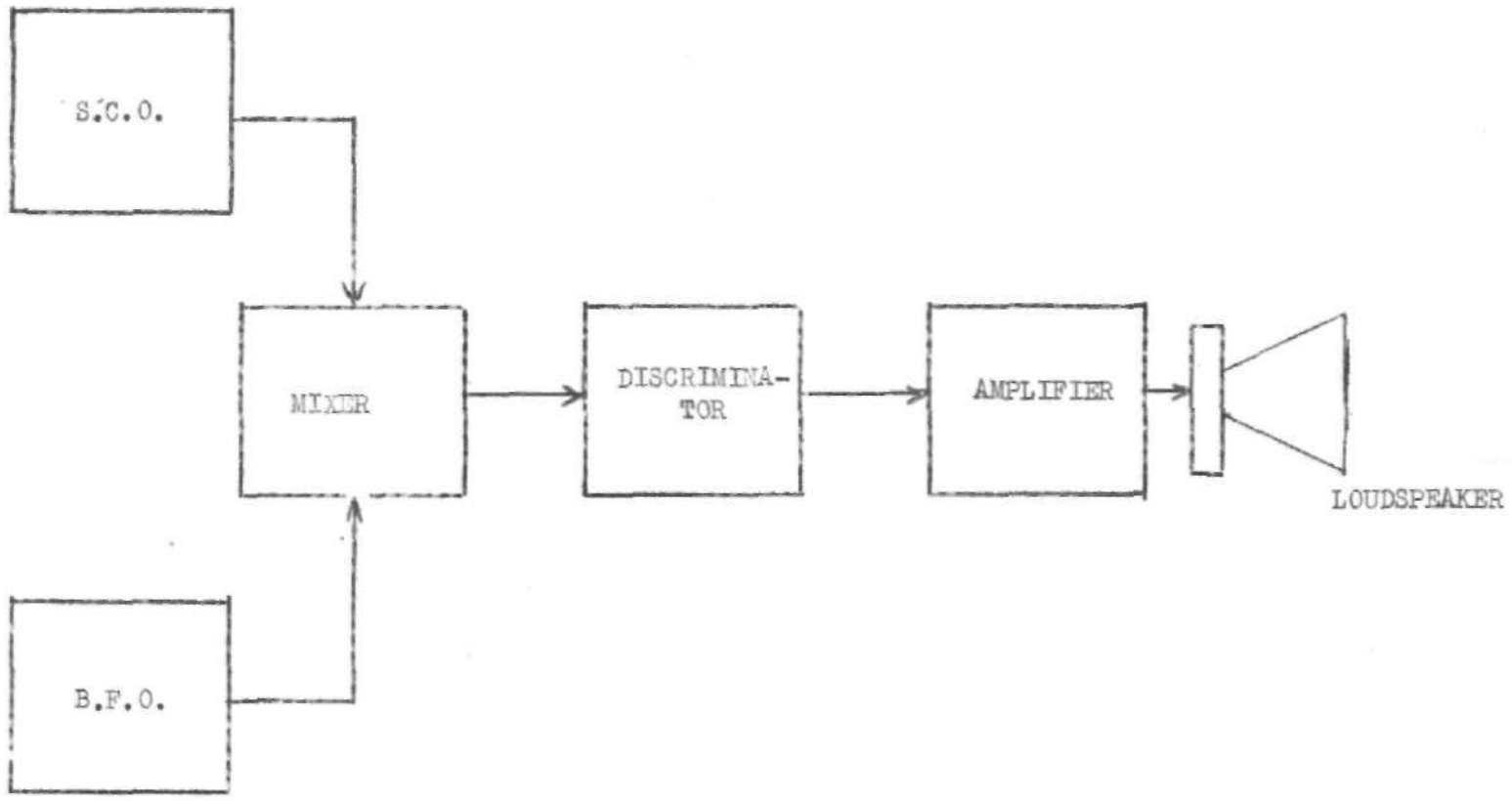


Fig. 2.1: The B.F.O. Metal Detector

together and the difference is selected by a suitable low pass filter. The difference frequency produces an audio tone after amplification. When the search coil is brought near a metal, its associated oscillator frequency (S.C.O) changes hence causing a change in the difference frequency and giving a different audio tone.

The principal disadvantage of this method is that the oscillator frequencies must be very stable and must not drift due to external variations. This type of stability is usually difficult to achieve at low cost leading to poor sensitivity of the method.

Another method which is often employed to provide improved sensitivity is the Induction Balance, fig.2.2. In this method, two coils with virtually no coupling between them are used in the search head. A modulated radio frequency signal is fed into one coil while the other serves as the receiver. When a metallic object is brought near the search head, eddy currents are induced in the object and subsequently an a.c. field is induced according to Lenz's Law which now off sets the initial induction balance between the two coils. The receiver coil then picks up a signal which is amplified and demodulated. This method while providing a better sensitivity, is noisy and difficult to adjust for zero balance.

The comparison and detection of the phase difference of two oscillator frequencies could also be used to sense the proximity of a metallic body. This method is similar to the B.F.O. type in thattwo oscillators are needed. However, in this case the search coil oscillator and the reference oscillator

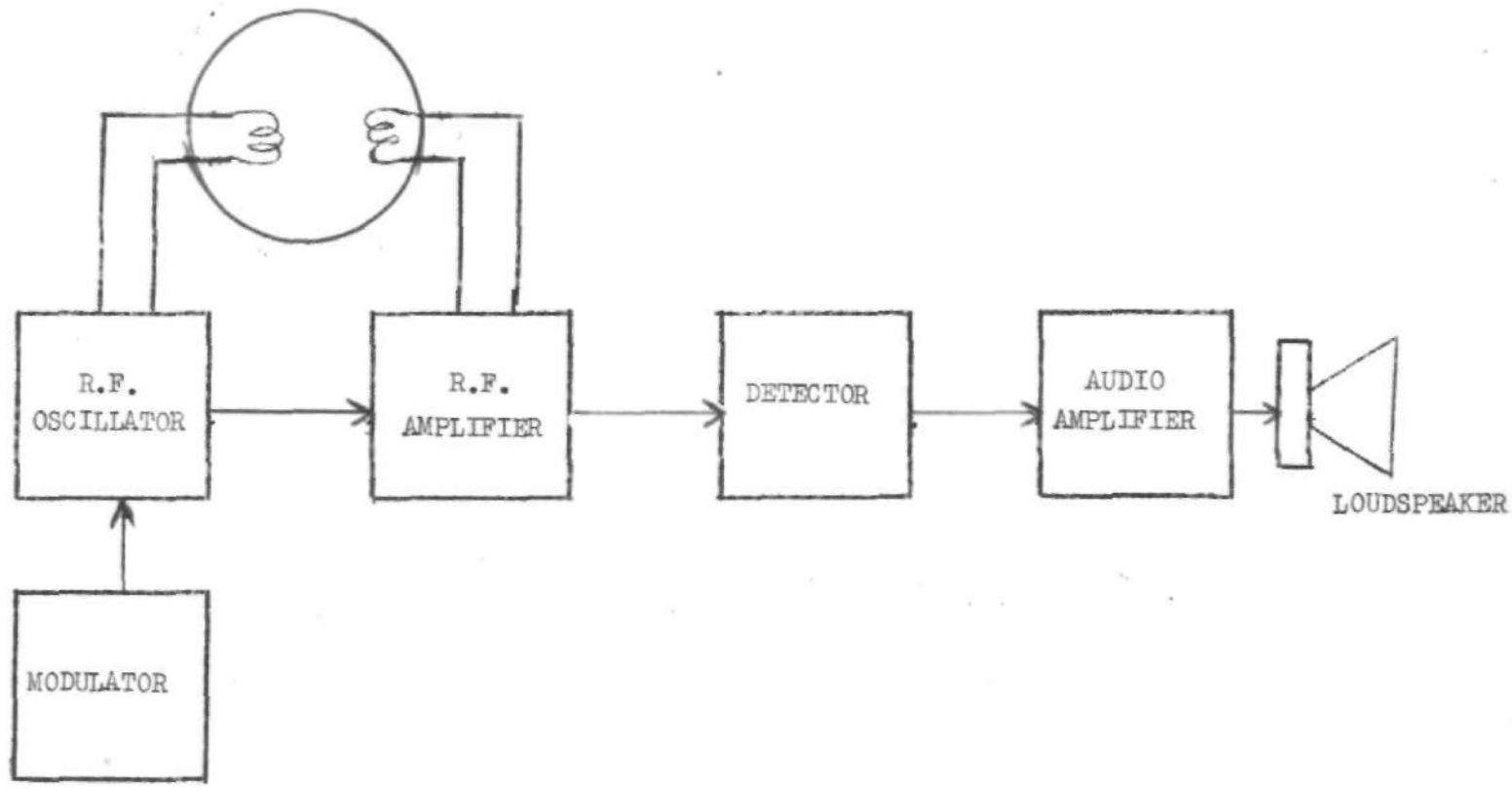


Fig. 2.2: The Induction Balance Metal Detector

are initially tuned to the same frequency. The outputs of these oscillators are compared by a phase comparator. When two frequencies differ from each other by say Δf, there is a phase difference $\Delta \phi$ in time T between them given by (11)

$$\Delta\phi = 360\ \Delta T\ \Delta f.$$

The detection of this phase difference has been used to locate metals with light emitting diode (L.E.D.) readout. This is phase logic method.

2.4 The Phase Lock Loop Proximity Detector

In both the B.F.O. and the phase logic methods, two stable oscillators are required to enhance good sensitivity. The level of stability required is not usually easy to achieve with cheap oscillators since this involves identical drift characteristics with the two oscillators.

In the induction balance method, apart from the need of two coils, it is very difficult to establish a zero offset in quiescence. A method which solves these difficulties and simultaneously provides for sensitivity and reliable metallic presence detection employs the phase lock loop technique.

A phase lock loop is an electronic feedback system which consists of a phase detector, low pass filter and a voltage controlled oscillator. Its feedback nature gives it the ability to adjust to environmental variation yet without loss of accuracy and sensitivity. Environmental

variation effect on the detector is even more pronounced for the Traffic Detector where the loop is buried in the road. Only one coil is needed to build the Detector Oscillator (D.O). The second oscillator is in-built in the loop and hence its stability is greatly improved. It possesses the ability to track small drifts in the D.O. due to environmental variations with no sacrifice in the noise immunity characteristics. It is relatively easy to build and the components are available now either as a single chip loop e.g. Signetics NE 565 or as a separate building blocks.(36)

The Detector Oscillator is tuned to the same frequency as that of the voltage controlled oscillator of the PLL. This frequency is also the centre frequency of the loop and a capture range is defined to suitably accomodate drifts in the Detector Oscillator.

The capture range can be controlled by the selection of the low pass filter and consequently allowance can be made for any predicted drift in the D.O.

The output is a d.c. signal which is also the error voltage that forces the V.C.O. to synchronise with the loop oscillator. The capture range is inversely proportional to the overall gain or sensitivity of the loop. It is usually desirable to reduce the capture range to get better sensitivity. This is done either by using a large time constant filter or/and by employing the Narrow Band Operation technique. Further details are discussed in the ensuing chapters.

2.5 Summary

A survey of proxinity detectors has been outlined in this chapter. The possible choice of electrical property for use has been given. An insight has also been given into the use of frequency modulation in Treasure Finding. The various methods for such Locators have been compared and the justification of the selected method for the Traffic Detection application has been outlined.

CHAPTER THREE

THE DETECTOR OSCILLATOR

3.0 Synopsis

This chapter discusses the Detector Oscillator. The construction of the Loop Coil is presented. The performance of the oscillator so designed using this coil under varying weather conditions are also enumerated.

3.1 Introduction

The Loop Detector

This is the main sensing component of the Vehicle Detector which must be properly designed and installed for reliable operation. The inductive loop is made part of an oscillator, the Detector Oscillator, D.O. In Metal or Treasure hunting applications, this loop coil, termed the Search Coil is usually much smaller and mobile. It is moved from point to point to detect buried 'Treasures'. In this application, as in many other industrial applications, it may be fixed. The principle of operation is however, the same in that a conducting body coming into the neighbourhood of a current carrying coil alters the coils inductance and induces eddy currents. The coil in this application is buried giving rise to problems that are unique to the application. The problems will be discussed in a later part of this chapter.

3.2 Construction of the Loop Detector

The structure of the Loop Detector must satisfy both physical and electrical conditions. The specifications of the coil includes its length, width, number of turns and wire diameter.

The physical dimension must be such that it is possible to discriminate between vehicles on opposing sides of the road. This imposes a constraint

on the width of the coil. A study of the width of single carriage roads around Zaria, Kaduna State, Nigeria shows that it varies from about 3.0 meters for local roads to about 3.66 meters for Urband primary roads. In addition, the average widths of vehicles (rim to rim) of some vehicles plying Zaria roads are shown in Table 3.1.

	Vehicle Type	Rim-to-Rim Width (meters)
1	Morris Minor	1.38
2	Cortina (Ford)	1.42
3	VW Pasalt	1.48
4	VW 1500 cc	1.48
5	Lada	1.49
6	Peugeot 404 Saloon	1.49
7	Morris Marina	1.50
8	Peugeot 504 L	1.56
9	Toyota Dyna	1.57
10	Opel Record	1.58

Table 3.1: Average Width of some Vehiceles

From these considerations, a dimension of 0.6 x 2.0 meters was used in the design. The choice of the number of turns is subject to two constraints. The frequency of operation of LC-Oscillators is fixed generally by the

inductance, L and capacitance C according to the equation

$$f = \frac{1}{2\pi\sqrt{L\,C}} \quad \ldots\ldots \qquad 3.1$$

which implies that the change in frequency is given by

$$df = \left(\frac{\partial f}{\partial L}\right)_C dL + \left(\frac{\partial f}{\partial C}\right)_L dC \quad \ldots\ldots \qquad 3.2$$

where $\left(\frac{\partial f}{dL}\right)_C$ is the partial differential of frequency with respect to the inductance when the capacitor is held constant.

A large number of turns will give stronger field and larger change in frequency when the Vehicle crosses. However, the larger the number of turns, the smaller the value of circuit capacitance that can be used. A large number of turns will in general increase the self capacitance of the coil which is less predictable and subject to strong variations in environmental conditions.

The inductance and consequently the number of turns is determined in part by the optimum expected self capacitance to circuit capacitance ratio. This is discussed later in this chapter. Generally, a few number of turns is used and the sensitivity is compensated for by the use of narrow band operation of the phase-lock-loop.

The wire guage is selected by considering the skin effect[8,9] which is important at the operating frequency. The skin effect becomes appreciable

when the ratio of diameter to skin depth is equal to 2.0. The depth of penetration for a copper conductor of circular cross section is

$$\delta = \frac{6.6}{\sqrt{f}} \text{ cm} \qquad \ldots\ldots\ldots\ldots \quad 3.3$$

Therefore the wire diameter at a frequency of f (Hz) is

$$d = \left(\frac{d}{\delta}\right) \frac{6.6}{\sqrt{f}} \text{ cm} \qquad \ldots\ldots\ldots\ldots \quad 3.4$$

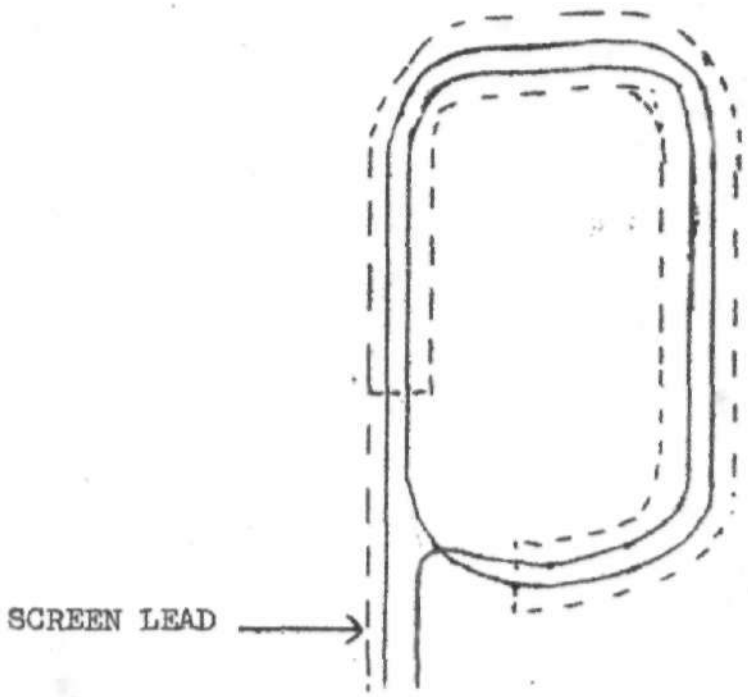
for an operating frequency of 100 kHz a standard Wire Gauge No. 27 with a diameter of 0.417 cm is adequate.

The turns of this rectangular coil are held together by an insulating tape. A strip of Aluminium foil is wound round with a gap at one end - see fig. 3.1. This serves as a screen which stabilizes the effect of electrostatic field on the coil thereby stabilising the effect of the stray capacitance . The introduction of the screen increases the self capacitance of the coil and reduces its inductance due to the eddy currents induced in it.

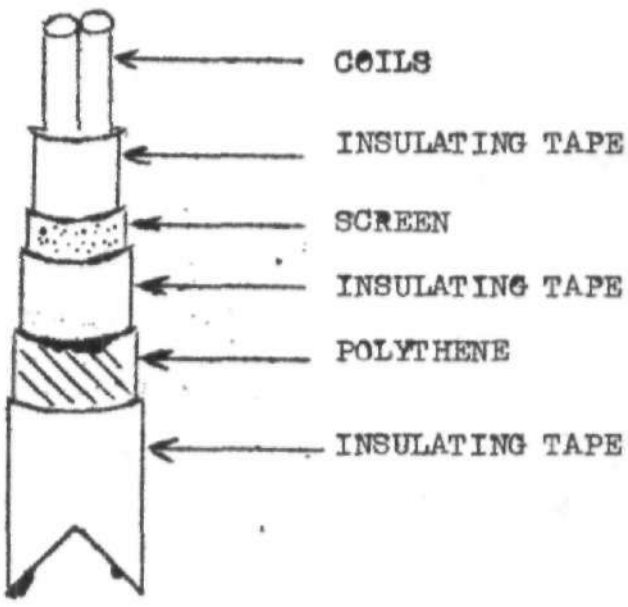
The response of the coil to the Vehicle presence is not significantly affected since the depth of penetration of the signal at 100 KHz in Aluminium is much greater than its thickness.

For Aluminium, the depth of penetration is given by

$$\lambda = \sqrt{\frac{2}{\omega \eta \mu \sigma}} \qquad \ldots\ldots\ldots\ldots \quad 3.5$$



(a)



(b)

Fig 3.1: CONSTRUCTION OF THE LOOP COIL

where

ω = angular frequency in radians

η = wave impedance

μ = absolute permeability of medium

σ = conductivity of Aluminium

or $$\delta = \frac{8.3}{\sqrt{f}} \quad \ldots\ldots\ldots\ldots \quad 3.6$$

Therefore the depth of penetration for aluminium at 100 KHz is 0,3 mm whereas the thickness of the foil used is .02 mm.

The screened coil is covered with insulating tape. For improved operation the coil must be insulated from moisture. This is a difficult task. Water molecule migration will affect the oscillating frequency stability. This is one of the reasons why a phase lock loop discriminator is selected for this Detector. The coil is wrapped with polythene. The construction is shown in Fig. 3.1. The whole assembly is buried in a small slot cut on the road surface. The depth of the slot should be such that it is well above the water table and close to the top so that there can be greater interaction between the electromagnetic field and the rim of the Vehicle. About 4 cm below the asphalt is suitable. A more expensive way of installation can however be applied to check the influence of water on the coil characteristics. For instance, the slot could be sealed with cement or/and asphalt.

The resultsoof measurements as the coil is treated is shown in Table 3.2.

Treatment	Frequency (KHz)	Percentage Drop in Frequency	Stability °/oo per deg. C Ambient
1. Coil alone	141.6	0	0.26
2. Screened	132.05	6.7	0.25
3. Screened & grounded	126.05	11.0	0.16
4. Covered with Polythene	119.38	15.7	0.34
5. Buried on the road	92.66	34.6	7.0

Table 3.2: Treatment of the Coil, and Performance Measurements

The screen has increased the self capacitance, and so decreased the oscillating frequency. Further compression of the screen due to taping further increases the self capacitance. In the following sections, some oscillator structures considered for the Loop Detector Oscillator will be discussed.

3.3 The Detector Oscillator: General

Since the variation in the frequency of the Detector Oscillator, (D.O.) is the primary indication for the presence of a Vehicle, the D.O. is a very important part of this type of Vehicle Detector. The oscillator must

possess several important characteristics. The most important is that it must include the detector coil whose characteristic (inductance) can be influenced by the Vehicle presence in the neighbourhood.

A large number of oscillators can be defined using the Detector Coil as one of the frequency selecting elements. Usually, a capacitor is used as the other frequency selecting component in order to achieve a stable and variable oscillating frequency. In some restricted instances, the self-capacitance of the coil can be used but this leads to very unstable and troublesome oscillations. The family [14,17] of Inductance-Capacitance, LC Oscillator is shown schematically in fig. 3.2. This can be broadly classified into relaxation and harmonic oscillators. The harmonic class whixh is of interest in this investigation can be broken into two types - the negative resistance and feedback oscillators.

There is not much distinction between the two in actual sense. A negative-resistance oscillator is a one-pair terminal device which includes an active component with a part of its voltage-current characteristic 'negative in slope'. Such active devices include the Tunnel diode [17,18]

For oscillation to persist, it is required that the negative resistance be equal or greater than the resistance of the associated passive network. A useful method of determining oscillation conditions is by polar plots. The polar plots are plotted to the same scale on the same axis and the point of intersection of the two curves is an indication of oscillation conditions. This method is used for the analysis later on in the chapter.

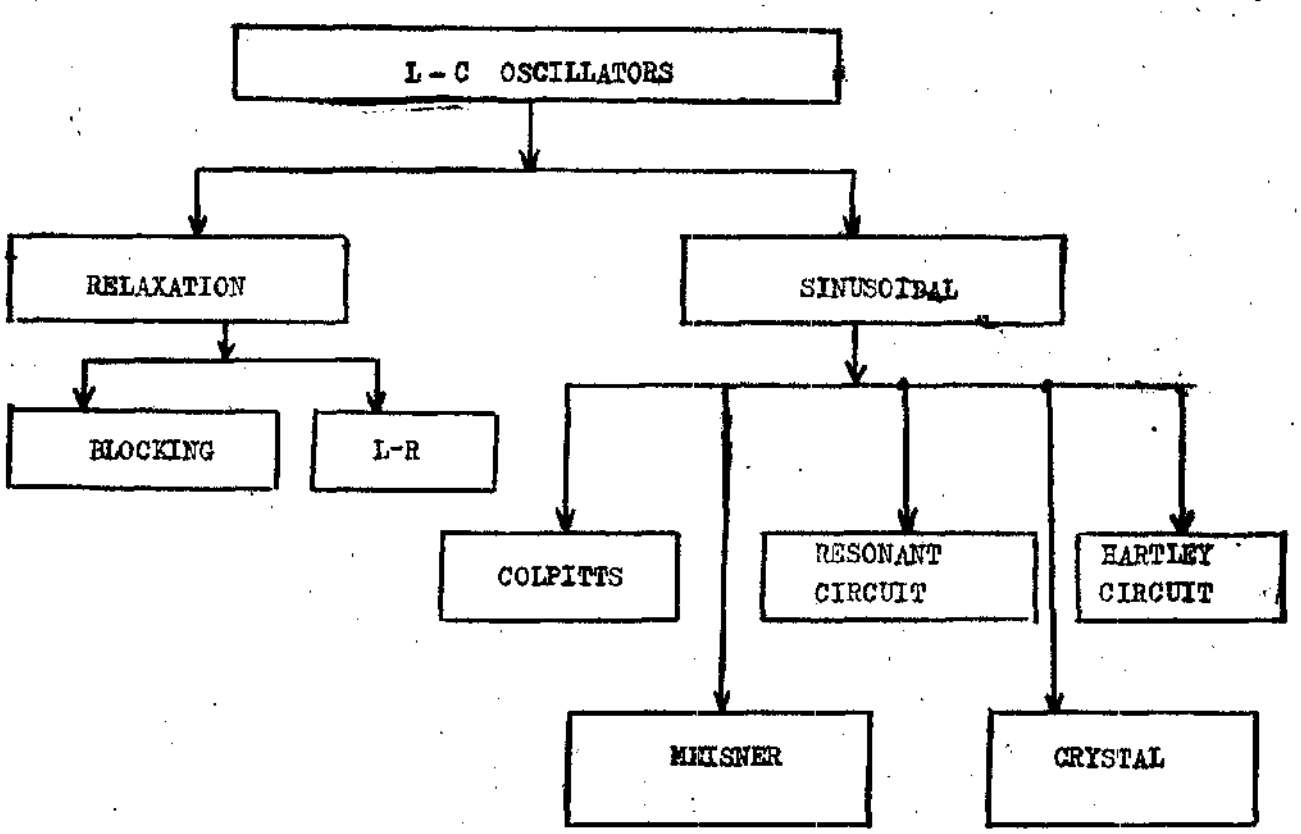


FIG. 3.2: FAMILY OF L C OSCILLATORS

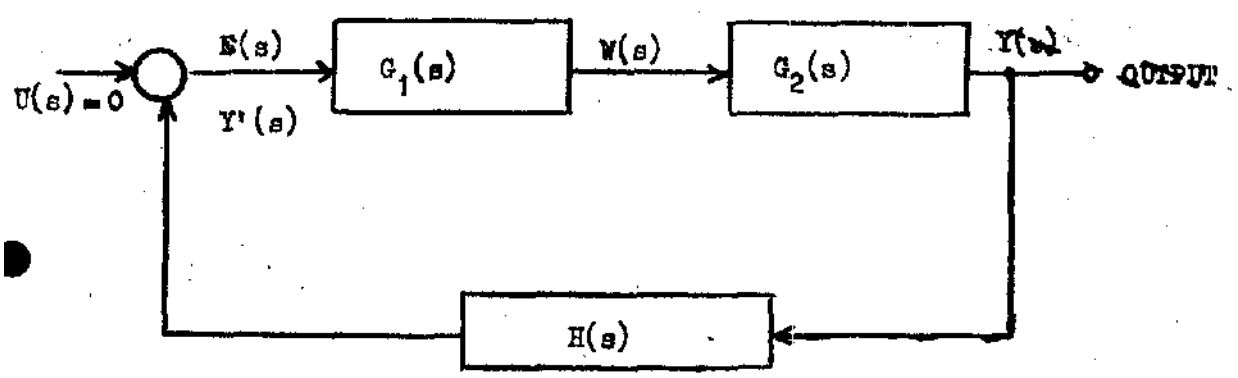


FIG. 3.3: OSCILLATOR AS A FEEDBACK SYSTEM

There are two-terminal pair devices such as transistor and triodes which could be used to build oscillators of the feedback type. The principle is illustrated in fig. 3.3. which involves positive feedback. A portion of the output is fedback into the input with the same phase as the input. Hence the oscillator is a self-fed amplifier converting d.c. into a.c. power.

The d.c. power drain is very inportant particularly if the Vehicle Detector is to be operated from a battery. The forward path elements, $G_1(s)$ and $G_2(s)$ of fig. 3.3 may not be always seperatle as in transistor oscillators. In most cases the feedback element H(s) is usually unity. The closed loop transfer function is given by (15)

$$T(s) = \frac{G_1(s)\ G_2(s)\ H(s)}{1 + G_1(s)\ G_2(s)\ H(s)} \qquad \ldots\ldots\ldots \qquad 3.1$$

The stability of the systen is determined by the characteristic equation

$$G_1(s)\ G_2(s)\ H(s) + 1 = 0 \qquad \ldots\ldots\ldots\ldots \qquad 3.2$$

If the polar plot of equation 3.2 encloses the critical plot $-1 + j0$ oscillation is probable. The point of oscillation is the intercept of the root-locus with the imaginary axis.

The application of these analysis criteria to the specific oscillators design is presented in the following sections. Many possibilities for oscillators are known but only two are presented in this thesis.

3.4 The Transistor Colpitts Oscillator

A Transistor Colpitts Oscillator was designed and tested with the Detection Coil. The common emitter transistor colpitts oscillator is shown in fig. 3.4 with its equivalent circuit. The junction capacitances are indicated neglecting the high frequency interbase resistance.

C_{bc} = Base collector junction capacitance

$C_{c'e}$ = Collector emitter junction capacitance

$C_{b'e}$ = Base emitter junction capacitance.

Due to the low impedance of the transistor input, the method of analysis is to load the output with the input impedance and break the feedback loop at point X. A test voltage V_1 is applied which gives a base current of V_1/h_{ie}. The collector current is $\frac{h_{fe}}{h_{ie}} V_1$

Applying standard theorems leads to fig. 3.4 c where

$$Z_1 = h_{ie} \,//\, X_{c1}'$$

$$Z_2 = X_{c_2}' \,//\, \frac{1}{h_{oe}} \qquad \ldots\ldots\ldots 3.3$$

$$Z_3 = X_L \,//\, X_{c3}$$

and

$$X_{c1}' = \frac{1}{j \omega C_1'}$$

$$X_{c3} = \frac{1}{j \omega C_3}$$

$$X_L = j \omega L$$

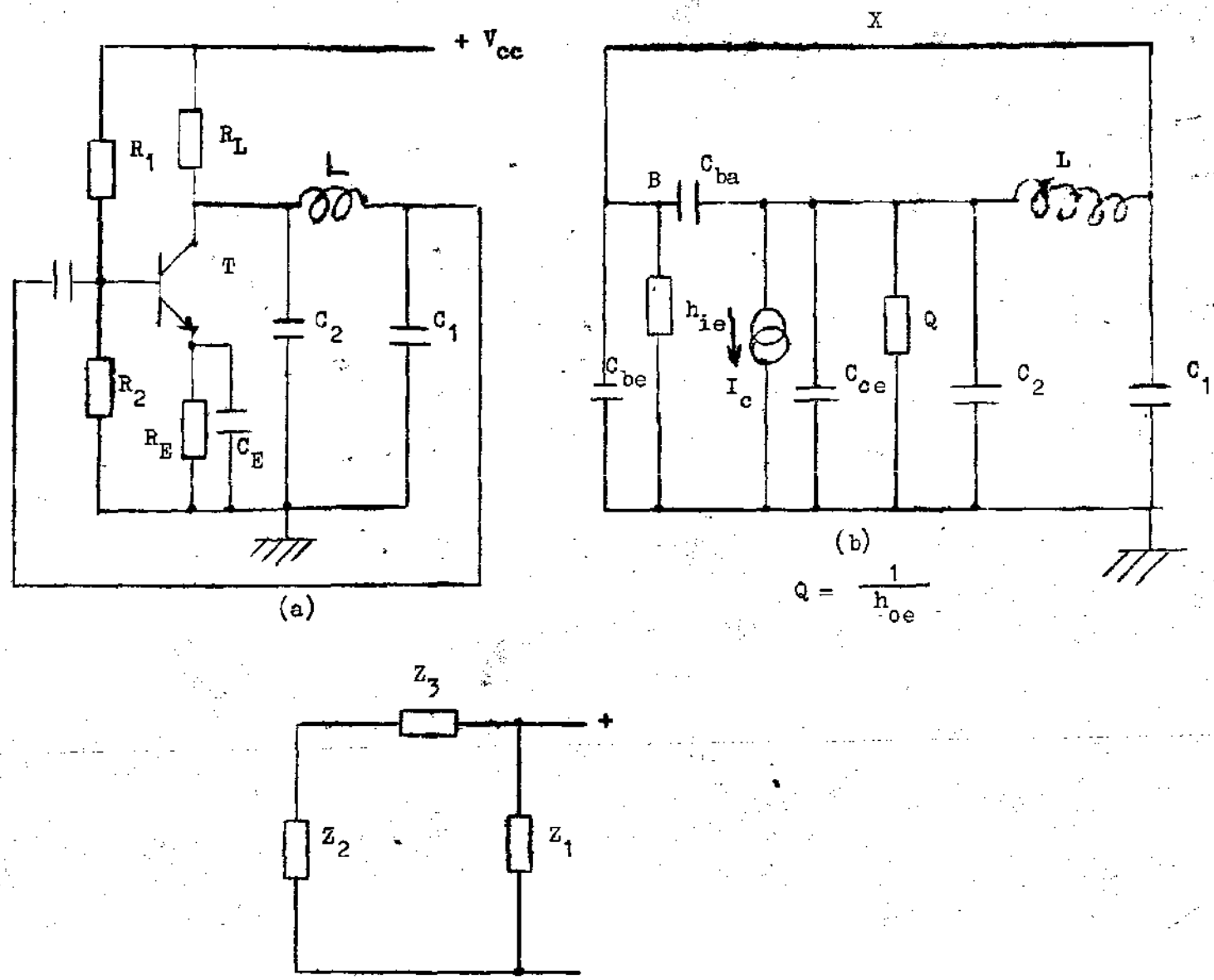


Fig. 3.4: TRANSISTOR COMMON EMITTER COLPITTS OSCILLATOR.-

$$X'_{c2} = \frac{1}{j\omega C_2'}$$

$$C'_1 = C_1 + C_{be}$$

$$C'_2 = C_2 + C_{de}$$

$$C_3 = C_{bc} + C_s$$

where C_s is the self capacitance of the coil.

The output voltage V_o in fig. 3.4 c is given by

$$V_o = \frac{Z_1}{Z_1 + Z_2 + Z_3} \; gm \; V_1 \; Z_2 \qquad \ldots\ldots\ldots \qquad 3.4$$

or

$$\frac{V_o}{V_1} = \frac{Z_1 \; Z_2 \; gm}{Z_1 + Z_2 + Z_3} \qquad \ldots\ldots\ldots\ldots \qquad 3.5$$

From equation 3.2, the transfer function in equation 3.5 must be equal to -1

That is

$$Z_1 + Z_2 + Z_3 = - Z_1 \; Z_2 \; gm \qquad \ldots\ldots\ldots \qquad 3.6$$

If the values in equation 3.3 are substituted into equation 3.6, then

$$1 + j\omega h_{ie} C'_2 + h_{ie} h_{oe} + j\omega h_{ie} C_i + \frac{j\omega L}{1 - \omega^2 L C_3} \left((h_{oe} + j\omega C'_2 + j\omega h_{ie} h_{oe} C'_1 - \omega^2 h_{ie} C'_1 C_2') \right) = - gm \; h_{ie}$$

Equating the imaginary parts of equation 3.7

$$h_{ie}(C'_2 + C_1)\frac{L}{1-\omega^2 L C_3}(h_{oe} - \omega^2 h_{ie} C'_1 C'_2) = 0 \quad \ldots\ldots 3.8$$

$$\therefore \omega^2 = \frac{h_{oe}/h_{ie}}{(C'_1 + C_2')(C_3 + \frac{C'_1 C'_2}{C'_1 + C'_2})} + \frac{1}{L(C_3 + \frac{C'_1 C'_2}{C'_1 + C'_2})}$$

$$= \frac{1}{C_3 + \frac{C_1' C_2'}{C_1' + C_2'}}\left(\frac{h_{oe}/h_{ie}}{C'_1 + C'_2} + \frac{1}{L}\right) \quad \ldots\ldots 3.9$$

Equating the real parts of equation 3.7 gives

$$1 + h_{ie} h_{oe} - \frac{\omega^2 L C'_2 + \omega^2 L h_{ie} h_{oe} C'_1}{1-\omega^2 L C_3} = -\beta h_{ie} \quad \ldots\ldots 3.8$$

Since $h_{fe} = \beta h_{ie}$

$$\therefore \omega^2((h_{fe} - h_{ie} h_{oe} - 1)(C_3 - C'_2 - h_{ie} h_{oc} C'_1)$$

$$= -(1 + h_{fe} + h_{ie} h_{oe}) \quad \ldots\ldots\ldots\ldots\ldots\ldots 3.9$$

If $h_{ie} h_{oe} \ll h_{fe}$ 3.10

(which is true for BC 108 (19) where h_{ie} = 2.7 kilo ohms, $h_{oe} = 18 \times 10^{-6}$ micro mho at I_c = 2 mA and h_{fe} is about 150),

then

$$\omega^2 L(h_{fe} C_3 - C'_2) = -h_{fe} \quad \ldots\ldots\ldots\ldots 3.11$$

Substituting

$$\omega^2 = \frac{1}{\left(C_3 + \frac{C'_1 C'_2}{C'_1 + C'_2}\right) L}$$

from equation 3.9 by neglecting the first term inside the bracket of the right hand side.

$$\therefore \quad h_{fe} \geq \frac{C'_2}{C'_1} \qquad \ldots\ldots\ldots\ldots \qquad 3.12$$

3.5 Design Criteria

In this section, some design inequalities and considerations for a Transistor Colpitts Oscillator are presented. The specifications for the oscillator are the

Operating frequency	=	100 KHz
d.c. power supply	=	9.0 volts

The coil is constructed as shown in an earlier section. The inductance of the coil was first of all made equal to 340 microhenrys. This value did not allow the use of a large circuit capacitance that will invalidate the variation of the self capacitance on the frequency stability.

The inductance was then reduced to about 100 microhenry by reducing the numbers of turns.

It is necessary to select a transistor which can provide satisfactory operation. Satisfactory operation implies high h_{fe}, alpha-cutoff, f_T and low inter-electrode capacitances. Transistor BC 108 was selected with the following characteristics: (32)

$h_{fe} \triangleq 220$ at collector current of 2 mA

$h_{ie} \triangleq 2.7$ kilo ohms at collector current of 2 mA

$h_{oe} < 30$ micro mho at collector current of 2 mA

Collector capacitance, $C_{b'c} < 6$ pF at 1.0 MHz

Emitter-base Capacitance, $C_{b'e} = 8$ pF at 10 MHz

Gain bandwidth Product, $f_T > 150$ MHz at collector current of 10 mA.

A collector amount of 6.5 mA was selected which will allow the oscillator to develop enough power for the expected load

The capacitor: values are found by considering equation 3.12 and the equation

$$f = \frac{1}{\sqrt{\frac{L\, C_1 C_2}{C_1 + C_2}}}$$

which is a practical starting point assuming that the junction capacitances are negligible together with the self capacitance of the coil. The capacitor type should be selected by considering the temperature coefficient of frequency of the oscillator.

The temperature coefficient of frequency is related to the temperature coefficient of inductor and capacitance since

$$f = f(C_1, C_2, L, h_{fe})$$

If $f = \frac{1}{2\pi\sqrt{LC}}$ 3.13

(33)

Then change in frequency is given by the partial differential equation.

$$df = \left(\frac{\partial f}{\partial L}\right)_C dL + \left(\frac{\partial f}{\partial C}\right)_L dC \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad 3.14$$

Since L and C also varies with temperature

$$\frac{df}{dT} = \left(\frac{\partial f}{\partial L}\right)_C \frac{dL}{dT} + \left(\frac{\partial f}{\partial C}\right)_L \frac{dC}{dT} \quad \ldots\ldots\ldots\ldots\ldots \quad 3.15$$

where

$\left(\frac{\partial f}{\partial L}\right)_C$ is the partial differential of f with respect to L when C is constant.

$$\left(\frac{\partial f}{\partial L}\right)_C = -\frac{f}{2L} \quad \ldots\ldots\ldots\ldots\ldots\ldots \quad 3.16$$

and

$$\left(\frac{\partial f}{\partial C}\right)_L = -\frac{f}{2C}$$

$$\therefore \quad \frac{1}{f}\frac{df}{dT} = -\frac{1}{2}\left(\frac{1}{C}\cdot\frac{dC}{dT} + \frac{1}{L}\frac{dC}{dT}\right) \quad \ldots\ldots\ldots \quad 3.17$$

$$\alpha_f = -\frac{1}{2}(\alpha_C + \alpha_L) \quad \ldots\ldots\ldots\ldots \quad 3.18$$

where

α_f is the temperature coefficient of frequency

α_C is the temperature coefficient of capacitor

α_L is the temperature coefficient of inductor

The optimum design would require

$$\alpha_f \rightarrow 0.$$

In practice α_f is minimised by choice of capacitor types which can offset the coefficient of the inductance which is positive tending to reduce the right-hand side of equation 3.18.

For an oscillator using two capacitors C_1 and C_2 in series

$$C = \frac{C_1 C_2}{C_1 + C_2}$$

then

$$\left(\frac{\partial C}{\partial C_1}\right)_{C_2} = \left(\frac{C_2}{C_1 + C_2}\right)^2 \quad \ldots\ldots\ldots\ldots \quad 3.19$$

and

$$\left(\frac{\partial C}{\partial C_2}\right)_{C_1} = \left(\frac{C_1}{C_1 + C_2}\right)^2 \quad \ldots\ldots\ldots \quad 3.20$$

so

$$\frac{dC}{dT} = \left(\frac{C_2}{C_1 + C_2}\right)^2 \frac{dC_1}{dT} + \left(\frac{C_1}{C_1 + C_2}\right)^2 \frac{dC_2}{dT} \quad \ldots\ldots \quad 3.21$$

or

$$\alpha_C = \frac{C_2}{C_1 + C_2} C_1 + \frac{C_1}{C_1 + C_2} C_2 \quad \ldots\ldots\ldots \quad 3.22$$

If $C_2 \gg C_1$

then $\alpha_C \simeq \alpha_{C_2}$ $\ldots\ldots\ldots\ldots$ 3.21

Equation 3.21 shows that if

$$C_2 \gg C_1$$

capacitor C_2 will have a dominating influence on the frequency as temperature varies. Hence C_2 must be carefully selected.

However, α_C is not so simple since there is another 'capacitance' C_3 which represents the varying self-capacitance of the coil.

In fig. 3.5 is shown the variation of frequency with ground and Ambient temperature. 'The capacitance' C_3 is responsible for the fact that frequency variation did not follow ambient temperature faithfully.

Assuming the capacitors are in parallel, the temperature coefficient of capacitor can be found since now,

$$C = C_1 + C_2 + C_3 \quad \ldots\ldots\ldots \qquad 3.22$$

$$\frac{\partial C}{\partial C_1} = \frac{\partial C}{\partial C_2} = \frac{\partial C}{\partial C_3} = 1$$

So equation 3.21 becomes

$$\frac{dC}{dT} = \frac{C_1}{C_1 + C_2 + C_3} \cdot \frac{1}{C} = \frac{dC_1}{dT} + \frac{C_2}{C_1 + C_2 + C_3} \cdot \frac{1}{C_2} \frac{dC_2}{dT} + \frac{C_3}{C_1 + C_2 + C_3} \cdot \frac{1}{C_3} \frac{dC_3}{dT} \qquad \ldots\ldots\ldots 3.23$$

$$C = \frac{C_1}{C_1 + C_2 + C_3} \alpha_{C_1} + \frac{C_2}{C_1 + C_2 + C_3} \alpha_{C_2} + \frac{C_3}{C_1 + C_2 + C_3} \alpha_{C_3} \qquad \ldots\ldots 3.24$$

which shows the difficulty of stabilizing the frequency of the oscillator when C_3 is taken into account. The third term of equation 3.24 is responsible for the troublesome behaviour of the oscillator with humidity and temperature.

In the first place C_3 varies with humidity and temperature and the direction of variation is determined more or less by the humidity of the medium.

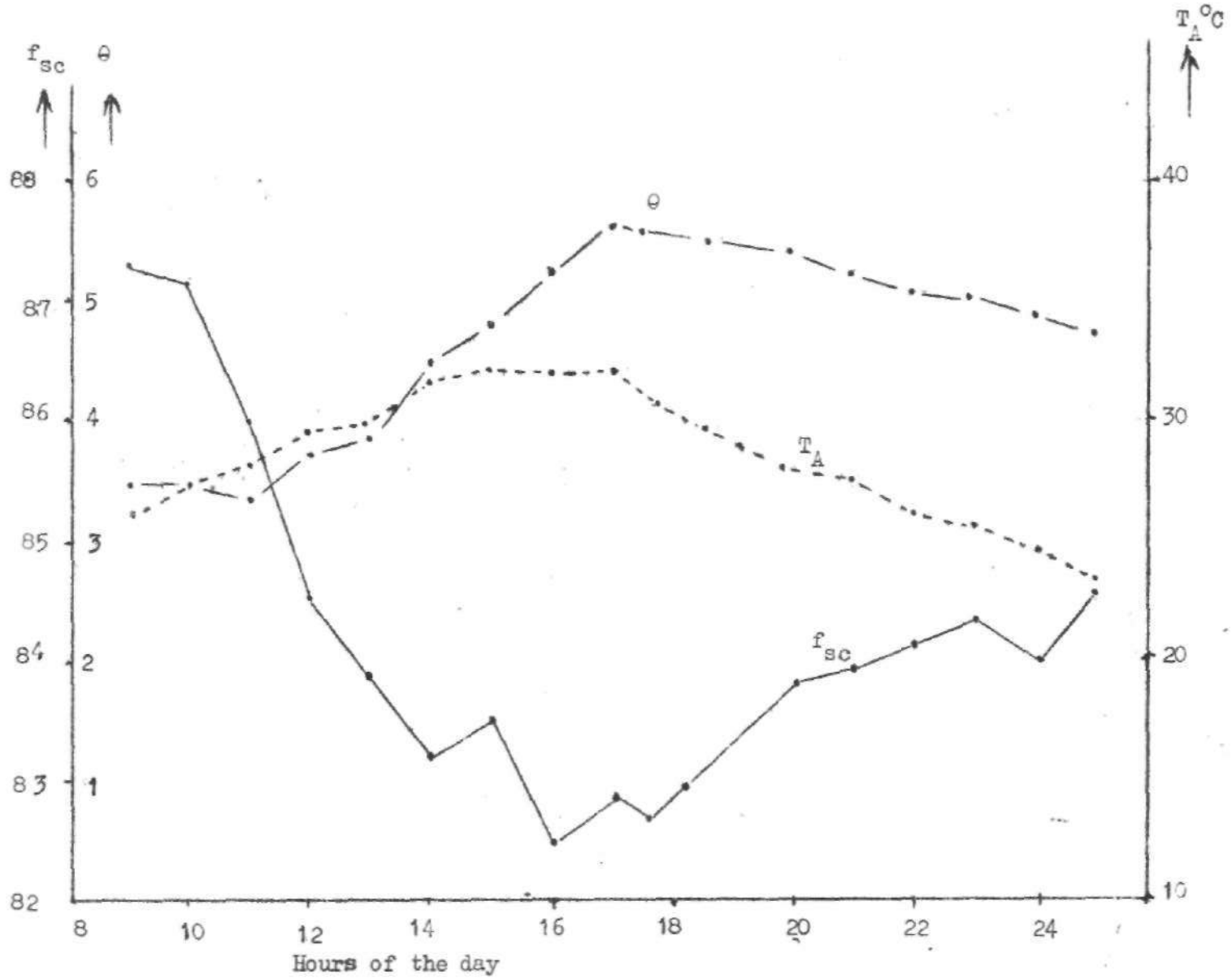


Fig. 3.5: Frequency Variation with Ground and Ambient Temperatures

It is therefore necessary to make

$$C_3 \ll C_1 + C_2$$

so that equation 3.24 reduces to

$$\alpha_c \simeq \frac{C_1}{C_1 + C_2}\alpha_{c_1} + \frac{C_2}{C_1 + C_2}\alpha_{c_2} \qquad \ldots\ldots\ldots \quad 3.25$$

and careful selection of C_1 and C_2 will minimise the frequency drift with temperature.

3.5 Colpitts Oscillator Using a Non-linear Element

A Colpitts oscillator built around an electronic relay element is shown in fig. 3.6. It consists of a nonlinear element and a frequency selecting network. The Describing Function[18,20,21,22] approach is used to analyse these types of oscillators. This approach is justified if the harmonics generated can be neglected without introducing appreciable error.

The Describing Function, D.F., is defined as an equivalent transfer function of the nonlinear system. This is used to replace the nonlinear system so that the entire system reduces to a linear one which could be analysed by standard linear control theory.

The nonlinear element selected here is an electronic relay with a deadband. This is the Transistor- Transistor Logic (TTL) inverter gate 7404 whose transfer characterisitc is shown in fig. 3.7. The characteristic shows that the output during logic '0' is greater than 0.0 volts. The

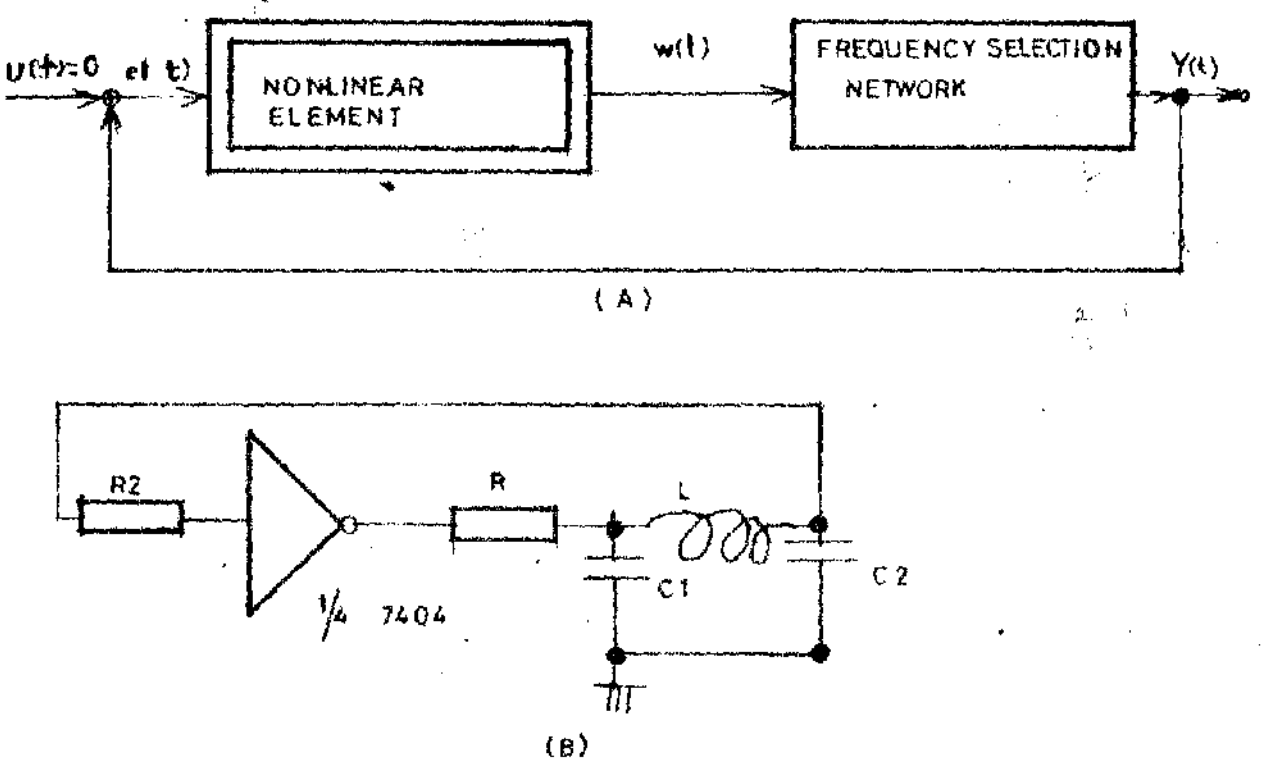


FIG 3 6 COLPITTS OSCILLATOR USING NONLINEAR ELEMENT

(A) BLOCK DIAGRAM (B) CIRCUIT DIAGRAM FOR 7404

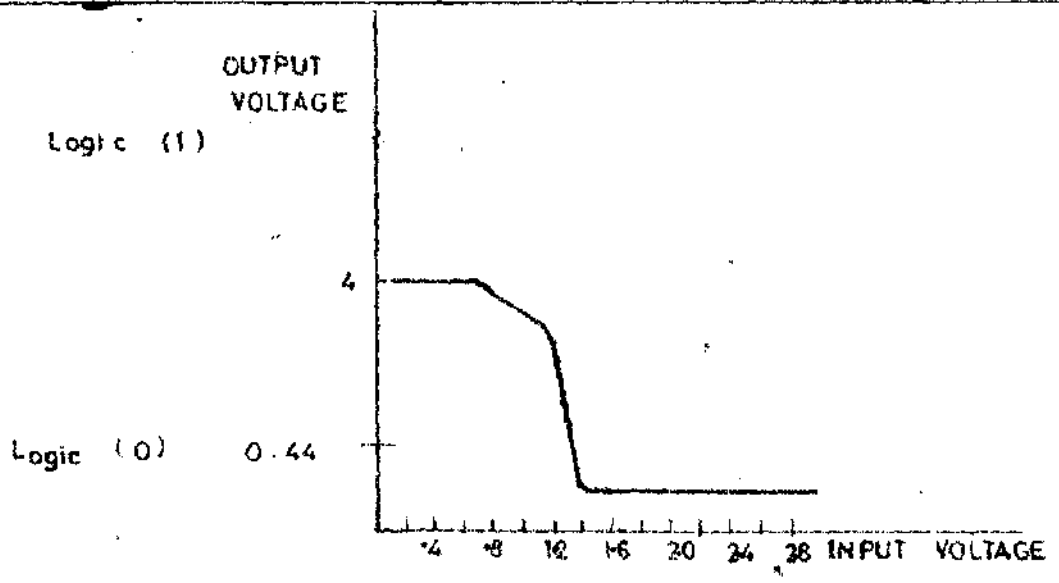


FIG 3.7 TRANSFER CHARACTERSTIC OF TTL 7404

system is not odd symmetric and so there is a d.c. signal which is fed round the loop, and back into the input of gate.

There are two inputs now to the non-linear element - a d.c. signal and a sinusoidal signal.

Consequently the usual assumptions made in the derivation of the D.F. that the d.c. component of the Fourier series is zero do not hold here. The analysis is therefore performed by the Dual Input Describing Function (D.I.D.F.) method.[26,27,28]

The D.I.D.F. for this nonlinear element can be derived by considering the diagram in fig. 3.8 which shows the input - output wave forms of the TTL 7404 gate. For convenience, the gate is treated as consisting of a buffer and an 'inverter'. The input to the gate is

$$e_o(t) = \alpha + \beta \sin \omega t \qquad \ldots\ldots\ldots\ldots \qquad 3.26$$

the deadband is

$$D = \alpha + \beta \sin \theta_1 \qquad \ldots\ldots\ldots\ldots \qquad 3.27$$

$$\therefore \quad \theta_1 = \arcsin \left(\frac{D - \alpha}{\beta}\right) \qquad \ldots\ldots\ldots\ldots \qquad 3.28$$

$$\theta_2 = \pi - \theta_1$$

The output is

$$Y = A \quad \theta_1 \leq \omega t < \theta_2$$

$$0 \quad \theta_2 \leq \omega t < \theta_1 + 2\pi \qquad \ldots\ldots\ldots \qquad 3.29$$

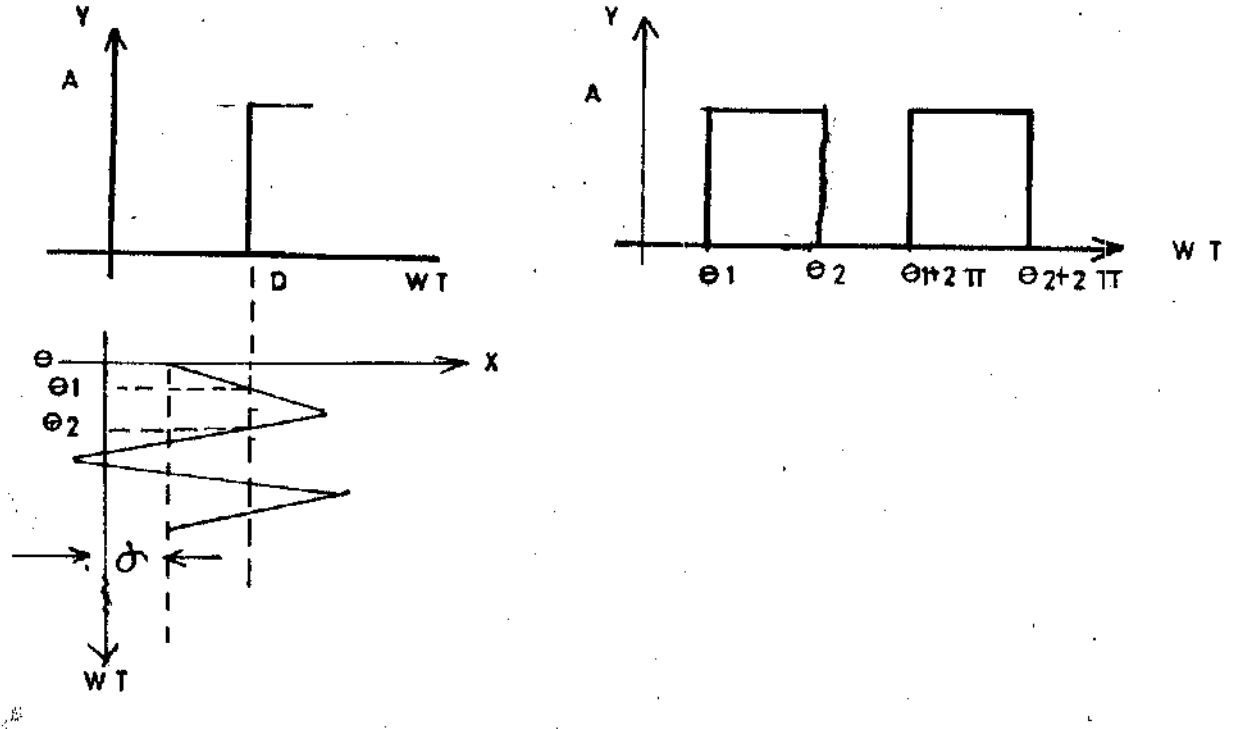


FIG 3.8 INPUT — OUTPUT WAVEFORMS OF TTL 7404 GATE

Applying Fourier series,

the d.c. term is

$$P_o = \frac{1}{2\pi}\int_{\theta_1}^{\theta_1+2} \omega(t)\, d\omega t \qquad \ldots\ldots\ldots \qquad 3.30$$

$$= \frac{A}{2\pi}(\pi - 2\theta_1) \qquad \ldots\ldots\ldots\ldots \qquad 3.31$$

According to the assumptions of the D.F., the higher terms in the Fourier series can be neglected leaving only the average value and the fundamental.

For n = 1, the cosine term is zero

For nn = 1, the sine term is $P_1 = \frac{1}{\pi}\int_{\theta_1}^{\theta_1+2} \omega(t) \sin \omega t\, d\omega t$

$$= \frac{2A}{\pi} \cos\theta_1 \qquad \ldots\ldots\ldots\ldots\ldots \qquad 3.32$$

$$\therefore P_o = N_o(\alpha_1, \beta) = \frac{A}{2\pi}\left(\pi - 2\arcsin\left(\frac{D-\beta}{\beta}\right)\right) \qquad \ldots\ldots\ldots \qquad 3.33$$

and

$$P_1 = N_1(\alpha_1, \beta) = \frac{2A}{\pi\beta}\sqrt{1 - \left(\frac{D-\alpha}{\beta}\right)^2} \qquad \ldots\ldots \qquad 3.34$$

These values are now ready to be used in conjuction with the transfer function of the frequency selecting network.

The Transfer Function of the linear part of the frequency determining network is obtained by simple circuit analysis and is

$$G_1(s) = \frac{1}{(1 + R\,S\,C_1)(1 + S^2\,L\,C_2) + R\,C_2\,S}$$

$$= \frac{1}{S^3\,R\,L\,C_1\,C_2 + S^2\,L\,C_2 + S\,R\,(C_1 + C_2) + 1} \qquad \ldots\ldots\ldots 3.35$$

Equation 3.35 did not take into consideration the self capacitance of the coil. This self capacitance is very important as far as reliable operation is concerned. The modified transfer function, taking this into consideration becomes

$$\frac{1 + S^3\,L\,C_3}{S^3(R\,C_2\,(C_3 + C_1) + C_3\,(R\,C_1 + 1)) + S(R(C_1 + C_2)) + S^2\,L\,C_2) + 1} \qquad \ldots\ldots\ldots 3.36$$

The analysis is completed by solving these two equations to obtain the oscillation conditions:

$$G_0 N_0 (\alpha, \beta) = -1 \qquad \ldots\ldots\ldots\ldots \qquad 3.37$$

where $$G_0 = \lim_{\omega \to 0} G_2(j\omega) \qquad \ldots\ldots\ldots\ldots \qquad 3.38$$

The evaluation of equation 3.37 is done for various values of to get corresponding values of B. These values of and B are then used to solve equation 3.38 in graphical form.

The frequency of oscillation is determined by the point where the plot of $G_2(j\omega)$ intersects the $-\frac{1}{N_1(\alpha, \beta)}$ axis and the amplitude

of oscillation is specified by the value of $\frac{1}{N_1(Q,\beta)}$ at that point.

Design Criteria for the Oscillator

The coil is prepared as earlier on discribed. The frequency design starting point can be obtained by ignoring C_3 and equating the imaginary part of equation 3.35 to zero.

$$\omega^2 = \frac{C_1 + C_2}{L\ C_1\ C_2}$$

or if

$$f = \frac{1}{2\pi\sqrt{\frac{L\ C_1\ C_2}{C_1 + C_2}}} \qquad \ldots\ldots\ldots \quad 3.37$$

The choice of capacitor follows the same routine as discussed in section 3.4.

3.6 Performance Tests of the Detector Loop Oscillator

Two types of Colpitts Oscillators have been designed and tested. This section discusses the relative merits of these two methods and in addition reports the performance tests as circuit components were varied.

Whereas the integrated circuit (I.C.) type may not be able to withstand more than 70°C ambient temperature (commercial type), the transistor type can withstand higher temperature than this.

Table 3.3 shows that the frequency change when a car passes is almost the same for the two types oscillating at about the same frequency.

The table also shows that the higher the frequency of operation the greater the change.

Treatment	Frequency (KHz)	$\frac{\Delta f}{f}$ %
TTL 7404 Type	57.7	1
TTL 7404 Type	104.3	1.3
Transistor Type	47.30	1

Table 3.3: Change in Frequency with Vehicle Presence.

For regular TTL power dissipation per gate is about 10 millivolts but the dissipation in transistor circuits can be controlled by the choice of biasing current. The TTL type can only withstand a supply variation of about ± 5% around 5.0 volts while the transistor type can withstand more than this depending on the bias arrangement. For class A bias it can withstand up to ± 20 % variation. [23,24]

A TTL 7404 Colpitts Oscillator was selected for further investigation in order to provide more information about its behaviour and provide a better understanding of its performance.

However, to obtain a cleaner waveform, a Schmitt trigger inverter, TTL 7414 was used. Due to its hysteresis [25], it has better noise immunity

though the frequency of operation is slightly lower, table 3.4, because of phase shift introduced by the hysteresis.

The analysis is similar to that given for 7404 earlier but the phase shift due to the hysteresis must be taken into consideration.

	C_1 nF	C_2 nF	Frequency (KHz)	
			7404	7414
1.	47.0	10.0	152.27	148.24
2.	47.0	22.0	125.35	122.00
3.	47.0	36.0	109.910	105.86
4.	47.0	47.0	103.311	98.700
5.	47.0	68.0	91.777	88.821

Table 3.4: Using 7404 and 7414 as Nonlinear Element.

The band of frequency allocated to "Metal Detector"[10] is 16 - 150 KHz with a forbidden band from 90 to 110 KHz to protect the "Standard Frequency" of 100 KHz.

The prefered band is however from 85 to 90 KHz and 110 to 144 KHz. The lowerprefer band is chosen since it will give less drift with temperature.

The effect of the self capacitance on the Detector Loop Oscillator has been discussed earlier. It has also been argued that the use of higher circuit capacitance will give improved operation. Table 3.5 gives the variation of frequency with temperature for various Inductance to Capacitance ratio.

L microhenry	C_1 nF	C_2 nF	F KHz	Δf KHz	ΔT ^{o}C
340	5.1	2.5	88.05	4.95	7.6
100	6 8	47	89.80	0.256	5.7

Table 3.5: Frequency Drift with Temperatu e for various Inductance to Capacitance ratio.

It can be seen that the use of smaller inductance provides more stable operation.

Capacitor C_1 and C_2 are both responsible for the frequency of operation although it has been suggested that (32) C_1 is responsible for the amplitude level while C_2 determines the frequency. Table 3.6 shows that both C_1 and C_2 infact influence the frequency but C_2 has greater influence.

C_1 nF	C_2 nF	Frequency KHz
33	33	117.64
33	47	112.67
47	47	99.51
47	68	89.76
68	68	83.53

Table 3.6: Influence of C_1 and C_2 on Frequency

It is therefore desirable to make C_2 big in comparison to the expected self capacitance.

In section 3.5 it was shown that the temperature coefficient of frequency is a function of the temperature coefficient of the inductor and the capacitors. Two capacitors are at ambient temperature while the inductor and the third 'capacitor'are at ground temperature. Consequently both the ambient and ground temperature affect the frequency of operation as shown in table 3.7 and 3.8.

Day	Time	Frequency (MHz)	Temperature °C		Remarks
			Ambient	Ground	
1	8.00 pm	83.37	22.2	35.5	C_1 = 5.1 nF C_2 = 2.4 nF L = 340 microhenry
	9.00 pm	84.074	23	35.25	
	10.00 pm	84.334	28.3	34.89	
	11.00 pm	84.465	22.6	34.84	
2	8.100 am	87.33	22.2	30.72	
	4.40 pm	85.851	23.8	35.33	
	5.00 pm	83.804		35.49	
	6.00 pm	83.806	23.0	35.55	
	7.00 pm	83.762	23.0	35.3	
	8.00 pm	84.054	23.0	34.83	
	9.00 pm	84.604	23.0	34.79	
3	8.30 am	87.80		30.3	
	9.30 am	87.77		30.44	
	10.00 am	87.40		30.58	
	11.00 am	87.07		30.72	
	12.00 pm	85.61		30.86	
	1. 00 pm	84.44		31.69	

Table 3.7:

Table 3.8 shows the variation of frequency with both ambient and ground temperature for different but smaller ratio of C_3 to $(C_1 + C_2 + C_3)$ ratio.

The ground temperature is measured with the aid of a thermocouple inserted on the road by the side of the coil. The other end is kept at a reference temperature. The temperature difference between the two junctions is measured by a Galvanometer - fig. 3.9.

The calibration curve is shown in fig. 3.8.

This curve can be described by the equation

$$t = \alpha\theta + T_{ref}$$

where T_{ref} is the reference temperature for the calibration of the galvanometer.

θ = galvanometer reading at a temperature t ^{o}C.

α = slope of the curve in deg. $C/_{\theta}$.

It is not necessary to return the cold end to T_{ref} each time a reading is to be taken if it is seen that

$$t = \alpha\theta + T_{ref} + (T_N - T_{ref})$$

where T_N is the new reference temperature which may be the ambient temperature.

Summary

In this chapter, the design of the Detector Loop and its oscillator were presented and their performance under normal operating conditions

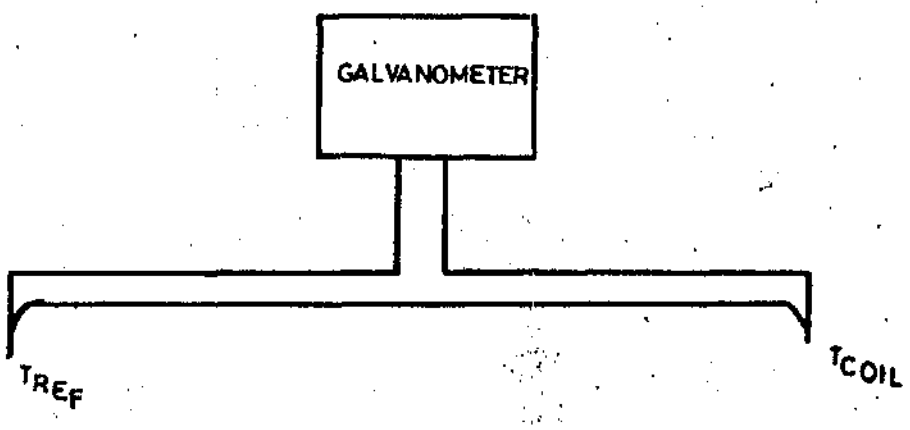


FIG 3.9 MEASURING COIL TEMPERATURE

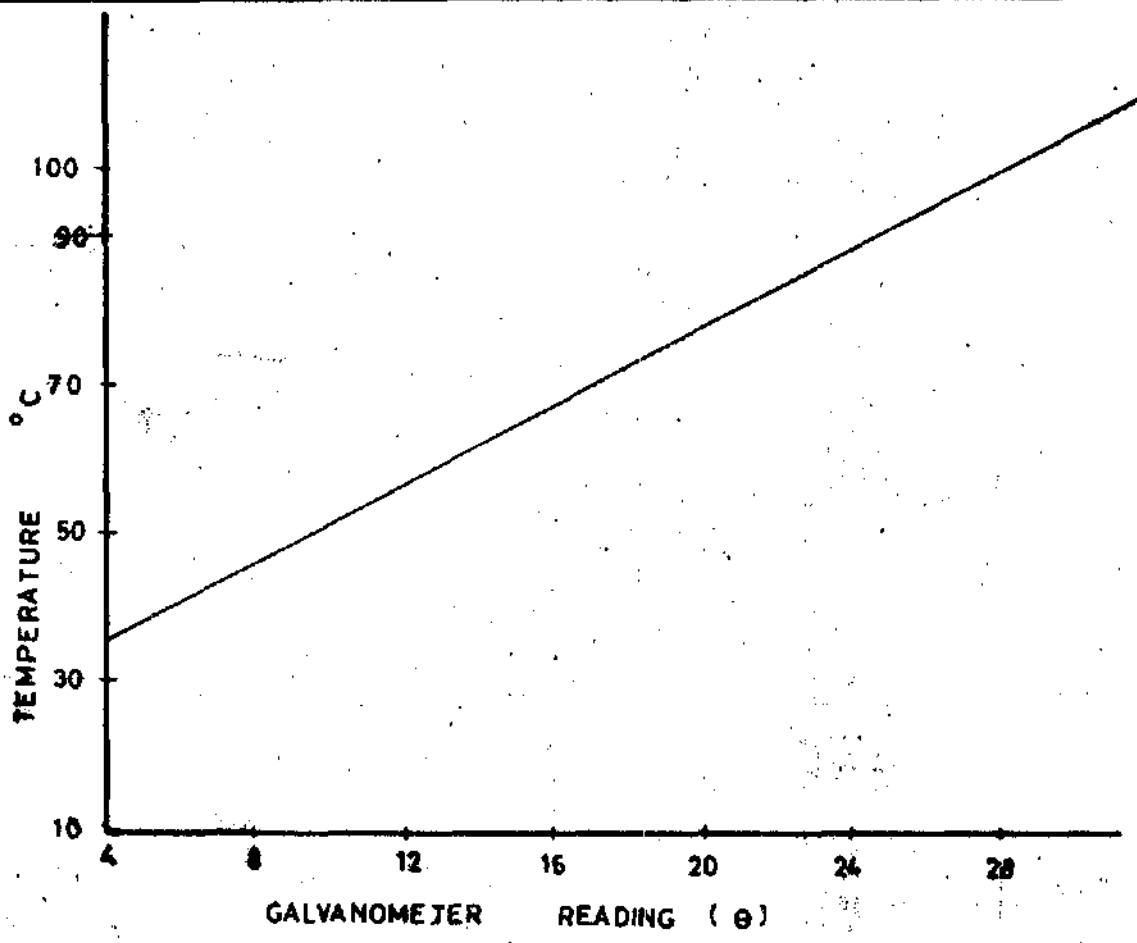


FIG 3.10 CALIBRATION CURVE

Day	Time	C_1 68 / C_2 68	Frequency in Kild Hertz 68 / 47	47 / 47	33 / 47	33 / 33	Temperature °C Ambient	Ground	Remark
1	10.00 am		89.760	99.570	112.727	117.724	21.9		L = 100 μH
	11.37 am		89.651	99.513	112.657	117.637	22.0		
	12.70 pm		89.330	99.335	112.357	117.260	22.5		
	7.39 pm		89.640	99.388	112.627	117.500	24.5	30.4	
2	9.40 am	83.581	89.766	99.705	112.800	117.870	22.0	24.4	
	11.07 am	83.551	89.741	99.581	112.860	117.790	22.2	24.4	
	12.06 pm	83.393	89.370	99.251	112.460	117.379	23	24.5	
3	10.21 am	83.521	89.721	99.660	112.910	117.866	21.5	24.5	
	12.47 pm	83.184	89.180	99.182	112.310	117.190	24.2	26	
	2.50 pm	83.144	89.157	99.035	112.210	117.110	28.3	26	
4	7.48 am	83.597	89.800	99.795	113.125	118.116	18.8	25	
	9.38 am	83.562	89.750	99.647	112.957	117.944	21.0	25	
	12.00 pm	83.420	89.574	99.460	112.671	117.620	23.3	25	
	3.00 pm	83.220	89.272	99.116	112.304	117.140	25.0	29	
	6.22 pm	83.427	89.544	99.340	112.560	117.542	24.5	31	
5	7.30 am	83.584	89.808	99.797	113.190	118.187	16.5	24.5	
	8.30 am	83.572	89.792	99.751	113.082	118.091	18.9	25.0	
	12.00 pm	83.415	89.565	99.440	112.691	117.635	24.0	25.5	
	3.00 pm	83.090	89.000	99.907	112.061	116.935	29.5	25.5	

Table 3.8: Variation of Frequency with Temperature and L/C ratio

were discussed. Various physical and electronic constraints were translated to design inequalities and suitable design procedures were outlined.

In the next chapter, the design of a PLL discriminator for the detection of the Loop Oscillator frequency variation with vehicle presence is presented.

CHAPTER FOUR

THE PHASE-LOCKED-LOOP (PLL)

4.0 Synopsis

The design of a phase-locked-loop discriminator is the subject of this chapter. It introduces the PLL as a building block and goes on to use it to build a frequency discriminator suitable for the performance characteristic of the Detector Oscillator of the previous chapter. A suitable output circuitry for interfacing with other circuits, and in particular connection for a vehicle counter with L.E.D. display is presented.

4.1 Introduction to the PLL

The analysis and results of the previous chapter indicate that although oscillator frequency can be changed by the passage of a vehicle over the loop an undesirable component of frequency change is also presented.

The Phase-Locked-Loop is used in this chapter first as a frequency discriminator to detect the sudden changes in the frequency of the loop oscillator due to the presence of a vehicle over the coil and secondly to track Detector Oscillator slow drift due to environmental conditions. This ability of the phase-locked-loop to track the drift in frequency is what determines the superiority of this method to the other methods discussed in chapter two. The need to successfully track the drift and also the rate of change of the frequency prescribe the important parameters of the PLL such as the capture range and lock range.

The fundamental concepts of the phase-locked-loop has been know by the early 1930. Its use however, was limited by cost and complexity of discrete

form implimentation.

The advent of integrated PLL implemented as either a simple chip or multi-chip[36] configurations has aimplified the design and widened the application areas of the PLL. Applications now include use in T.V., tracking of satellite signals, frequency synthesis and discrimination, and control.

4.2 Analysis of PLL Response

Basically, a phase locked loop, PLL is an electronic feedback system which consists of a phase detector, a low pass filter, an amplifier and voltage controlled oscillator, VCO, as shown in fig. 4.1.

A simplified qualitative picture of the operation of the PLL can be described as follows: The phase discriminator selects the phase change between the input signal and the voltage controlled oscillator which is selected by the low pass filter and amplified by the amplifier. The amplified error signal is used to control the frequency of the voltage controlled oscillator. In control terminology, the V.C.O. is the controlled system whilst the phase detector/amplifier/low pass filter perform the error detection and conditioning.

It is quite evident that if the system were linear and the error is such that it reduces the difference in frequency and hence phase, the V.C.O. would tend to track the changes in incoming signal frequency. The PLL is said to be in lock when the frequency of the incoming signal, the PLL are equal and with $\pi/2$ radians phase shift. In practice, a small deviation in phase is present between the two system because the system loop gain is finite.

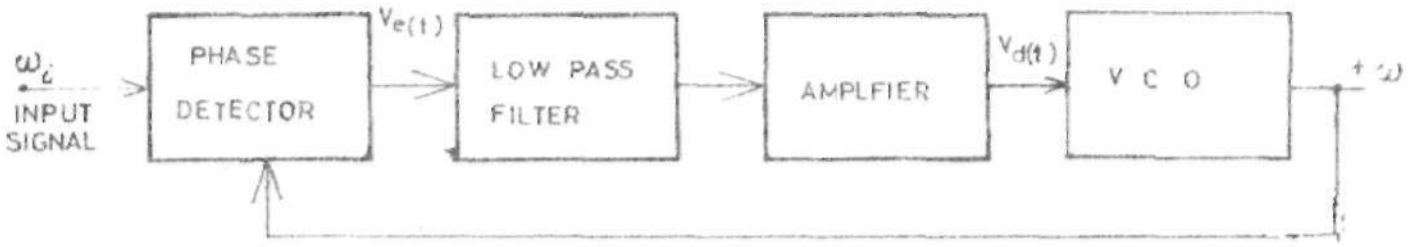


FIG 4.1 PHASE LOCK LOOP

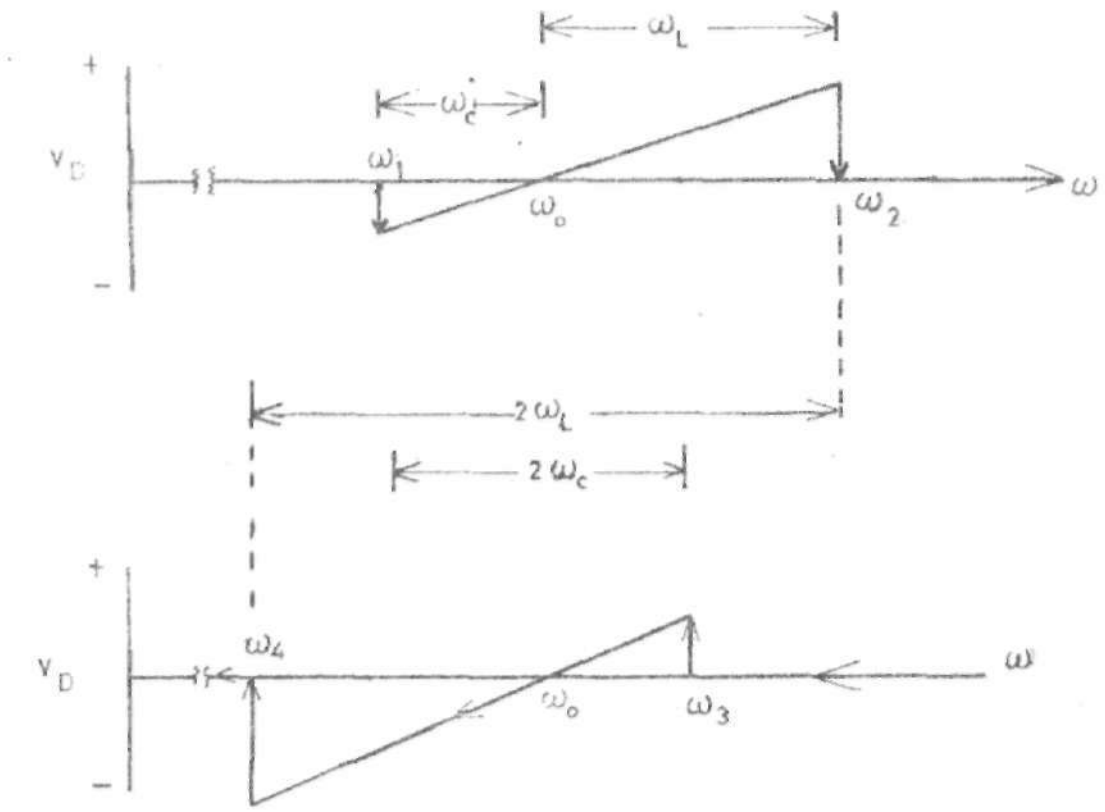


FIG 4.2 FREQUENCY TO VOLTAGE TRANSFER PLOT OF PLL

The phase discriminator in practice is a multiplier circuit which generates the sum and difference of the input frequencies.

$$e_1 = \frac{A}{2}\left[\sin(\omega_i - \omega_o) + \sin(\omega_i + \omega_o)\right] \quad \ldots\ldots \quad 4.1$$

where

ω_i = input frequency in radians

and ω_o = V.C.O. centre frequency in radians.

The low pass filter, LPF eliminates the sum frequencies such that for small difference in frequencies the output of the LPF can be represented as follows:-

$$e_2 = K(\omega_i - \omega_o) \quad \ldots\ldots\ldots\ldots\ldots \quad 4.2$$

from (4.2) it is possible to see that phase the discriminator and the LPF are equivalent to a simple frequency comparator. Depending on the characteristics of the LPF, amplifier and V.C.O. response, the response for small signals can be studied using standard servomechanism theory. This can be used to establish the tracking characteristics of the PLL system.

The ideal frequency-voltage characteristic of the PLL is shown in fig. 4.2 to illustrate its behaviour with varying input signal.

The difference frequency $(\omega_i - \omega_o)$ is filtered with the sum frequency $(\omega_i + \omega_o)$ when it is greater than the upper band edge of the LPF. However, as ω_i approaches ω_o, the difference $(\omega_i - \omega_o)$ falls

into the passband of the LPF and produces an error voltage which shifts the V.C.O. and hence reduces the value of $(\omega_i - \omega_o)$ until lock is attained. When the input frequency ω_i is increased from zero the operating frequency remains the same until ω_1 is reached when the loop 'captures' the input signal and lock is established. The operating frequency then follows the input signal frequency until the upper end of the lock range is reached at ω_2 where lock is lost and V.C.O. returns to ω_o.

On the other hand when the frequency is decreasing towards ω_o, the loop 'captures' the input signal at ω_3 and V.C.O. tracks the input signal until the lower end of the lock range is reached where lock again is lost and V.C.O. returns to ω_o.

The capture range which is the range of input frequencies over which the loop can acquire lock is given by $(\omega_3 - \omega_1)$ is determined by the low pass filter.

The low pass filter L.P.F. plays an important role in the operation of the P.L.L. and its choice depends on particular application. Whereas a large time constant provides an increased memory effect for the loop. The tracking rate is lowered. A large time constant, however provides increased immunity to noise and out-of-band signals. Under this condition, the capture range is reduced and the capture transient, the settling time, becomes longer.

A single-pole RC filter is the simplest filter for a second order loop, fig. 4.3.

When used with a PLL, for example, Signetic PLL NE 565, the resistor R is internal and the capacitor is connected between pin 7 and ground, fig 4.4. If transient has died down, the system may be considered linear and the open loop transfer function of the PLL is

$$T(S) = \frac{K_v \, F(s)}{S} \qquad \ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad 4.3$$

where

K_v = $K_o k_d$ A

K_o = V.C.O. Conversion gain rad/sec./Volt.

K_d = Conversion gain of phase detector

A = Amplifier gain

F(s) = Transfer function of the filter.

The closed loop transfer function is

$$H(s) = \frac{T(s)}{1 + T(s)} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \quad 4.4$$

for the single pole RC-filter of fig. (4.3a).

$$F(s) = \frac{1}{1 + \tau S} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \quad 4.5$$

where τ = RC

Then the closed loop response is

$$H(s) = \frac{K_v}{S\,(S + \tau S)} \qquad \ldots\ldots\ldots\ldots\ldots \quad 4.6$$

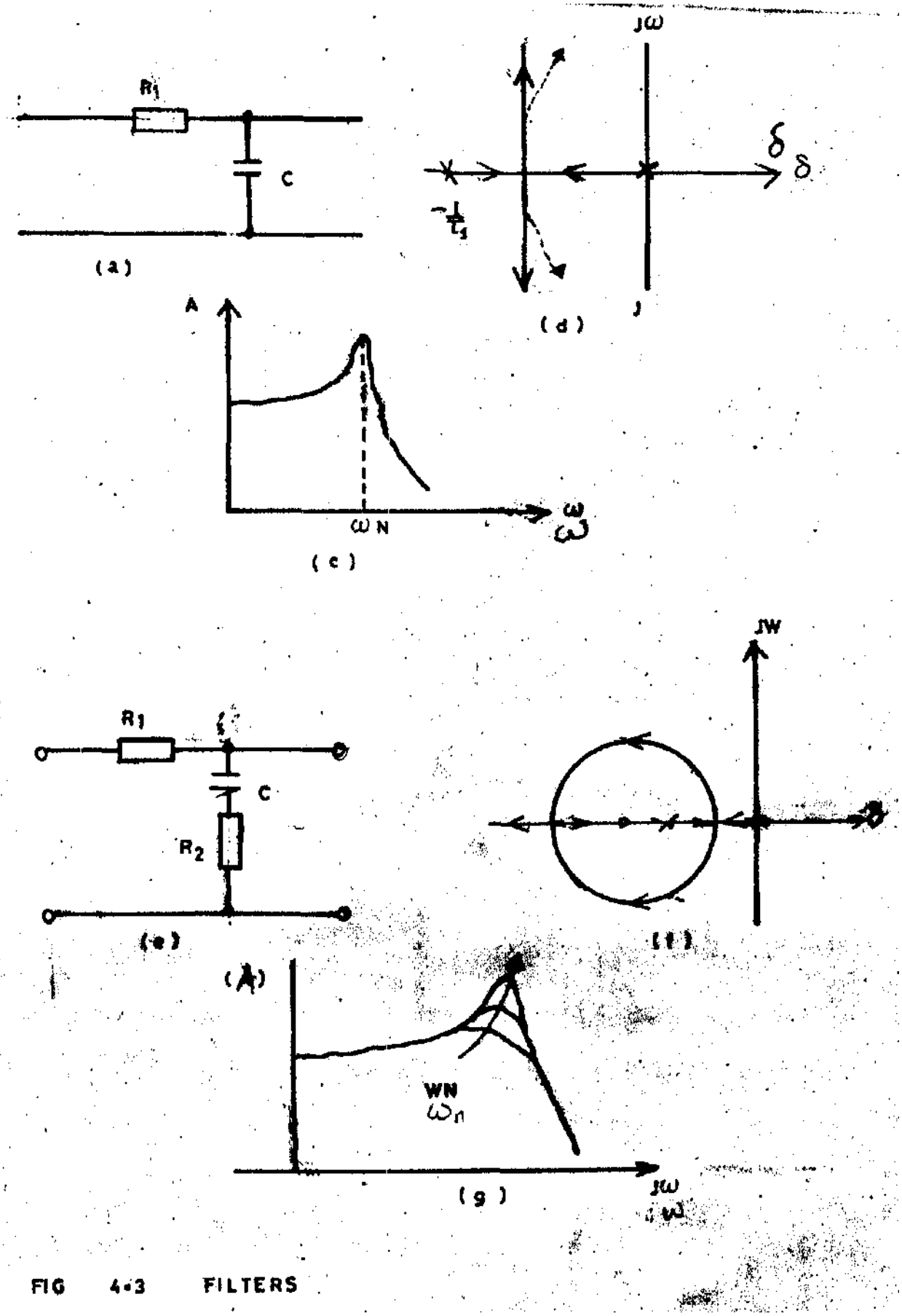


FIG 4·3 FILTERS

which gives the root locus and frequency response shown in fig. 4.3 (b and c) respectively.

The PLL system is stable but may become unstable due to a parasitic time constant or inappropriate increase in K_v. In this case, the root locus bends towards the imaginary axis (see dotted lines of fig. 4.3 (b)). This tendency towards instability is usually eliminated by the use of a lag-lead filter of the type shown in fig. 4.3 (e - g) with the associated root locus and frequency response.

Here

$$F(s) = \frac{1 + \tau_2 s}{1 + (\tau_1 + \tau_2)s} \quad \text{.....} \quad 4.7$$

where $\tau_1 = R_1 C$

$\tau_2 = R_2 C$

The choice of τ_2 enhances stability and the frequency response now depends on the damping which can be controlled by τ_1 and τ_2. These results are used latter with the signetic NE 565 PLL.

The lock range which is the range of input frequencies over which the loop will remain in lock (once lock has been acquired), on the other hand is not a function of the low pass filter. This is a d.c. parameter which is a function of the range of error voltage that can be generated to produce a corresponding V.C.O. frequency deviation.

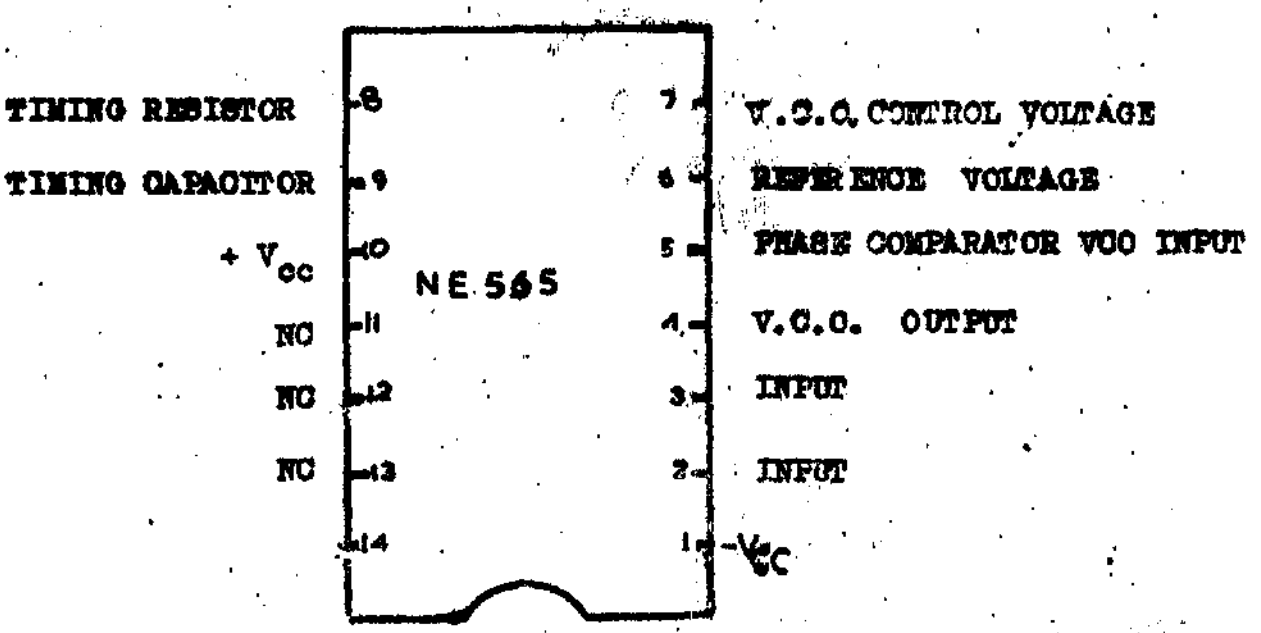
The V.C.O. has a maximum swing beyond which lock will be lost. The voltage developed at the output of the phase detector is a function of the product of the phase and the amplitude of the in-band component to which the loop is locked. Therefore, a decrease in signal amplitude must be accompanied by an increase in phase difference between V.C.O. and input signal if the output voltage (and consequently frequency deviation) is to be constant. Generally, a strong input signal will allow the maximum use of the V.C.O. swing capability.

The speed of response is a function of the low pass filter. The low pass filter limits the maximum rate at which tracking can occur because the LPF capacitor voltage cannot change spontaneously. For fast operation small time constant with high frequency cut-off is used but the sum frequencies will be presented at the output. The trade off selected will depend on particular applications.

The center frequency, $_o$, which is the free running frequency of the VCO when not in lock is given by expressions characteristic of each V.C.O. For signetic NE 565 PLL, the center frequency is given approximately by[35]

$$f_o \simeq \frac{1.2}{4 R_1 C_1} \text{ Hz} \qquad \ldots\ldots\ldots\ldots\ldots \quad 4.8$$

This center frequency can be tuned by the selection of capacitor C_1, fig 4.4. The phase detector is a multiplier which generates a voltage proportional to the difference in phase of the two input signals. This may be



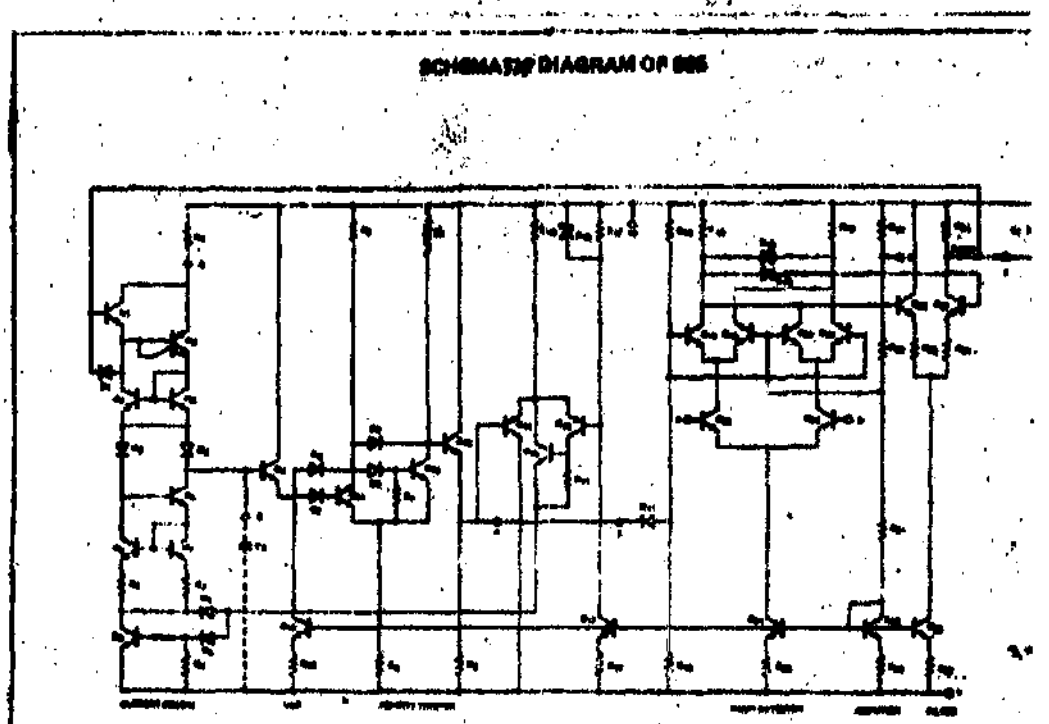


FIG 4.4 NE 565 P.L.L

sinusoidal or square signal phase detector employing diodes,[34] transistors[35] and integrated circuit gates.[11,35,36] The principle is similar and it is usually required that the phase difference/output voltage be linear.

4.3 The Signetic NE 565 PLL

This is a single chip PLL whose pin configuration and schematic diagrms are shown in fig. 4.4.

The Phase Detector

This is a double-balanced mixer-type, fig. 4.5 where the differential amplifier input stage has a differential gain of

$$A_d = \frac{R_c}{0.013\ I_o}$$

The input is multiplied by a unity square wave

$$\frac{4}{\pi(2n+1)} \sin\ ((2n+1)\omega_o t)$$

where $n = 0, 1, 2, 3, \ldots\ldots$

The effect of this multiplier action can be seen by considering input signal

$$V_i + V_k = V_i \sin(\omega_i + \theta_1) + V_k \sin(\omega_k t + \theta_k). \quad \ldots\ldots 4.9$$

The phase detector output is

$$V_d = \frac{4\,V_i}{\pi(2n+1)} \sin((2n+1)\omega_o t).\ \sin(\omega_i k + \theta_1)$$

$$+ \frac{4\,V_k}{\pi(2n+1)} \sin((2n+1)\omega_o t).\ \sin(\omega_k t + \theta_k) \ldots\ldots 4.10$$

which can be reduced to the following form by applying trigonometrical relations.

$$V_d = \frac{2A_d}{\pi} \sum_{n=0}^{\infty} \left[\frac{V_i}{(2n+1)} \cos((2n+1)\omega_o t - \omega_i t - \theta_i) - \frac{V_i}{(2n+1)} \cos((2n+1)\omega_o t + \omega_i t + \theta_i) + \frac{V_k}{(2n+1)} \cos((2n+1)\omega_o t - \omega_k t - \theta_k) - \frac{V_k}{(2n+1)} \cos((2n+1)\omega_o t + \omega_k t + \theta_k) \right] \quad \ldots\ldots 4.11$$

If $V_k = 0$ and V_i is close to ω_o, the first term (n = 0) has a low frequency component which feeds round the loop to bring about lock. At lock $\omega_o = \omega_i$ and the first term becomes

$$V_d = \frac{2A_d V_i}{\pi} \cos\theta_i \quad \ldots\ldots\ldots \quad 4.12$$

With $V_k = 0$, n = 0, the second term shows that the sum frequency is always present at the phase detection output. If $V_k \neq 0$, the third term indicates that the loop can be driven out of lock by another frequency difference particularly when at its extreme ends. The forth term has little effect.

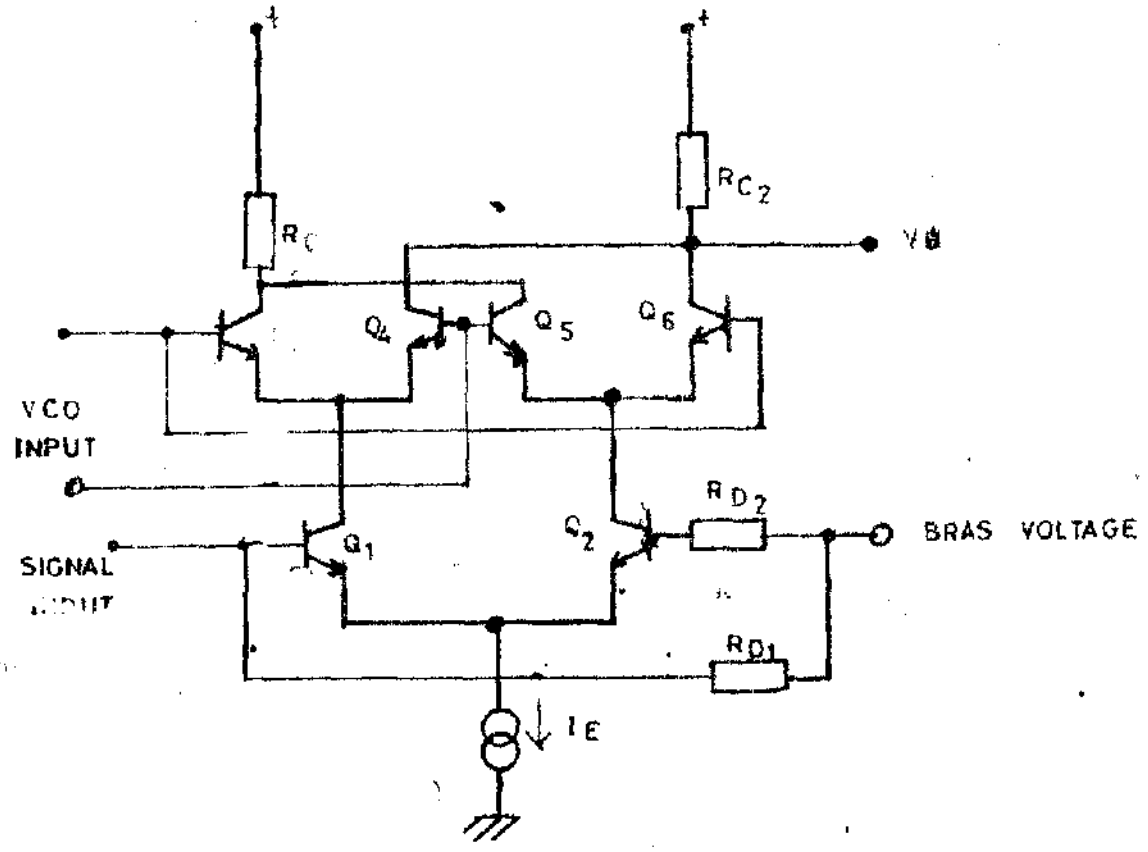
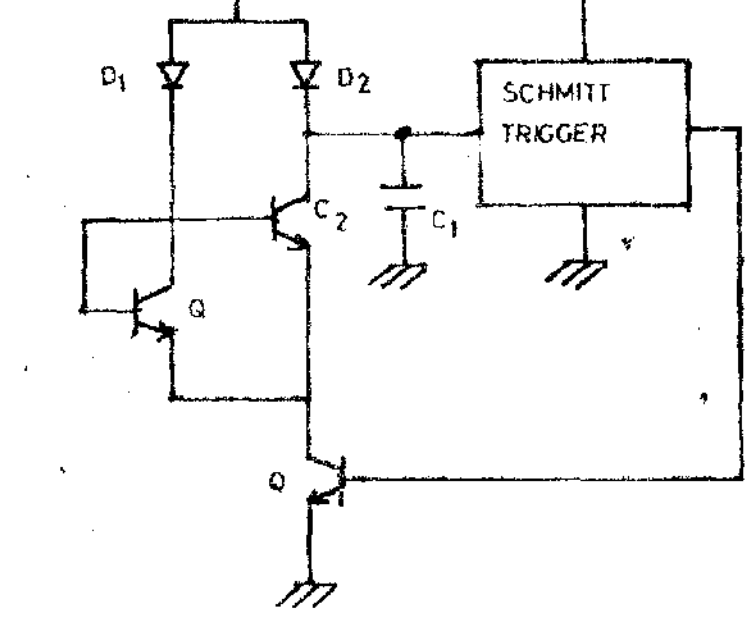


FIG4.5 INTEGRATED PHASE DETECTOR



The Voltage Controlled Oscillator (V.C.O.)

This consists of a precise current source and a non-saturating Schmitt trigger, simplified diagramatically in fig. 4.6. The operation may be understood as follows. If ihitially transistor Q_3 is OFF, the charging current I_1 due to the control voltage V_c charges the capacitor C_1 until the threshold level of the Schmitt trigger is reached when it changes state and transistor Q_3 is now turned ON.

In this state, transistor Q_1 and Q_2 are turned ON and the current I_1 flows through diodes D_1 and D_2. The portion flowing through D_2 discharges capacitor C_1 until the Schmitt trigger changes state and the cycle repeats itself.

The V.C.O. output is a function of the control voltage and has an integrating property which gives the closed loop transfer function a pole always at the origin.

In the operation as a PLL, the V.C.O. output, pin 4, must be connected externally to the phase detector input, pin 5.

4.4 The Design Considerations of the PLL Frequency Discriminator

The analysis of the previous section provides the explanation of the potential effectiveness of the PLL as a frequency discrininator for the Loop Detector. It is possible to track the slow changes (drifts) in the Detector Oscillator frequency by selecting the lock range. The vehicle-initiated changes which are much faster can be tracked by establishing

the capture range. These can be translated into the following criteria.

Lock Range Criterion

The theoretical development[35] indicates that the lock range is related to the centre frequency and the supply voltage by the expression

$$2 f_L \simeq \frac{16 f_o}{V_{cc}} \quad \ldots\ldots\ldots \quad 4.11$$

where V_{cc} = total supply voltage

f_o = centre frequency

For a supply voltage of $\pm$ 5 volts which is compactible with supply to other TTL IC chips, the lock range is, for f_o = 80 KHz.

$$26 \text{ KHz} \leq 2 f_L > 144 \text{ KHz} \quad \ldots\ldots. \quad 4.12$$

This covers, theoretically the band allocated to metal detectors including the forbidden band 90 - 110 KHz. The lock range must be reduced and the power supply values should remain the same. A method for reducing the lock range is given later in the chapter.

Capture Range Criterion:

The performance tests on the Detector Oscillator in chapter three indicates that the frequency drifts could be as high as 8 kilohertz for a dry day and about twice this value when it rained for a worse case selection of inductance to capacitance ratio. This implies that the

Capture range $\geq$16 kilohertz.

The maximum speed expected along the route may be taken as 100 km/hr. This speed can only be attained at off-peak periods in urban centres. At this speed, the time taken to cover an average car length of 5 meters is 180 mS.

This implies that the

$$\text{Capture time} \leq 180 \text{ mS}$$

This pull-in time depends on the initial frequency and phase difference between the two signals, the loop gain as well as the low pass filter bandwidth. A wider bandwidth implies a bigger capture time.

For a second order loop with a transfer function of

$$T(s) = \frac{\omega_n^2}{s^2 + 2\delta\omega_n s + \omega_n^2} \qquad \ldots\ldots\ldots 4.13$$

where ω_n = natural frequency

δ = damping factor

the system oscillatory tendency when the damping factor δ is small.

When

$$\delta \geq 1$$

there is no over-shoot and the loop captures instantaneously. The time for maximum output (22) is

$$t_{max} = \frac{\pi}{\omega_n \sqrt{1 - \delta^2}} \qquad \ldots\ldots\ldots\ldots 4.14$$

If $\delta = 1$, for critical damping

then

$$t_{max} = \frac{\pi}{\omega_n} \quad \ldots\ldots\ldots \quad 4.15$$

For this second order loop[34]

$$\omega_n = \sqrt{\frac{K}{\tau_1}}$$

$$\therefore \quad \tau_{max} = \sqrt{\frac{\tau_1}{k}} \quad \ldots\ldots\ldots\ldots \quad 4.16$$

From where it can be seen that the capture time increases with the LPF time constant.

Ideally

$$t_{max} \leq 180 \text{ nS.} \quad \ldots\ldots\ldots \quad 4.17$$

The capture range for this second order loop is given by[35]

$$2(2\pi f_c) \quad 2\omega_c \quad 2\sqrt{\frac{\omega_L}{\tau}} \text{ radians} \quad \ldots\ldots \quad 4.18$$

where $2\omega_L$ = lock range.

But the lock range is also given by

$$2 f_L \simeq \frac{16 f_o}{V_{cc}} \text{ Hz.} \quad \ldots\ldots\ldots \quad 4.19$$

where f_o is the center frequency

and V_{cc} is the total power supply

Hence

$$f_c \simeq \frac{1}{\pi} \sqrt{\frac{32\pi f_o}{\tau V_{cc}}} \quad \ldots\ldots\ldots \quad 4.20$$

which is true for

$$f_c < 1/3\ f_L$$

Fig. 4.7 shows the capture range as a function of f_o for the same filter.

The centre frequency design equation for NE 565 is given by the manufacturer as

$$f_c = \frac{1.2}{4\ R_1\ C_1} \qquad \ldots\ldots\ldots\ldots \qquad 4.21$$

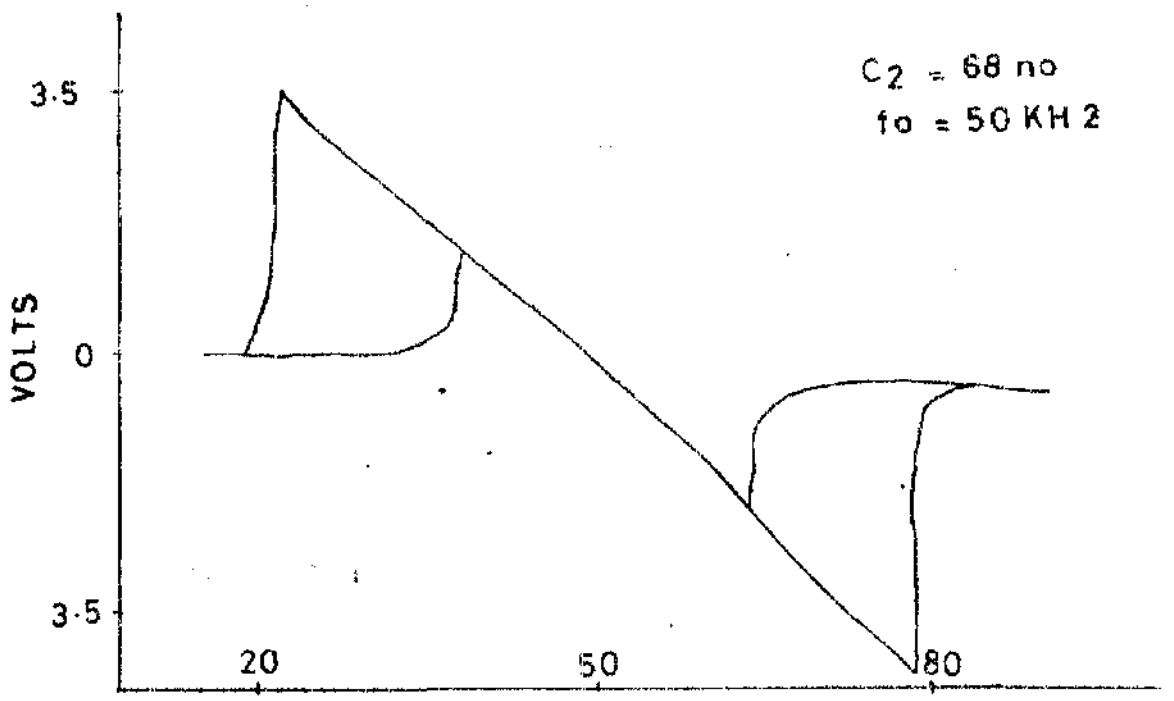
where R_1 = 3.6 kilohm and is internal

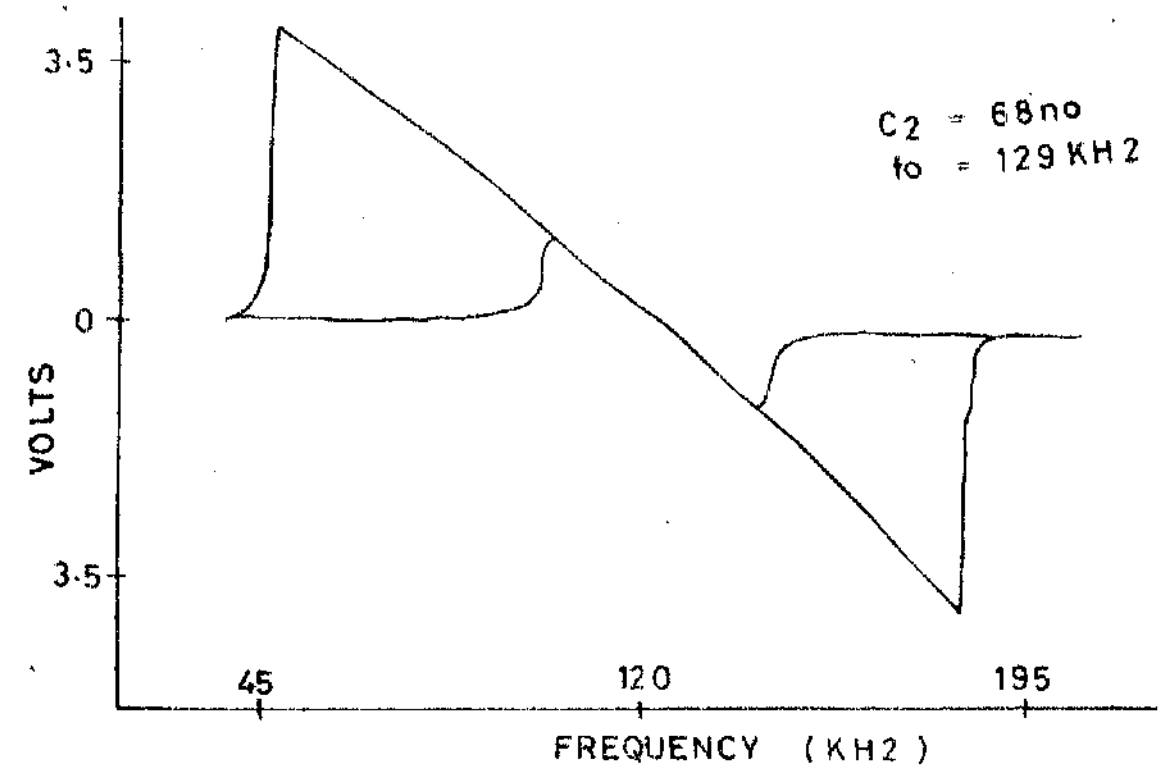
and C_1 is selected to give the required centre frequency.

The theoretical lock range and capture range expressions are given respectively in equations 4.11 and 4.20. The lock range and capture range are shown in fig.4.8a as traced by the X - Y plotter. The photographs of its oscilloscope display is in plate no. 2.

The capture range and the lock range shown in fig. 4.8a are obtained by the apparatus (35) set up of fig. 4.8 b. A voltage controlled generator is swept linearly through the frequency range of 10 to 200 KHz. The output of the generator feeds the input of the PLL. The output of the PLL is connected to the Y-axis while the sweep voltage is connected to the X-axis of an oscilloscope. The sweep frequency must not be too large, typically it should be between 1/100,000 to 1/1000 of the centre frequency.

A hard copy of this measurement is obtained by using the X - Y plotter. The frequency is hand swept using the offset control of the sweep generator. The reciprocal of the slope of line AB gives the V.C.O. coversion gain.





LOCK CAPTURE RANGE & CENTRE

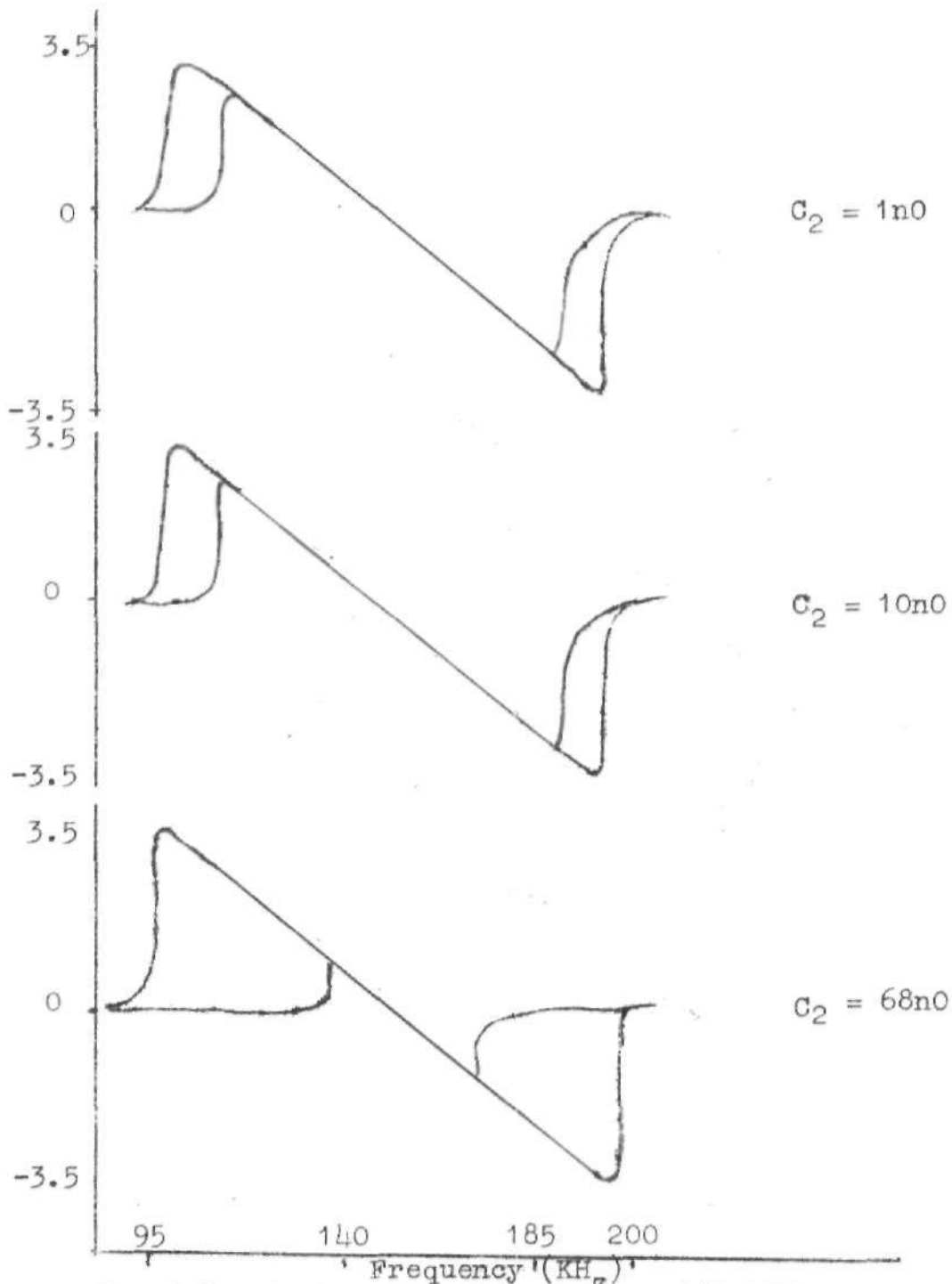


Fig 4.8a Lock and Capture Range with LPF

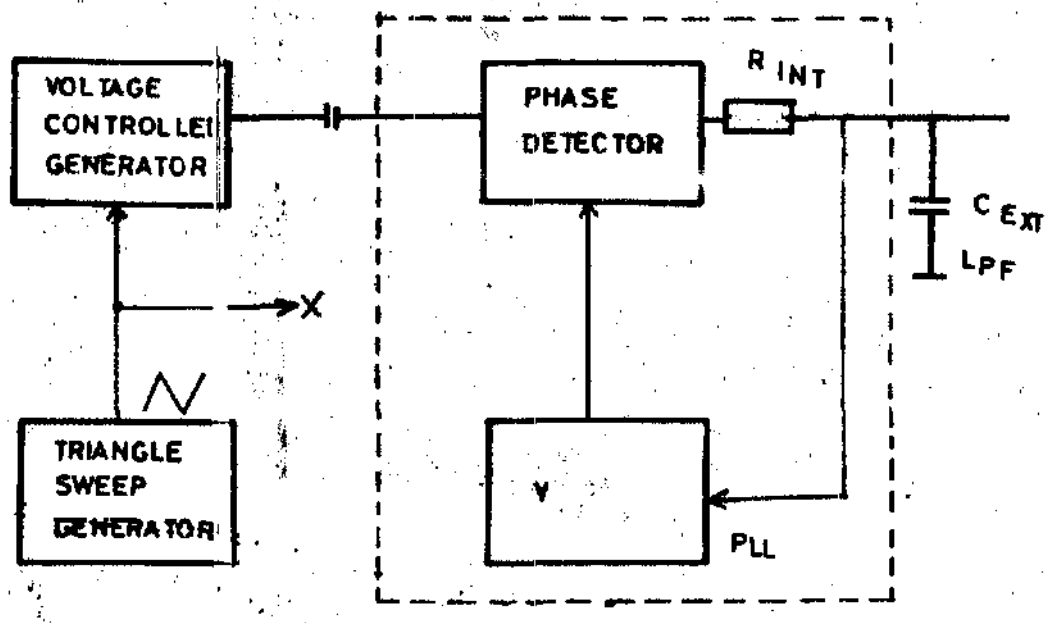


FIG 4·6b CAPTURE AND LOCKED RANGE MEASUREMENT SETUP

However, since the expected change in the Detector Oscillator frequency when the vehicle crosses is in the neighbourhood of 2%, a narrow band design is more acceptable. This is justified further by the expected value of the

$$\text{drift} \leq 16 \text{ KHz}$$

and the need to avoid the 90 to 110 KHz band.

The narrow band behaviour can be achieved by modifying the feedback path of the ME 565.[35,37]

The total loop gain is given by[35]

$$K_v = K_o k_a A$$

which can be changed by modifying the effective value of K_o, the V.C.O. conversion gain.

Modification of K_o involves injecting a constant current into pin 8, fig. 4.4 over and above that supply by R_1. This method is shown in fig. 4.9 where a constant current is injected into pin 8.

Transistor Q_2 is used as a diode for temperature stability. The loop circuitry establishes a control voltage V_c which makes the loop centre frequency to be independent of the supply voltage. This is quite an advantage in situations where the detector is supplied from a battery.

For a charging current of 1.0 mA

V_{R_A} is approximately 200 mV

and $R_A \simeq \frac{200}{2} \frac{\text{mV}}{\text{mA}} = 100 \text{ ohms}$

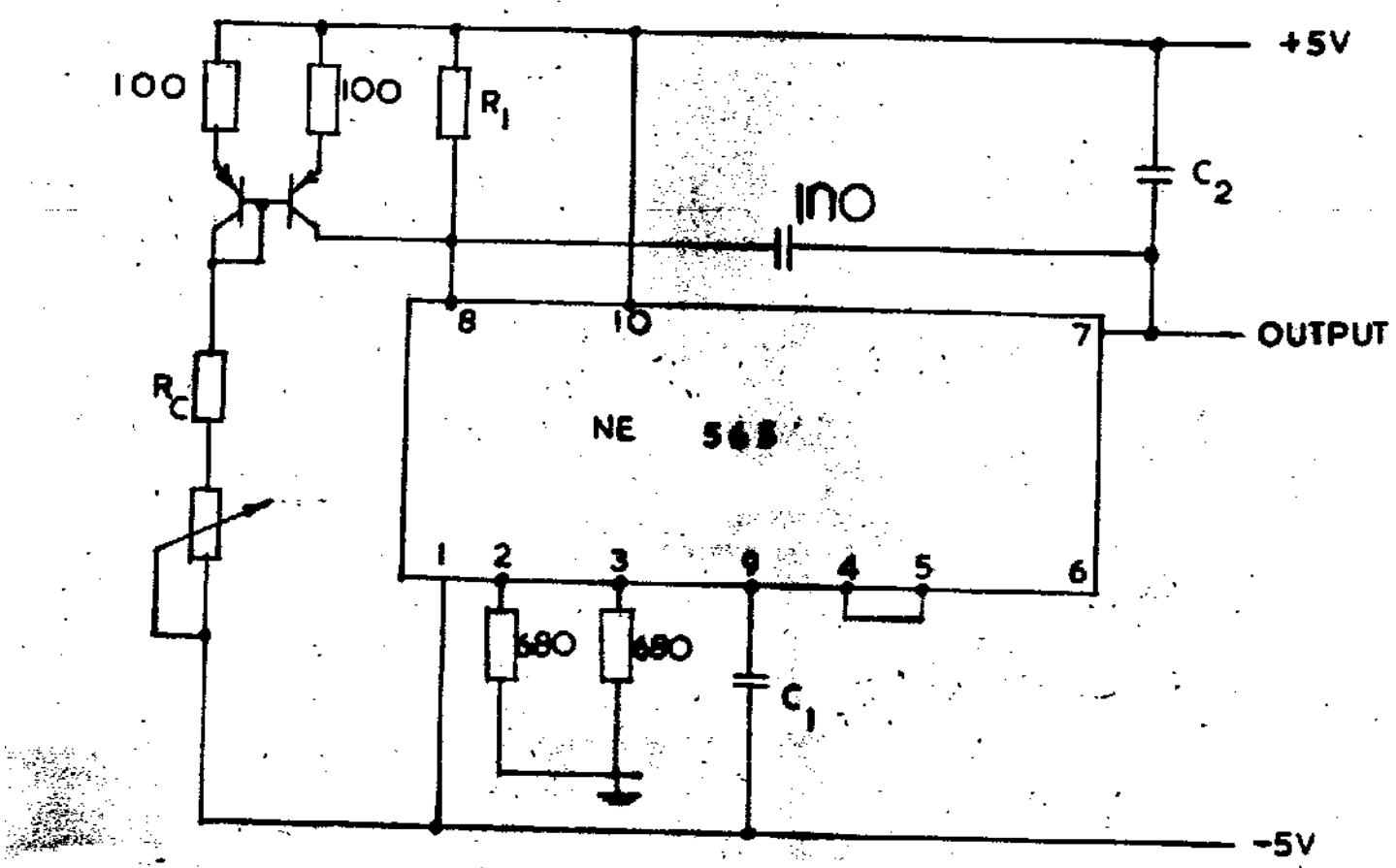


Fig 4.9 NARROW BAND OPERATION

(37)
The centre frequency design equation is

$$f_o = \frac{1.2 + 10 \frac{R_1}{R_B}}{4 R_1 C_1} \qquad \ldots\ldots\ldots \qquad 4.22$$

and the lock range is reduced by the factor

$$\frac{63}{50} \quad \frac{f_o}{(1.2 + 10\frac{R_1}{R_B})}$$

The new capture range and lock range for this narrow band operation are shown in fig. 4.10 for different values of LPF time constant. K_o has changed from 85.6 KHz/V to 6.44 KHz/V - table 4.

It can be seen generally that the LPF determines to a great extend the capture range both for normal and narrow band operations.

4.5 Output Circuitry

The output circuitry consists of an amplifier which increases the signal amplitude of the PLL discriminator output. This output can be converted into a clean pulse by a Schmitt trigger monostable integrated TTL circuit type 74121.

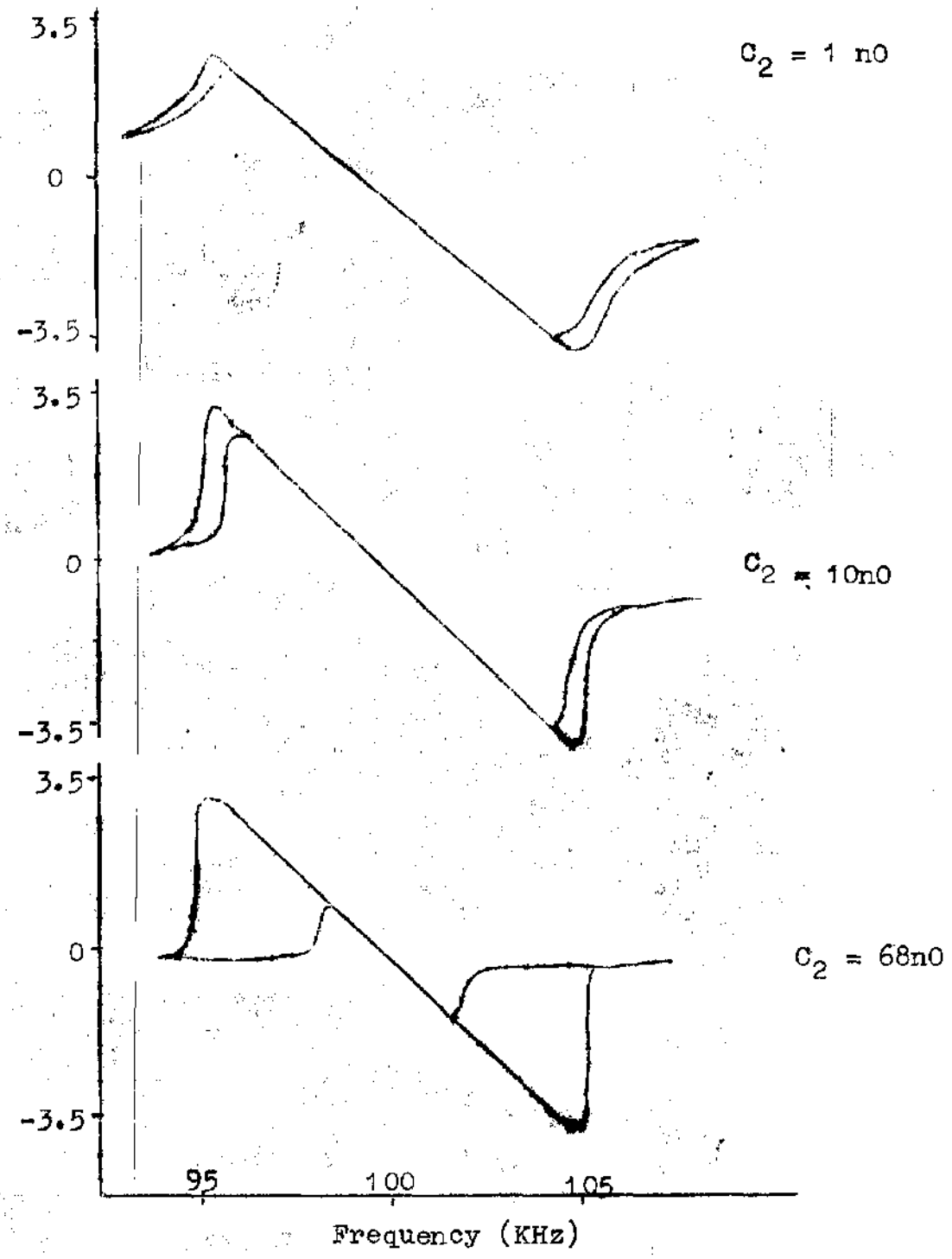


Fig. 4.10: Lock and Capture Rand with EPF (N.B.O.)

LPF nF	Mode	Frequency (KHz)				PLL Output Voltage				K_o KHz/V
		f_1	f_3	f_2	f_4	V_1	V_2	V_3	V_4	
1	Normal	42.0	65.11	182.52	196.0	4.45	4.28	2.85	2.65	85.6
	N.B.O.	95.09	95.72	105.2	105.5	4.45	4.3	2.75	2.8	6.3
2	Normal	42.53	104.7	149.48	196.0	4.45	3.83	3.3	2.65	85.3
	N.B.O.	95	96	104.56	105.5	4.45	4.35	2.6	2.75	6.44
3	Normal	43.87	119.79	113.27	195.85	4.45	3.65	3.5	2.65	84.4
	N.B.O.	95.49	96.34	102.25	105.5	4.43	3.94	3.25	2.8	6.0

Table 4.1 Comparing Normal and Narrow Band Mode

Many displays could be used. The output could drive a counter with Light Emitting Diode (LED) display or a mechanical counter (after a proper interfacing).

The output also could be interfaced with a chart recorder or a X- Y recorder (see latter) for graphical display of Vehicle density. It is also possible to interface the output with a telex which will give a hard copy of the traffic situation on paper.

The 741 operational amplifier is used to amplify the signal level. The configuration chosen is the capacitor coupled inverting type shown in fig. 4.11. This amplifier is an a.c. coupled variety aimed at eliminating the offset d.c. voltage which appears at the output of the PLL due to the drift in Detector Oscillator frequency. This capacitor together with the input impedance of the 741 stage will act as an external filter which will remove the sum of frequency if present.

The gain of the amplifier can be selected according to the threshold defined for the detector. Various thresholds can be defined for the bicycle, the motor cycle, motor car and multi-axile vehicle. In this thesis, the bicycle are not designed for. Hence, a gain is selected which will raise the output of the PLL to the voltage which can drive other TTL gates when ever a Vehicle crosses.

The V.C.O. conversion gain K_o is given for NE 565 by

$$k_o = \frac{2(\omega)_L}{K_d A \theta_d}$$

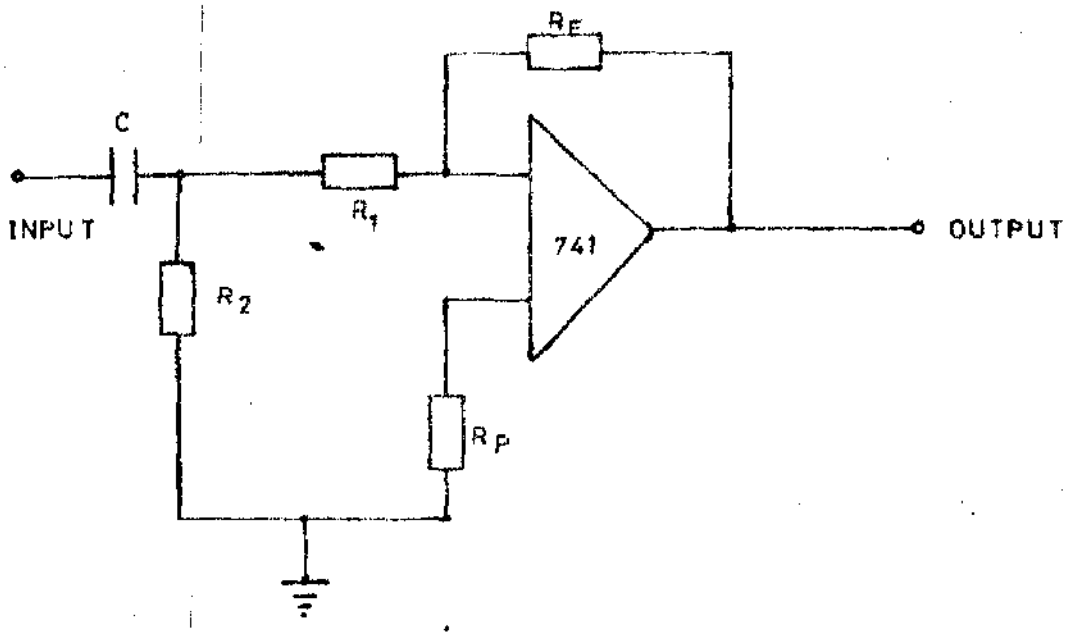


FIG 4.12 OPERATIONAL AMPLIFIER

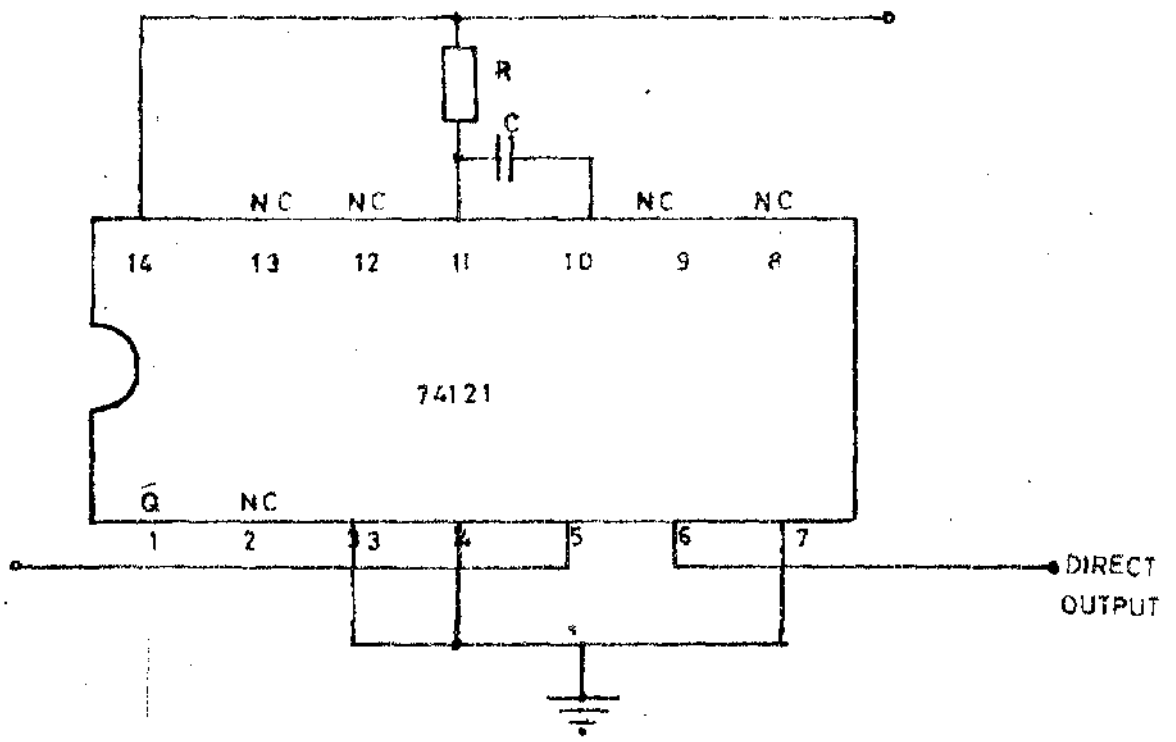


FIG 4.11 MONOSTABLE CIRCUIT

where $2\omega_L$ = lock range

$K_d = \frac{1.4}{\pi}$ volts/radians

$A = 1.4$

$\theta_d = \pi/2$

$K_o = \frac{2\omega_L}{\frac{(1.4)^2}{2\pi}}$

For a normal operation, the lock range is about 118 KHz of an operating frequency of 80 KHz.

This implies that

$$K_o \approx 1.3 \times 10^4 \text{ Hz/V}$$

for a frequency change of 600 Hz which is about 1% of f_o, when the vehicle passes, error voltage applied to the V.C.O. is about 38 mV.

A gain of about 35 has to be provided by the operational amplifier to raise this output to about 1.60 which drive other TTL gates.

The value of R_1 was selected such that the amplifier input impedance does not load the PLL stage.

The input resistance is approximately equal to R_1, provided of course that

$$R_2 \gg R_1$$

Assuming a value of 68 k for R_1 in order to permit a big choice of R_f and consequently the β - voltage transfer ratio will be small and the system's stability will be enhanced.

Note that

$$\beta = \frac{Z_{in}}{Z_{in} + Z_f} \qquad \ldots\ldots\ldots \quad 4.24$$

$\therefore$ R_f = 68 x 35 kiloohn $\simeq$ 2.2 Mega ohms.

The feedback resistor R_f provides the d.c. biasing to the non-inverting input. R_2 is made about 10 R_1 so that it will have little effect on the input impedance.

The longest transit time occurs when the car is moving at about 5 Km/hr and it is approximately 3.6 seconds. But the loop would have 'captured' the Detector Oscillator frequency before 180 mS.(milliseconds).

$\therefore$ C x $Z_{in} \leq$ 180 mS

or C $\leq \frac{180}{68}$ μ F

$\leq$ 2.6 μ F

C is selected to be 1.0 u F since the capture time is much smaller than 180 mS.

In order to minimise output offset voltage the resistor R_p was introduced such that

$R_p \simeq R_f$.

The operational amplifier output in general is distorted and can contain evidence of undamped behaviour. This signal is thus shapened by a Schmitt trigger monostable circuit giving a clean pulse for external use. The type used is the TTL 74121.[38,39]

The quarsistable time is determined by

$$R_{ext} \text{ and } C_{ext}$$

and satisfies the expression

$$T = 0.7\, R_{ext}\, C_{ext} \quad \ldots\ldots\ldots \qquad 4.25$$

For two vehicles travelling at 100 Km/hr, the distance between them should be ideally about 110 meters.

Let the distance be 50 meters

The distance is time therefore is

$$t = \frac{50}{10^5/3600} \text{ secs.}$$

$$= 1.8 \text{ seconds.}$$

The quarsistable time can then be selected according to this in equality

$$t \angle 1.8 \text{ seconds}$$

Let $t = 5$ mS

$\therefore$ $0.7\, R_{ext}\, C_{ext} = 5$ mS.

If R_{ext} = 24 kiloohm (which is less than 40 kiloohm maximum specified)[23]

then $C \triangleq 2\ \mu F$

This is connected in the level sensitive mode with snap action and hysteresis by grounding inputs A_1 and A_2 - fig. 4.12. The output of the operational amplifier is connected to input B, pin 5. The direct and indirect outputs are at pin 6 and 1 respectively.

The block diagram of the Vehicle Detector using LED readout is shown in fig. 4.13. The output of the 74121 is connected to a B.C.D. Counter which counts the number of vehicles that cross the Detector Loop. A LED display is obtained by using a BCD decoder driver to drive the way to achieve this using TTL 7448 and LED type Heulett Packard 5082 - 7740.

Summary

An introduction has been given into the PLL as a building block in this Vehicle Detector design. The various components of the PLL has been described with their role in the overall performance of the PLL stage. The various design criteria has been presented also with an output circuitry for connection to external load. A specific display, L.E.D. display is included as an example of various ways in which the output can be displayed.

In the next chapter, applications of the Vehicle Detector designed are discussed with respect to traffic survey and automatic traffic control at intersections.

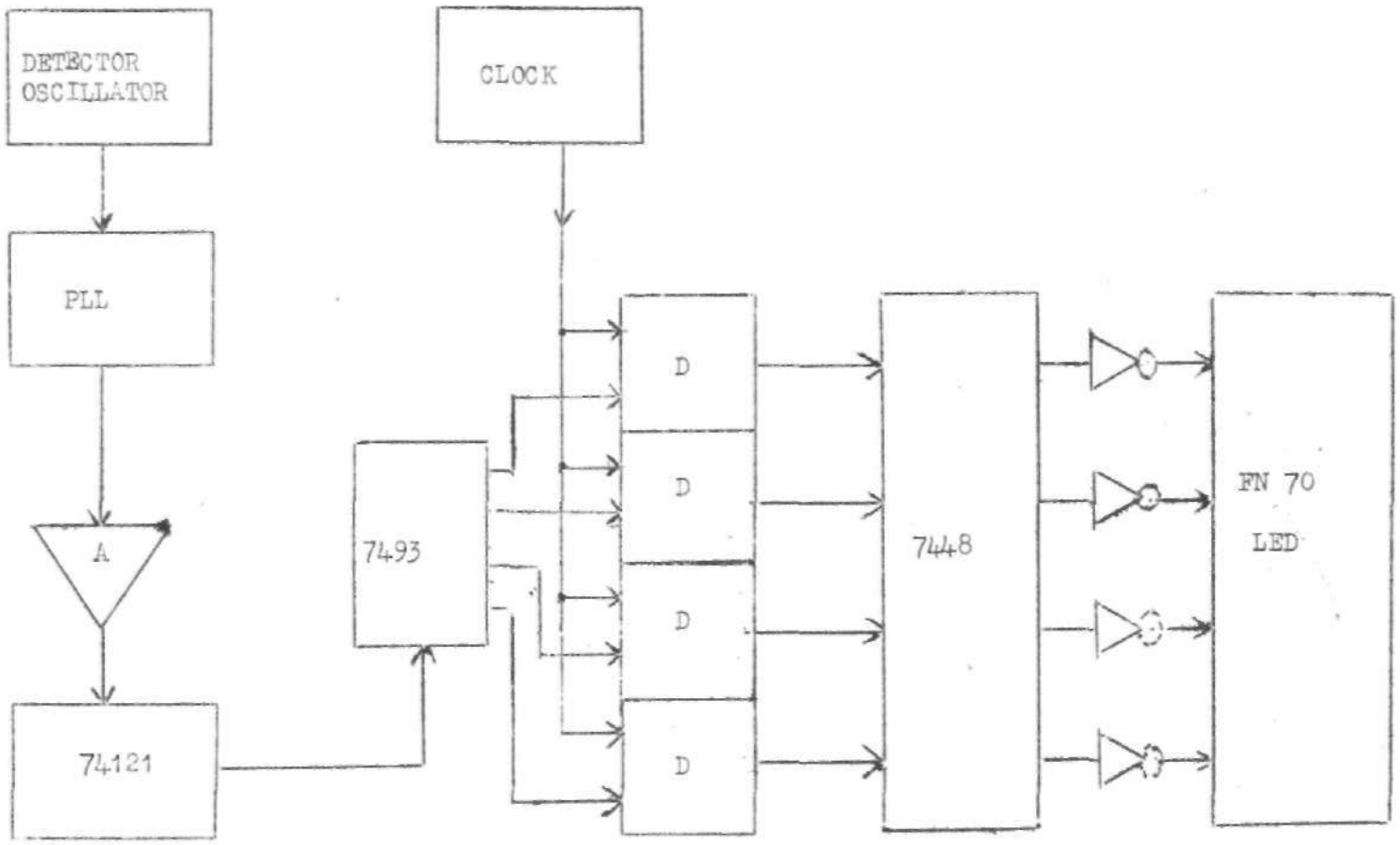


Fig. 4.13: Vehicle Detector Using LED Readout

CHAPTER FIVE

APPLICATIONS OF THE VEHICLE DETECTOR

5.0 Synopsis

The application of the vehicle detector to traffic survey and control is the subject of this chapter. It contains sections on Digital to Analog converter usuable with the X-Y recorder for the traffic data plotter. The concept of state machine is latter employed to design a model semi-vehicle-actuated controller.

5.1 Introduction

The versatility and the usefulness of the vehicle detector designed in the previous chapters will be demonstrated in this chapter by discussing two of its potential applications.

5.2 Application to Automatic Traffic Survey

In the last decade there has been a considerable increase in the volume of traffic on the national roads and the number of vehicles using the existing network is increasing more and more. This is undoubtedly related to the socio-economic activities of the society and as such effort must be made to see that maximum benefit is derived from automobile mobility. The authorities concerned need to know of the utility of existing networks and plan for future improvements. This involves collection of traffic data. This collection may be performed mechanically or manually for directional or non-directional, classified or unclassified cases.

A mechanical survey though may not be able to produce a readily classified data, it is undoubtedly prefered if long time and simulataneous monitoring of many locations are needed. The mechanical survey method involves the use of

vehicle detectors as given in chapter one. The number of vehicles that has crossed the Detector is stored in different ways viz:

(1) in mechanical counters, (2) in electronic counters and

(3) in graphical form.

The graphical presentation of traffic data has the advantage of showing the variational display of traffic per unit time.

Such a hard copy, graphical data could be obtained using a chart recorder together with a properly interfaced Vehicle Detector.

In this chapter, a hard copy of traffic data is made available with the Vehicle Detector already designed using a X-Y plotter. The schematic diagram of the Traffic Data Plotter (T.D.P.) is shown in fig. 5.1(a).

The number of Vehicles is counted by counter A and presents the value in digital form to the clocked parallel-in-parallel-out (PIPO) register. The time interval generator generates pulses at regular preset intervals. At the end of the interval the content of the counter is entered into PIPO register whose outputs are used for digital to analog conversion.

The X-Y plotter therefore experiences a Y-axis movement. The same pulse is delayed and used to reset the counter to zero.

Counter B records the number of time intervals. The output of Counter B is converted to analog signal so that the X-axis is marked in even spaces of time. Fig. 5.1(b) shows the timing diagram of the Traffic Data Plotter, and the flow chart is shown in fig. 5.1(c). Various components of the plotter are discussed in the following sections.

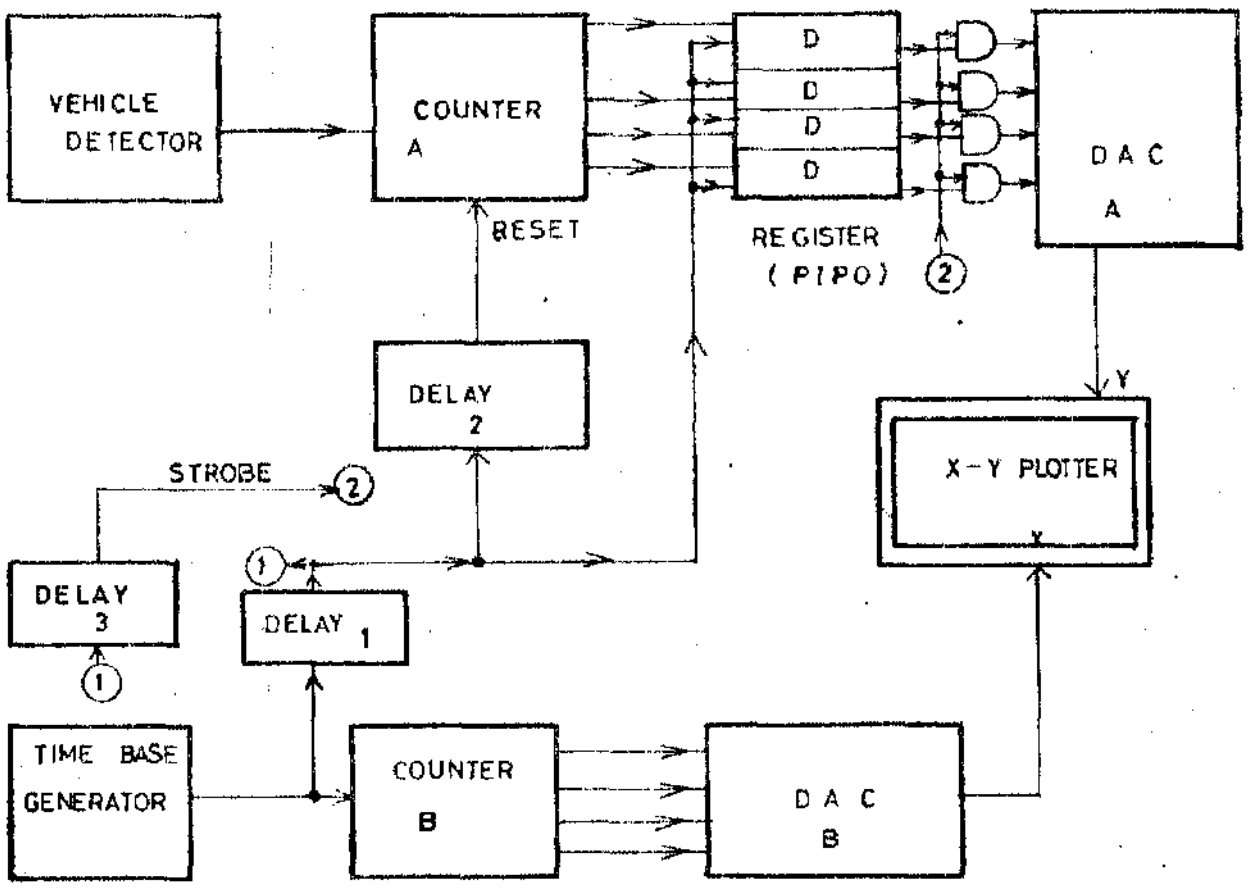


FIG 5.1A TRAFFIC DATA PLOTTER.

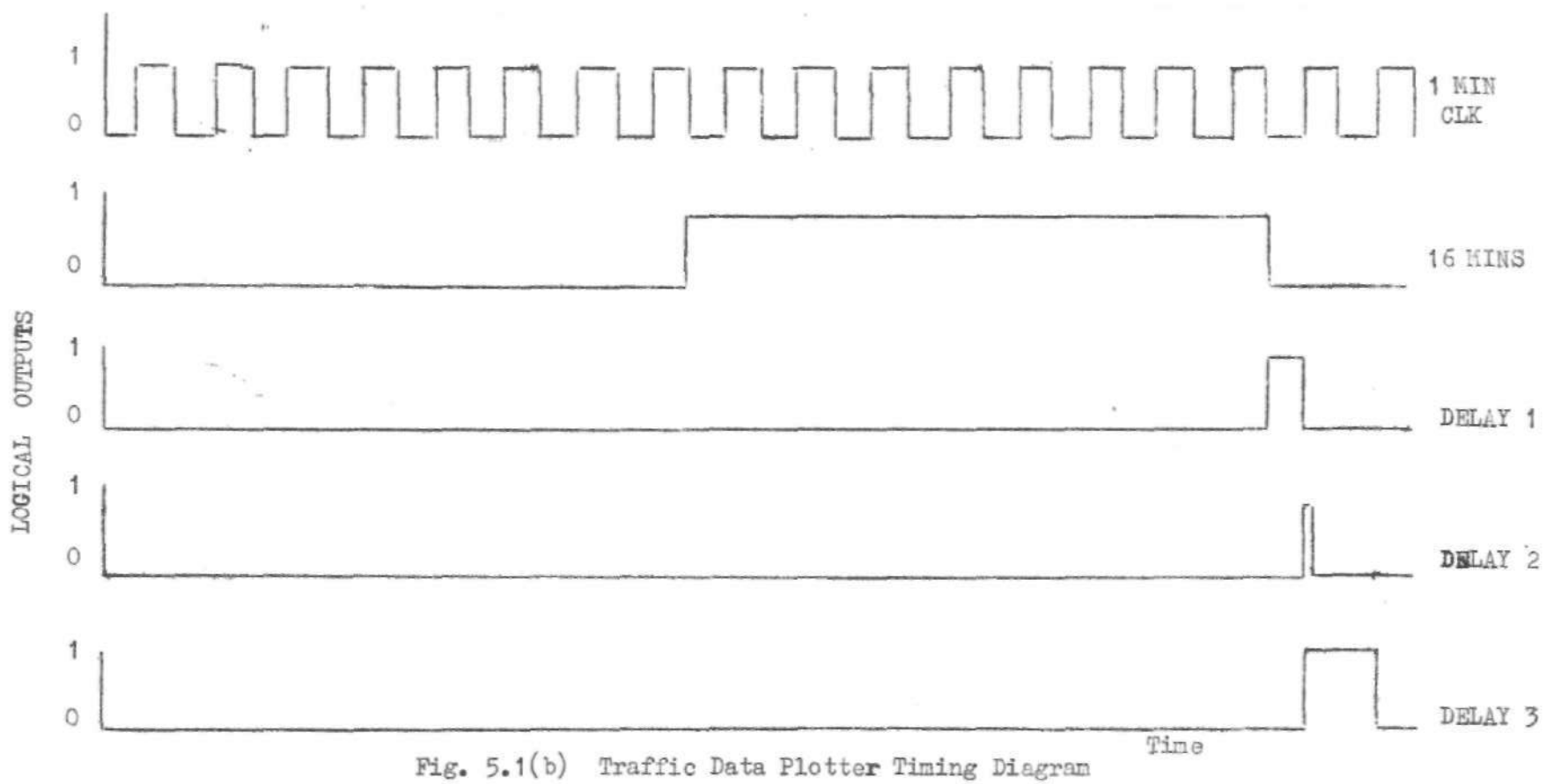


Fig. 5.1(b) Traffic Data Plotter Timing Diagram

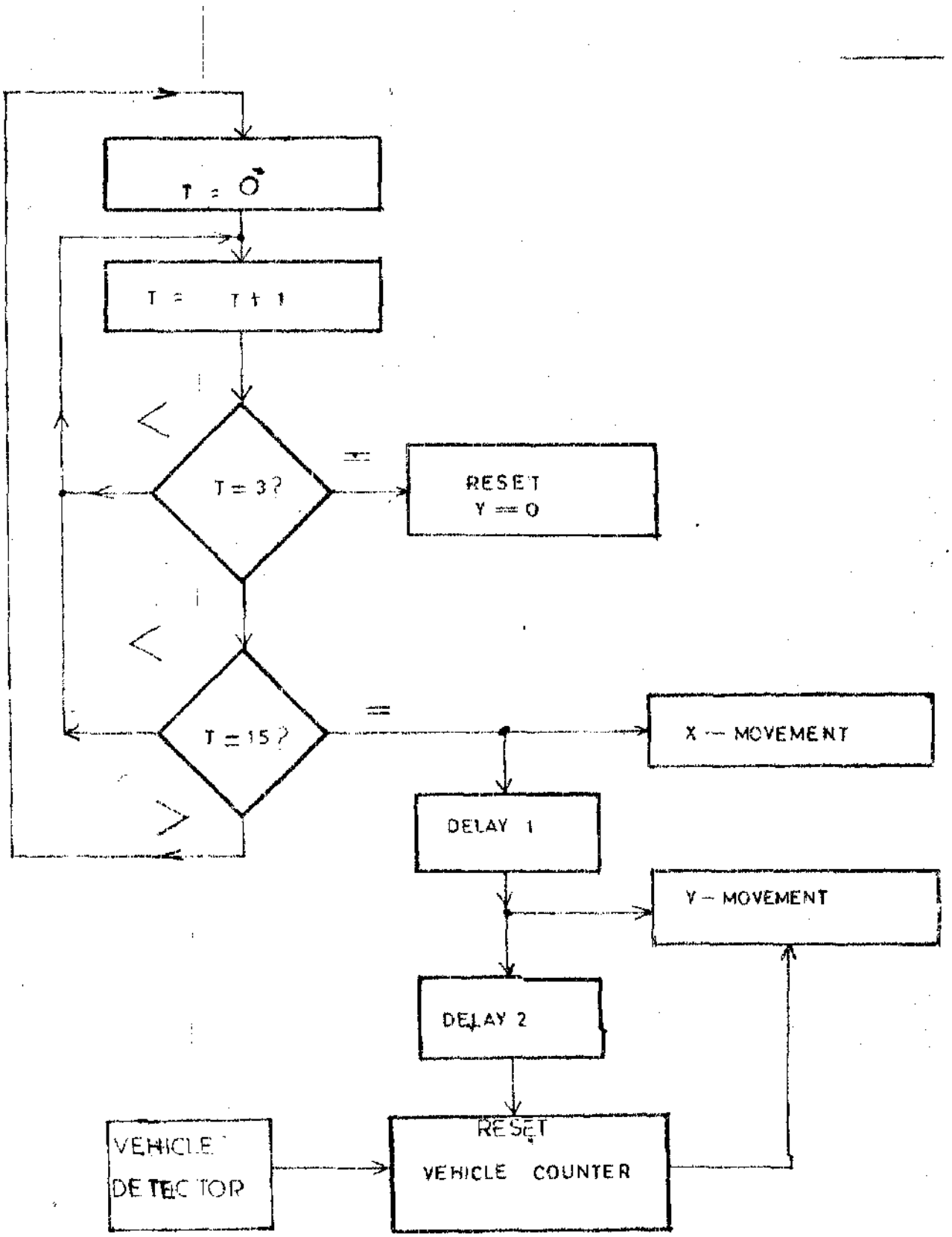


FIG 5 1C FLOW CHART OF TRAFFIC DATA PLOTTER

(b) Digital to Analog Convertion

Digital processing of information has many advantages[40] that efforts are on the increase to digitize most processes. The display of control and continuous systems with digital signals offer requires the generation of compactible analog signals. This procedure is what is known as Digital to Analog Conversion.

The Digital to Analog Converter, DAC accepts digital coded information and delivers a current or voltage proportional to the applied input. It possesses the ability to store information in digital form hence increasing immunity to drift and noise. A DAC can be designed using any of these methods:[41]

(1) the weighted resistor method

(2) the R - 2R ladder method

and the modification of the above which gives

(3) compromised weighted Resistor-Ladder method

(4) Inverted ladder method.

The general schematic diagram of a DAC is shown in fig. 5.2. The digital data is used to operate switches which connects a reference voltage V_R to the resistor networks. The contribution of each branch is then summed and amplified by the summing amplifier. The digital data is always stored in delay registers. This gives a stable analog signal

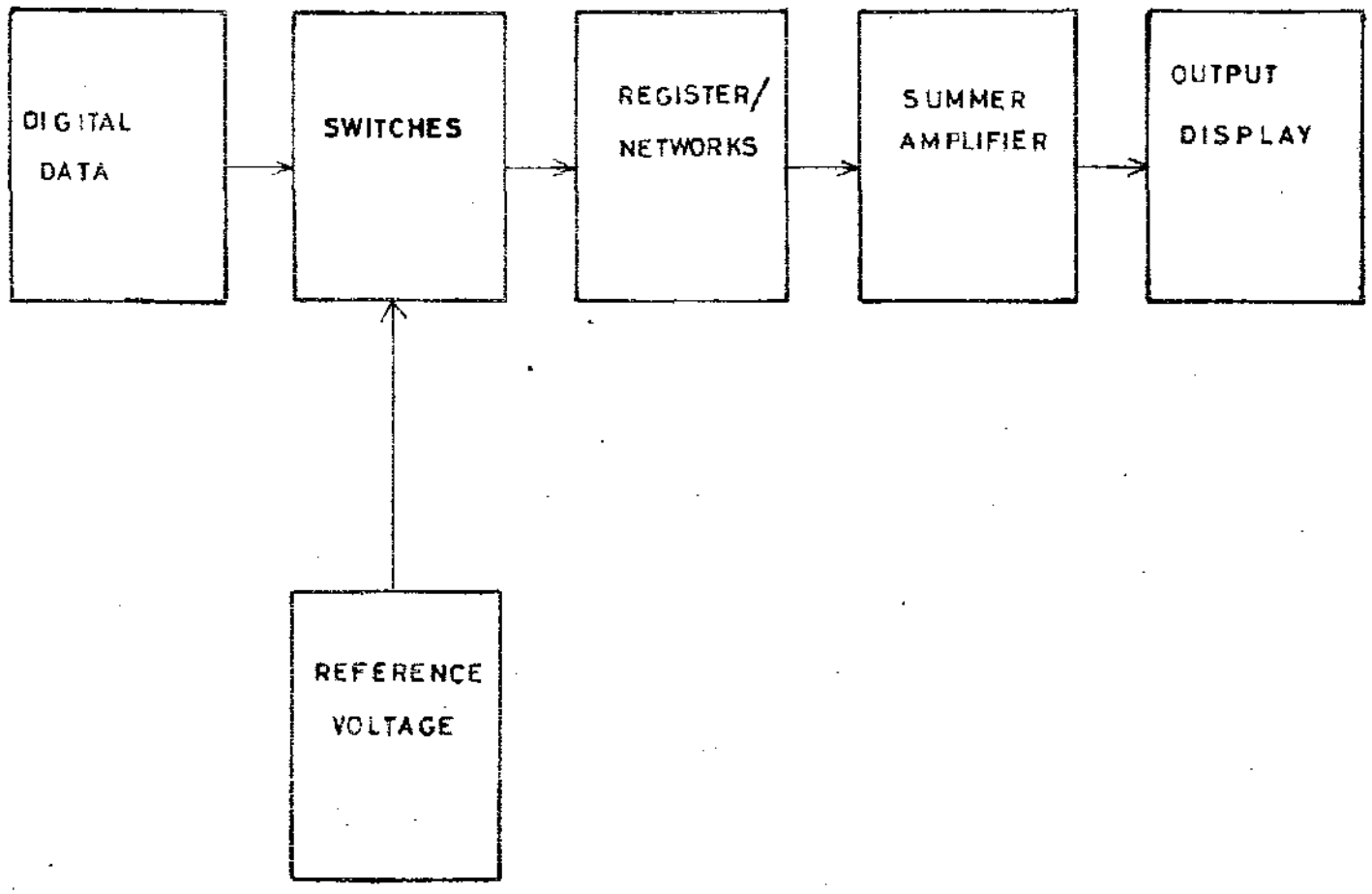


FIG 5.2 DAC BLOCK DIAGRAM.

and protects the information from noise and environmental effects.

The switches are double-pole type although diode and transistors types have been used,[40] the selection being dependent on its merits over others in particular applications.

(c) The Weighted-Resistor DAC[41,42]

The Weighted Resistor Digital to Analog Converter is shown diagrantically in fig. 5.3.

Generally a n - digit binary number

$$N = a_{n-1}\,2^{n-1} + a_{n-2}\,2^{n-2} + \ldots + a_1\,2^1 + a_0\,2^0 \quad \ldots\ldots \qquad 5.1$$

is used to operate the switches

$$S_{n-1},\ S_{n-2},\ \ldots\ldots\ldots\ S_1,\ S_0$$

and hence the reference voltage $V_{(1)}$ or $V_{(0)}$ is connected correspondingly to weighted resistors

$$R_i = 2^i R$$

for $i = 0$ to $(n - 1)$

where R is arbitrary and

$$R_0 = R$$

for most significant bit, m s b.

The current flowing in the summing element is

$$I = \frac{a_{n-1}\,V_R}{R} + \frac{a_{n-2}\,V_R}{2R} + \ldots\ldots \frac{a_1\,V_R}{2^{n-2}\,R} + \frac{a_0\,V_R}{2^{n-1}\,R}$$

$$= \frac{V_R}{2^{n-1}R} \sum_{i=0}^{n=1} a_i 2^i \qquad \ldots\ldots\ldots\ldots \qquad 5.2$$

The maximum current flows in the output when all switches are at '1' that is all a_i are '1' therefore (41)

$$I_{max} = \frac{V_R}{R} \frac{2^n - 1}{2^{(n-1)}} \qquad \ldots\ldots\ldots\ldots\ldots \qquad 5.3$$

This allows a discriminative selection of R to provide a certain maximum current. For a load of R_L, the new current is given by

$$I'_L = \frac{r}{r + R_L} I_L \qquad \ldots\ldots\ldots\ldots \qquad 5.4$$

where r is the output inpedance of the unloaded network which is the parallel conbination of all the weighted resistors

$$\text{i.e.} \qquad r = \frac{R}{2^n - 1} \qquad \ldots\ldots\ldots\ldots \qquad 5.5$$

The use of operational amplifier for summing is discussed later in the chapter.

(d) <u>The R-2R Ladder Converter.</u> (41,42)

One disadvantage of the weighted resistor DAC discussed in the last section is the wide range of resistor values to be used which is 2^{n-1} R for n-digit converter. In addition, the resistors must satisfy stringent tolerance limit if the syston is to function accurately. A method which reduces these complications is the R-2R ladder network.

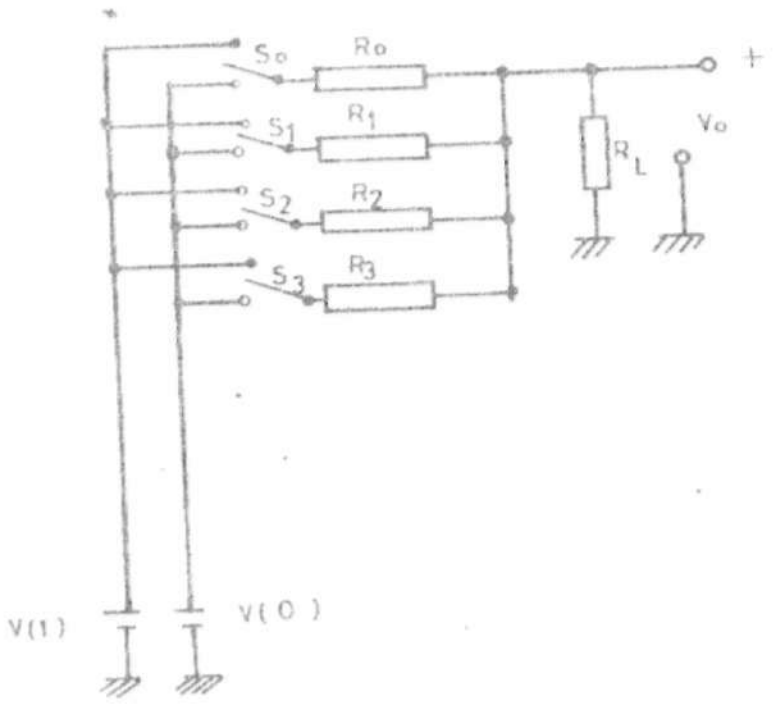


FIG 5.3 A WEIGHED RESISTOR DAC

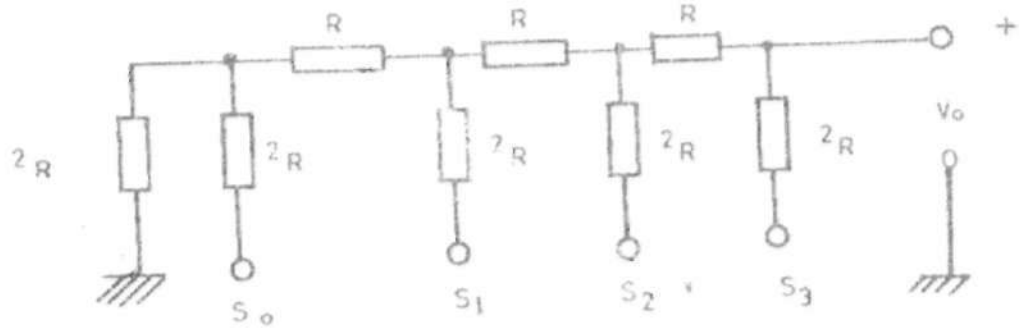


FIG 5.4 R — 2R LADDER DAC

It employs only two values of resistors R and 2 R, however, it requires twice the number of resistors used in the weighted-resistor DAC. A R - 2R ladder network is shown in fig. 5.4 for a 4-bit converter. It is a linear network where superposition theory is applicable for analysis. The contribution of each source to the output is independent of others and all contributions are linearly summed up. The switches have been represented by

$$S_3;\quad S_2,\quad S_1,\quad S_0$$

for a 4-digit number where S_0 is for least significant bit (lsb).
If

$$S_0 = 1$$

while

$$S_1 = S_2 = S_3 = 0$$

the voltage at the lsb node is $\frac{V_R}{2}$ and this voltage is attenuated by a factor of 2 at every successive node.

Consequently, the output voltage is

$$V_o = V_R \left(\frac{S_3}{2^1} + \frac{S_2}{2^2} + \frac{S_1}{2^3} + \frac{S_0}{2^4} \right)$$

$$= V_R(S_3\, 2^{-1} + S_2\, 2^{-2} + S_1\, 2^{-3} + S_0\, 2^{-4})$$

..... 5.6

or generally for n-digit number

$$V_o = V_R (S_{n-1} 2^{-1} + S_{n-2} 2^{-2} + \dots + S_1 2^{-(n-1)} + S_o 2^{-n}) \quad \dots\dots 5.7$$

$$= V_R 2^{-n} \sum_{i=0}^{n-1} S_i 2^i \quad \dots\dots\dots\dots 5.8$$

If the network is loaded by R_L, the relative contribution of each bit is not affected like in the weighted resistor network but the output is attenuated in a similar manner so that

$$V'_o = V_R 2^{-n} \sum_{i=0}^{n-1} S_i 2^i \frac{R_L}{R + R_L} \quad \dots\dots\dots 5.9$$

There is usually an inclusion of 2 R resistor which makes inpedance of each input 3R, the same value of resistor is used as the feedback resistor of the operational amplifier which is normally used to amplify the output of the ladder network.

e. <u>The Design Consideration for a Weighted-Resistor D/A C</u>

Despite the problems of the weighted resistor D/C, it was selected since a maximum of 4 bits is expected in which case the resistor range is 8 R.

The choice will also help in deciding a judicious value of R to provide a premeditated maximum output current. The output is amplified by an operational amplifier whose ideal representation is shown in fig. 5.5. The input impedance, Z_s is assumed ideally to be infinite and the input bias currents

$$I_{s\,1} = I_{s\,2}$$

and the output impedance

$$Z_o = 0$$

The point S is called the summing point. This is because for an amplifier with an infinite open loop gain A

$$V_s = 0$$

and point S is a virtual ground

$$I_1 = V_{in}/R_1$$

and $$I_f = I_1$$

For several currents at point S

$$I_f = \text{Sum of } (I_1)$$

and output voltage is

$$V_o = -R_f \sum I_1 \quad \ldots\ldots\ldots\ldots 5.10$$

For this design the following specifications are relevant.

(1) Op Amp supply voltage = ± 5 volts for uniformity with other TTL circuits.

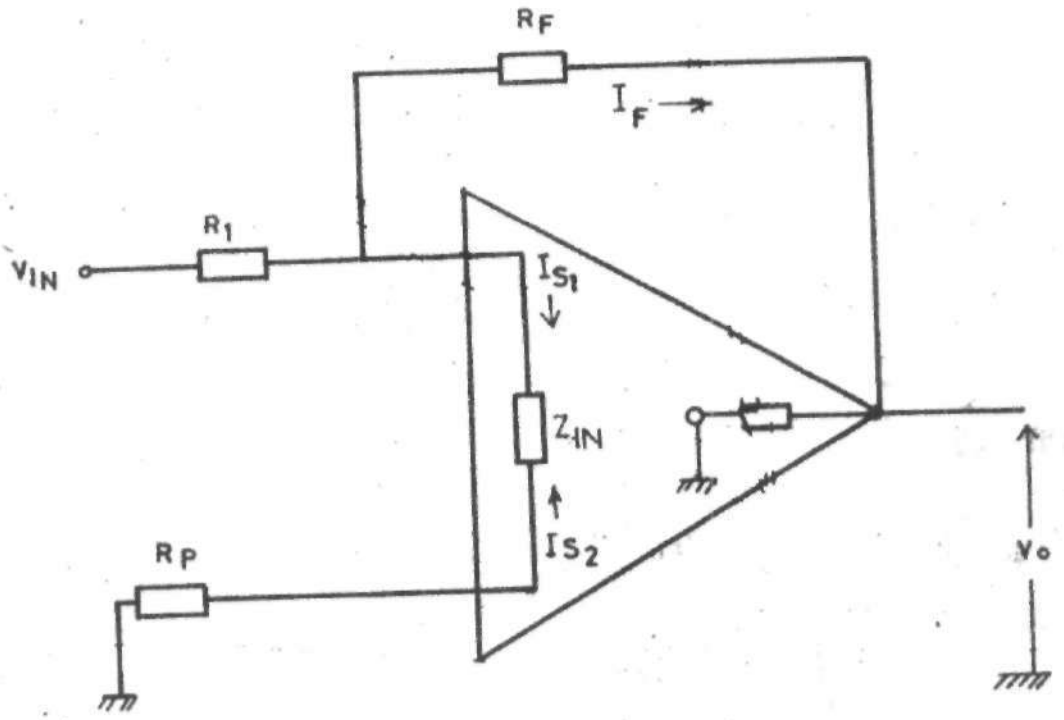


FIG 5.5 IDEAL OPERATIONAL AMPLIFIER

(2) Resolution = 6.2 % for a 4 bit word

(3) Maximum output at binary number 111 minimum = 4.0 volts

The minimum current at 0001 is given by (section b)

$$I_{min} = \frac{V_R}{2^{n-1}R} \quad \text{.....} \quad 5.11$$

for n-bit word.

Now for n = 4 and V_R = 5.0 V and I_{min} of 50 micro amps,

$$R = 12.5 \text{ kiloohms.}$$

and

I_{max} from equation 5.3 is

$$I_{max} = \frac{15 \times 5}{8R}$$

$$= 750 \text{ micro amps}$$

The feedback resistor is

$$R_f = \frac{V_o}{I_{max}}$$

$$= 5.33 \text{ kiloohms.}$$

$$V_{(0001)} = R_f \times I_{min} = 0.27 \text{ volts}$$

The resistance values are from msb to lsb

12.5, 25, 50, 100 in kiloohms.

The resistance R_p from the non-inverting input to the ground is such that the output error due to input bias current is zero.

Hence $R_p = R_1 // R_f$

f. Testing of the DAC

The weighted resistor DAC designed in the last section is tested with the apparatus set up of fig. 5.6. A pulse generator drives a ÷ 16 4-bit counter type TTL 7493. The outputs are stored in clocked parallel-in-parallel-out (PIPO) register assembled with D flip-flop type TTL 7474. The reference voltage was obtained from the output of TTL 7407 buffer/driver with open collector for uniform V_R. The 7407 gate also serve as switches. The output of the amplifier was measured as the count is advanced from 0000 to 1111. The results are plotted in the linearity curve of fig. 5.7. A display of the DAC voltage/bit transfer characteristic was also obtained on the scope by driving the counter at about 1 kilo Hz, plate No. 3

If the linearity of the converter is ϵ
then

$$|\epsilon| \leq \tfrac{1}{2}\Delta$$

where Δ = lsb.

From the graph Δ = 0.18
and ϵ = .04

It can be seen that since

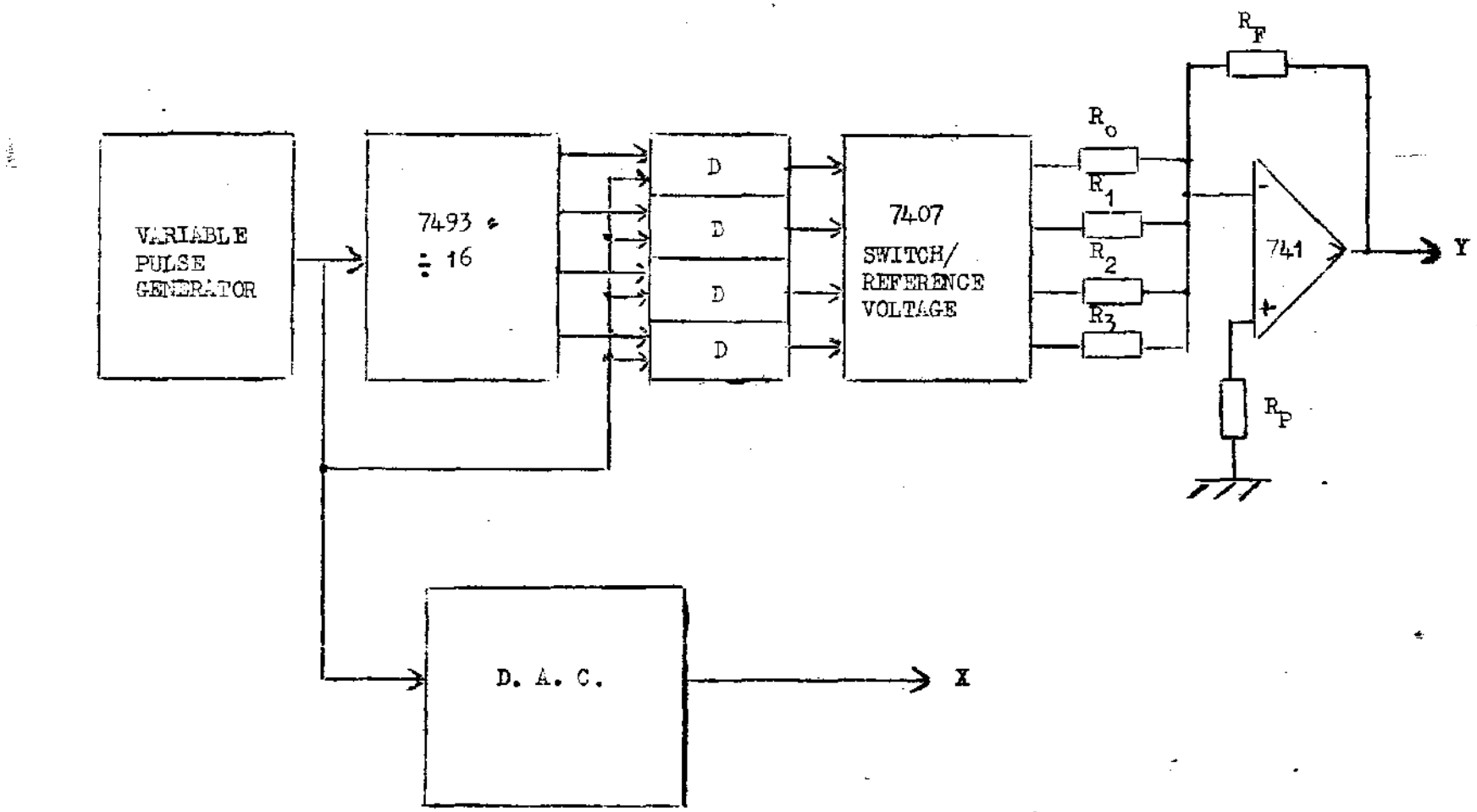
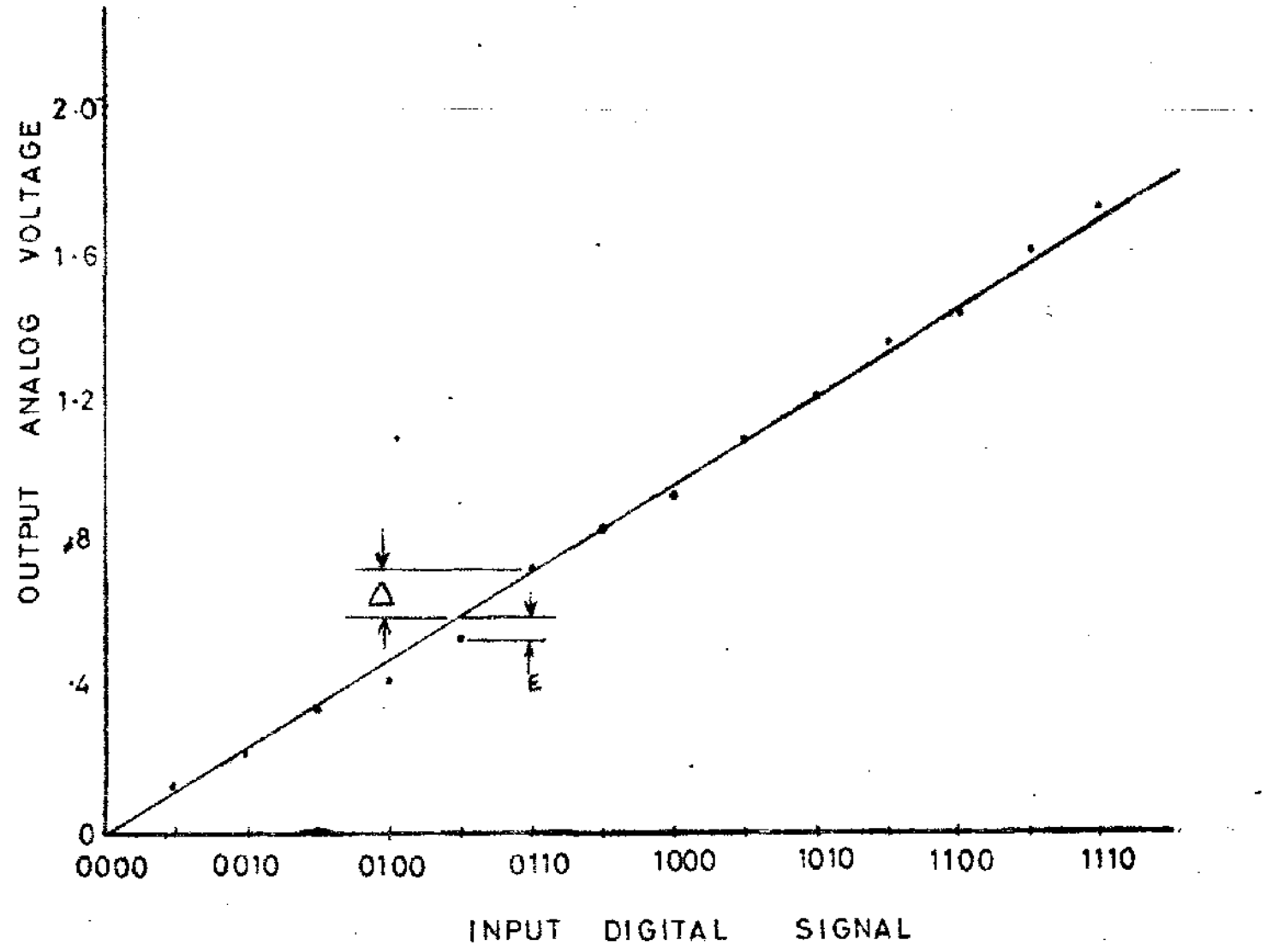


Fig. 5.6: Testing the D.A.C.

Fig 5·7 LINEARITY CURVE OF DAC



the DAC is acceptable and monotonic. The accuracy is defined as the difference between the actual analog voltage output and the ideal case is equal to

$$\frac{0.03}{4} \quad \text{or} \quad 0.75\ \% \text{ of full scale.}$$

g. The Time-interval Generator

The time interval generator is an astable multivibrator (AMV) followed by a frequency divider. The AMV is built around the 555 timer, (39) fig. 5.8. The capacitor C charges through R_1 and R_2 in series when power is switched on with a time constant of

$$C\ (R_1 + R_2)$$

Until 2/3 of the supply voltage is reached when the threshold comparator flops the internal circuitry so that C now discharges through R_2 with a time constant $C\ R_2$. The output remains low as C discharges to 1/3 of supply voltage when the circuit flips back to original state. The cycle then repeats itself. The frequency of operation is determined by R_1, R_2 and C.

The design equations for the periods are

$$\tau_{charge} = 0.7\ (R_1 + R_2)\ C$$

and

$$\tau_{discharge} = 0.7\ R_2\ C$$

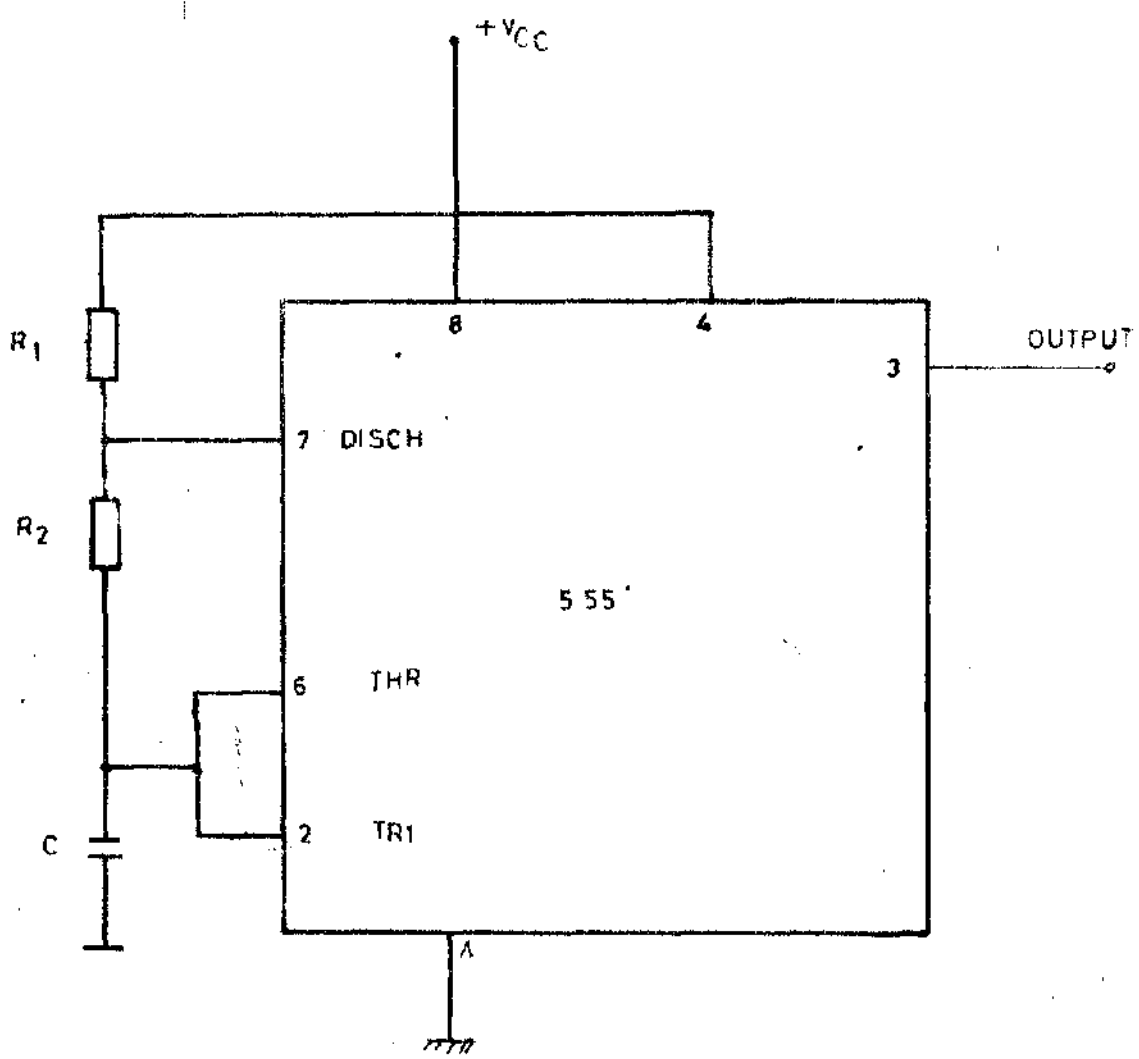


FIG 5.8 THE 555 TIMER PULSE GENERATOR

when C is charging the output is positive and grounded when C is discharging.

For a time of 60 seconds

$$(R_1 + R_2)C = 60$$

Let C = 47 microfarad

R_2 is selected to be small so that the discharge time is small and the negative edge pulse is sharp.

Discharge time is taken as 400 nS.

$\therefore$ R_2 = 1.2 k

and R_1 = 1.82 mega ohms

R_1 was constructed using a fixed 1.0 megaohm register together with correct trimmer.

The output of the timer can be divided to give 2, 8 or 1 · minutes by using the counter TTL 7493.

5.3 A Semi-Vehicle Actuated Traffic Controller

The need for efficient control of Urban traffic has been stressed in chapter one where reference was made to the use of automatic traffic controllers as being economic and versatile. Mention was made also of the advantages of a Vehicle-Actuated Controller over fixed time ones.

In this section, a Semi-Vehicle-Actuated Traffic Controller is designed with the Vehicle Detector already made. A full fledge Vehicle-Actuated Controller (44) is in the pipe line.

a. Control Algorithm

The Semi-Vehicle-Actuated Traffic (SVAT) Controller is designed for a T-junction. It is aimed at giving an improved performance over a fixed time controller.

The Vehicle Detector is to be installed on the side road alone. The split ratio of the right of way between the main road and the side road is 3 : 1. This is designed to reduce the accident rate associated with SVAT signallised junctions due to spontaneous interruption of the main road. In addition, when the detector circuit is faulty, the system will operate as a fixed time one. If there is a Vehicle on the side road after the Main road has taken a preset minimum green period, the right of way is transfered to the side road. The right of way is to be extendable for a maximum time if more vehicles arrive from the side road. A turn left provision is to be included to cater for traffic coming into the side road from the main road.

b. Implementation

The SVAT signals are implemented as shown diagramatically in fig. 5.9. This consists of a Vehicle Detector which is installed on the side road at a distance from the stop line related to the type of junction. The

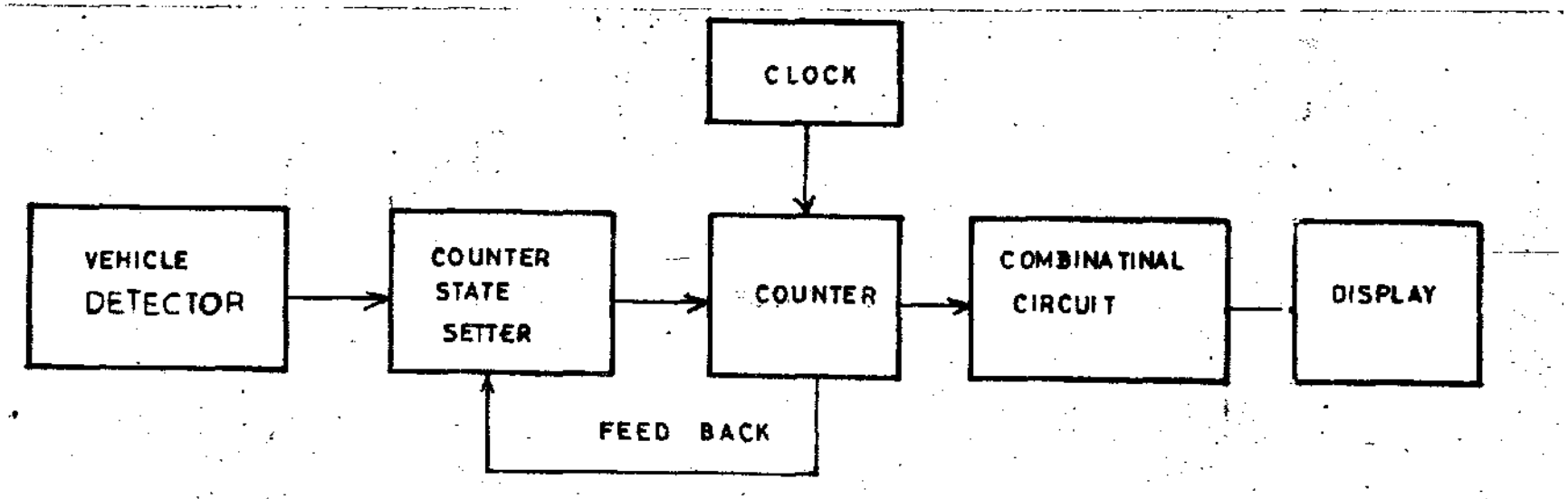


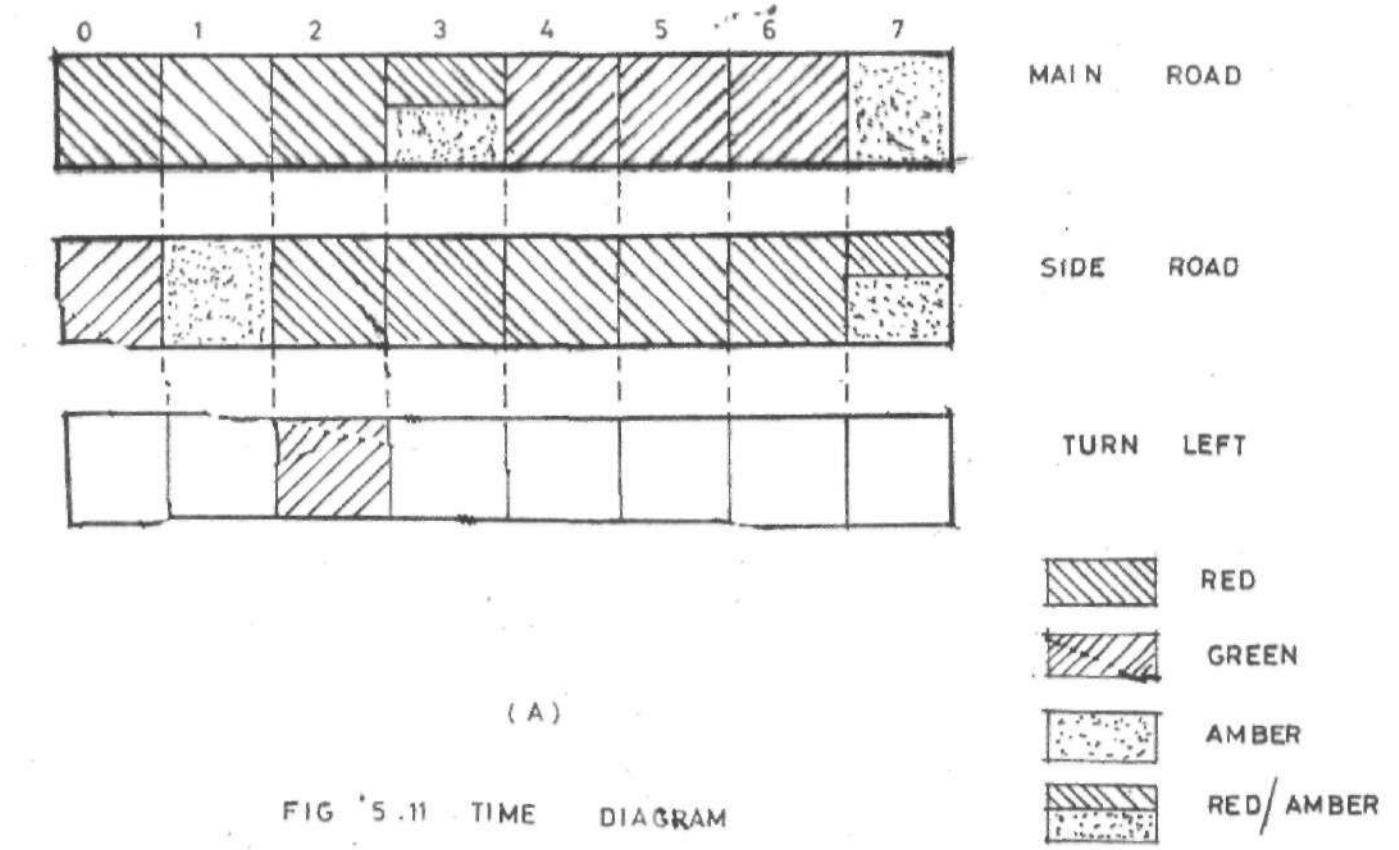
FIG 5.10 S.V.A.T. CONTROLLER IMPLEMENTATION

recommended distance[2] is shown below. For normal junction, the distance is about 40 meter and for junctions with narrow or winding approaches where speed is less than 50 Km/hr, the distance should be between 31 and 37 meters.

The Vehicle Detector feeds a decision circuit which sets the counter state according to relative traffic density on the side and main roads. The combinational circuit decodes the counter output for the display. The specification can be represented by the time diagram of fig. 5.10 for a T-junction - fig. 5.11 using an eight-state machine.[3, 43]

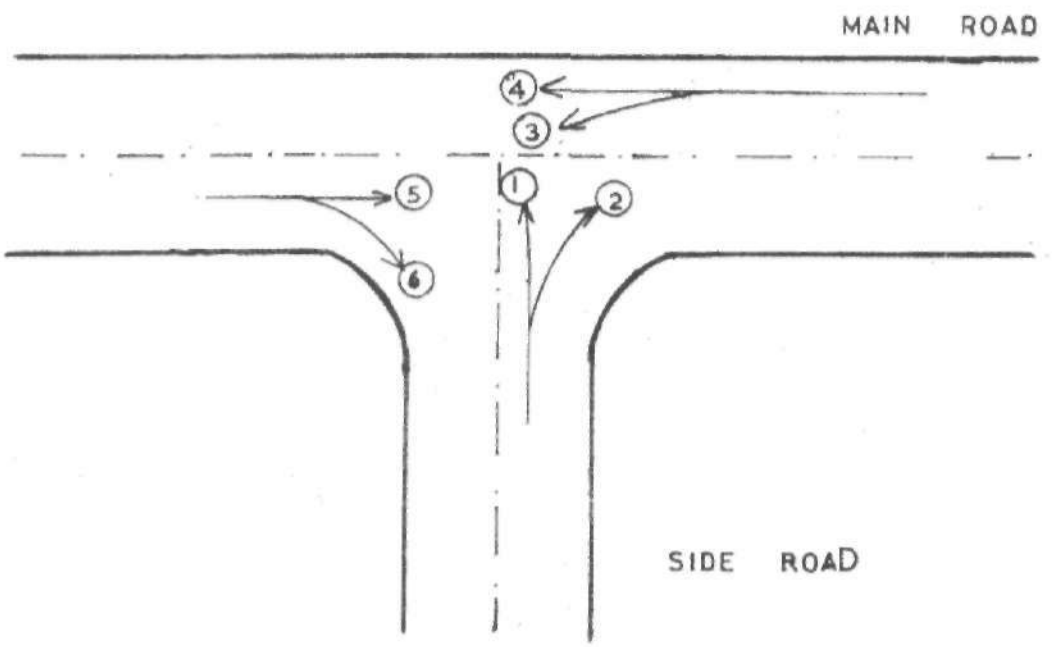
The algorithmic operational flow chart[43] of the eight-state machine is shown in fig. 5.12 and the Truth Table of operation is shown in fig. 5.13. An eight-state machine can be realised by using three flip-flops in series which generates the outputs shown in the Truth Table.

Since only three boolean functions are involved, the truth table can be minimised by using Karnaught map[41, 42, 43] method. The results are shown below.



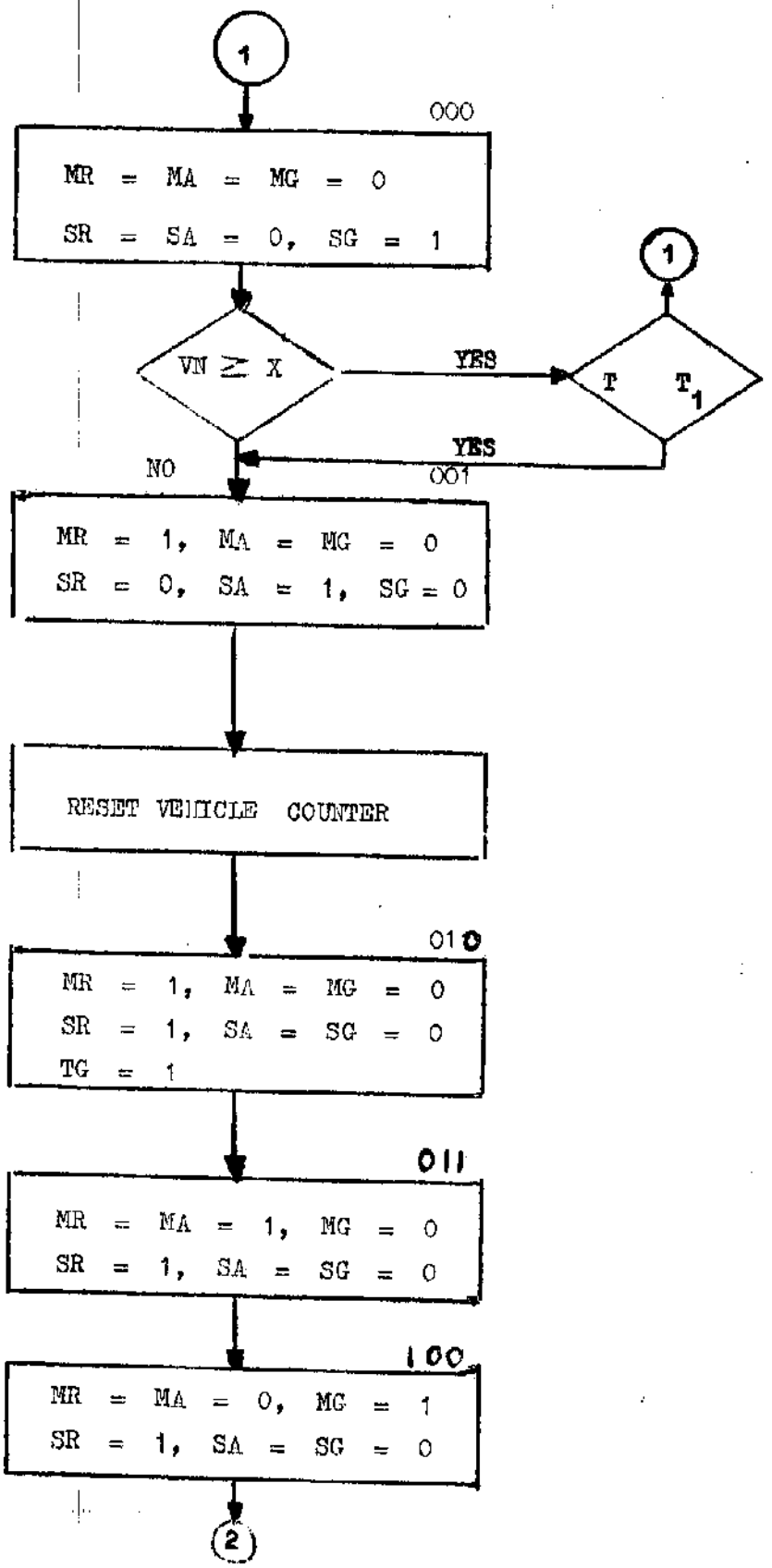
(A)

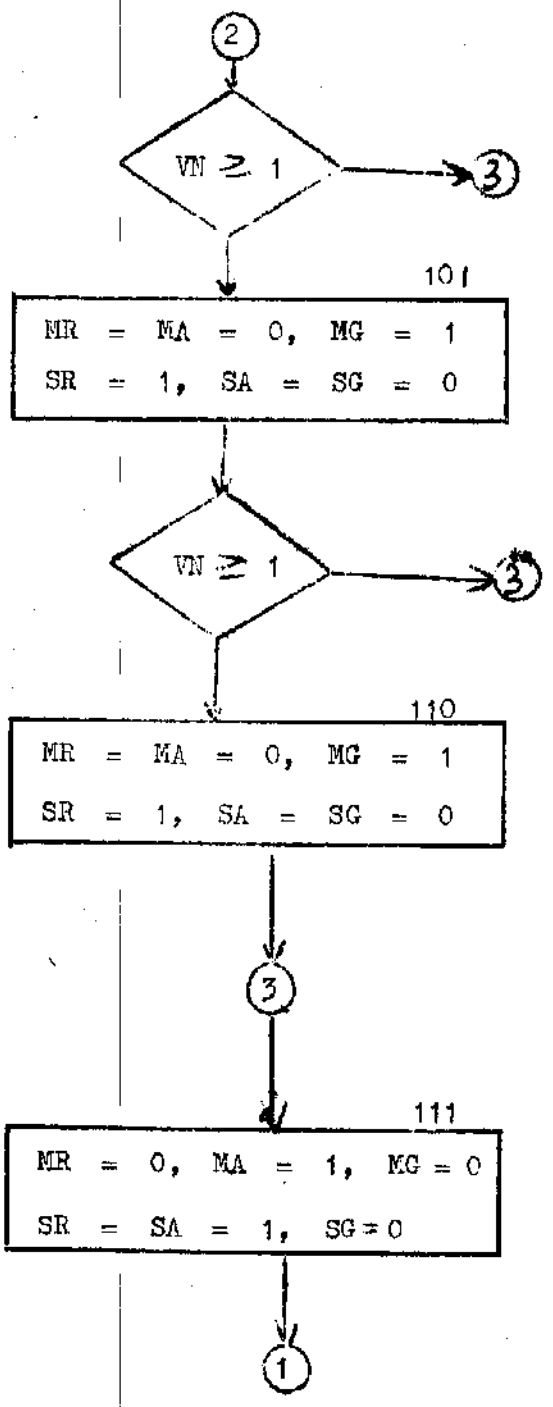
FIG 5.11 TIME DIAGRAM



(B)

FIG 5.12 T JUNCTION


1
000
MR = MA = MG = 0
SR = SA = 0, SG = 1
VN ≥ X
YES
1
T
T1
YES
NO
001
MR = 1, MA = MG = 0
SR = 0, SA = 1, SG = 0
RESET VEHICLE COUNTER
010
MR = 1, MA = MG = 0
SR = 1, SA = SG = 0
TG = 1
011
MR = MA = 1, MG = 0
SR = 1, SA = SG = 0
100
MR = MA = 0, MG = 1
SR = 1, SA = SG = 0
2



MR = MAIN ROAD RED
MA = " " AMBER
MG = " " GREEN
SR = SIDE ROAD RED
SA = " " AMBER
SG = " " GREEN
TG = TURN LEFT GREEN
VN = NO. OF VEHICLES
T_1 = MAXIMUM EXT. TIME

Fig. 5.12: Algorithm State Machine Flow Chart

Machine State	Counter State			OUTPUTS						
				Main Road			Side Road			Turn Left
	A	B	C	MR	MA	MG	SR	SA	SG	TG
0	0	0	0	1	0	0	0	0	1	
1	0	0	1	1	0	0	0	1	0	
2	0	1	0	1	0	0	1	0	0	1
3	0	1	1	1	1	0	1	0	0	
4	1	0	0	0	0	1	1	0	0	
5	1	0	1	0	0	1	1	0	0	
6	1	1	0	0	0	1	1	0	0	
7	1	1	1	0	1	0	1	1	0	

Fig. 5.13: Truth Table of the Machine

MR

AB / C	00	01	11	10
0	1	1		
1	1	1		

$MR = \bar{A}$

M A

AB / C	00	01	11	10
0				
1		1	1	

$MA = BC$

M G

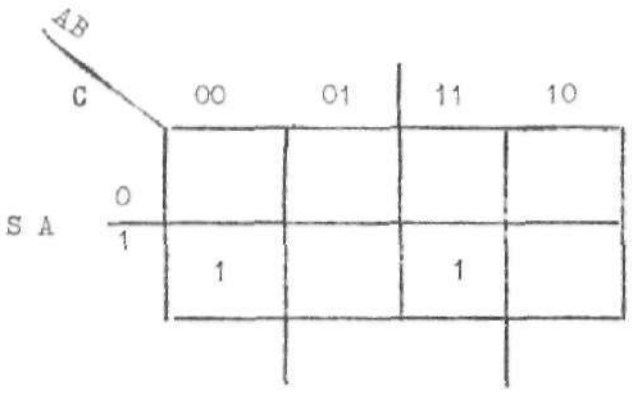
AB / C	00	01	11	10
0			1	1
1				1

$MG = A\bar{C} + A\bar{B}$

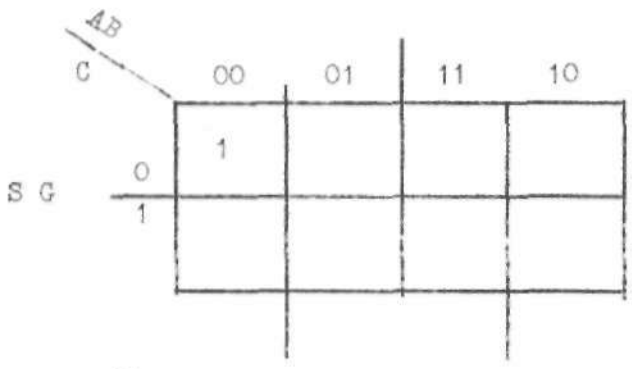
S R

AB / C	00	01	11	10
0		1	1	1
1		1	1	1

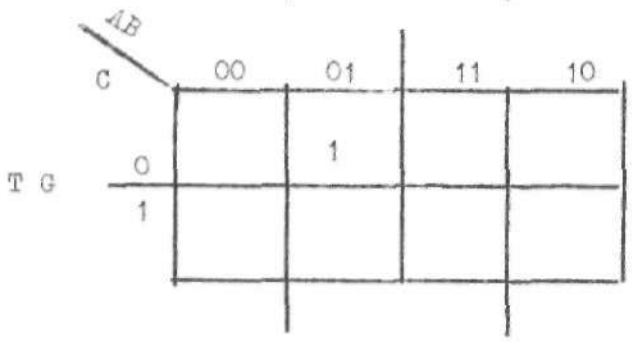
$SR = A + B$



$$S_A = \bar{A}\,\bar{B}\,C + A\,B\,C$$



$$S_G = \bar{A}\,\bar{B}\,\bar{C}$$



$$T_G = \bar{A}\,B\,\bar{C}$$

The combinational functions are realised with gates as shown in fig. 5.14

c. Counter State Setter

The circuit so far designed will produce a fixed time signal for the control of a T-junction with a 'left-turn' signal included. To achieve

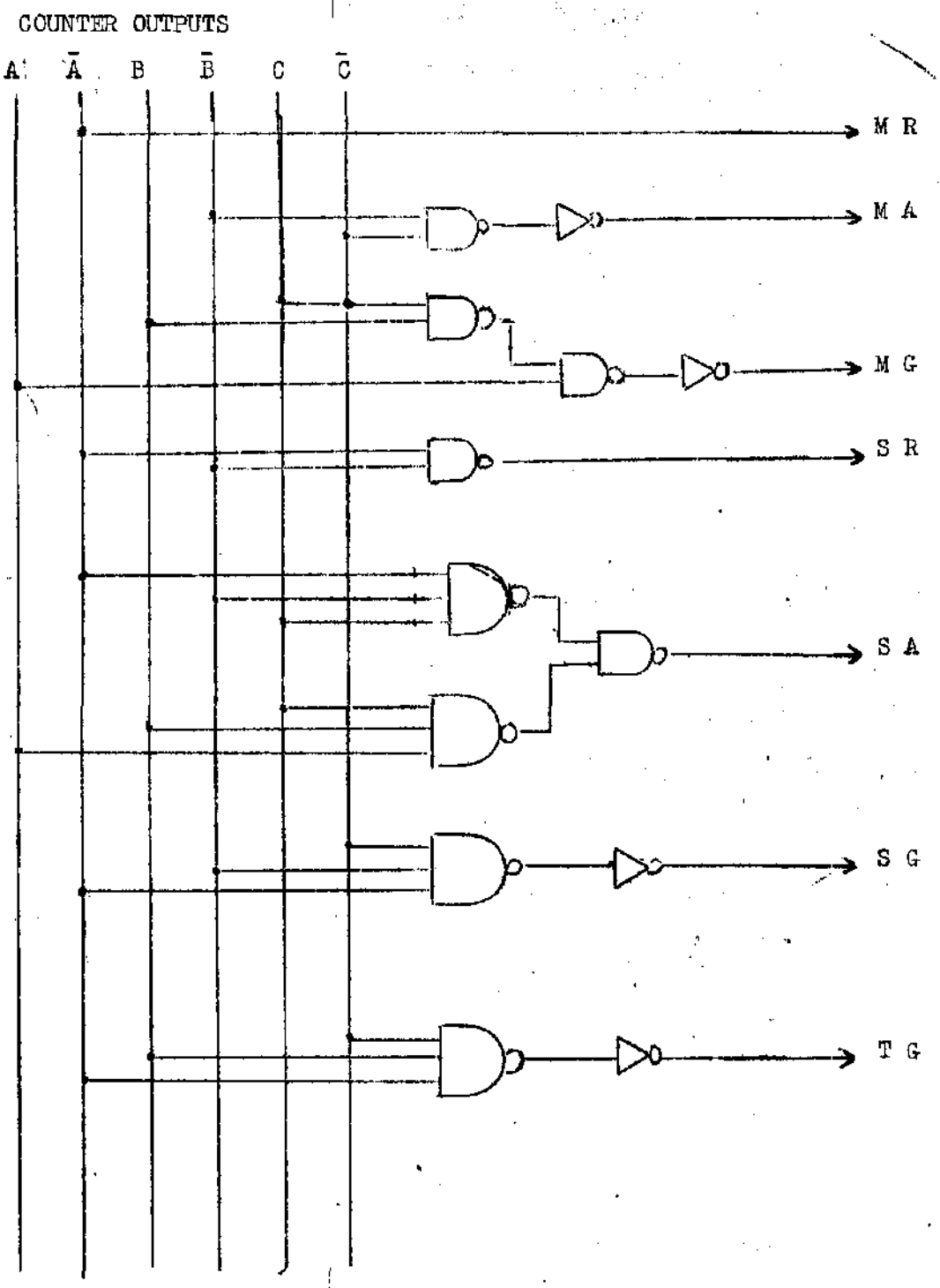


Fig. 5.14: DECODER SYSTEM

the flexibility already defined in section (a) a machine (counter) state setter is designed which works in conjuction with the Vehicle Detector.

The schematic diagram of the controller is shown in fig. 5.15. A study of this diagram with fig. 5.12 will provide a quick understanding of its operation.

With no demand signal from the side road, the machine controls the traffic with a split ratio of 3 : 1, the actual time allocation being dependent on the clock frequency. A demand from the side road when the machine is in state 5 to 6 transfers the right of way to the side road. This right of way is extendable for a predetermined period provided the number is greater than that specified for normal operation.

c. Assembly

A prototype display circuit is assembled using LED for this display. In all about 12 chips of integrated circuits consisting of 23 gates, 6 flip-flops and 3 monostable multivibrators have been used. This number of simple chips will facilitate fault finding and maintenance.

Summary:

This chapter has considered the application of the Traffic Detector designed to two traffic problems - traffic data collection and traffic control at intersections. It has demonstrated the versatility of the

Loop Detector which can be employed all the year round for monitoring and control of traffic both in urban centres and at any other locations of interest. In the next chapter, the linitations of, and suggestion for inprovement on this project are given in addition to the conclusions.

CHAPTER SIX

CONCLUSIONS

(a) Limitations of the Project

The design of the Traffic Detector using a phase lock loop as presented in this thesis has proved to be a useful instrument in solving one of the socio-economic problems associated with automobile movement. The concept with its potential for responding to varying environmental conditions is a very important step towards adaptive traffic control in Nigeria. Nevertheless, there are some limitations. The coil is buried and as such not easily mobile. If the Detector is needed for sample data from location to location cost and time of installation can be a limitation The buried Detector coil however cannot be readily attacked by rodent and children although it may be dug up and destroyed by unsuspecting road maintenance workers. If the integrated circuit Detector Oscillator is used, temperature of operation is a limitation. The temperature should be less than $70^{\circ}C$ for commercial integrated circuits. The Detector circuit has no means of detecting a stationary vehicle presence on the Detector Coil. The traffic data collection using the method described in the thesis is undoubtedly simple but like automatic devices of the type, it cannot classify readily the types of vehicles using the route.

(b) Performance Improvement and Suggestions for further Studies

The various limitations of the Vehicle Detector have been listed in the last section. The possible means of improving the performance of the Detector are discussed here.

The variations of the Detector Oscillator frequency with temperature have been shown to be strongly dependent on the coil preparation. The insulation of the coil from moisture appeared to be very critical, therefore the coil could be first covered with polythene, before other procedures described earlier. Wax could be substituted for polythene, it could also be covered with asphalt. In addition, the slot housing the coil could be cemented. These additional preparation will tend to increase the cost of installation but will provide a more stable oscillator.

The complete analysis of the Oscillator using the nonlinear element is an area where further investigation could be carried out by using the computer. The relationship between the influence of various components or the oscillator performance could then be evaluated more accurately.

Various thresholds could be defined for the output of the Detector in such a way that a classification could be made in terms of bicycles, two-axled and multi-axled vehicles. A decision circuit can also be included which will indicate the continuous presence of a vehicle on the coil as in cases of break downs, and parkings. The Detector

can also be adapted for other traffic measurements such as speed, with two detectors the speed of a vehicle could be determined.

(c) Conclusions

In this thesis the design of a Traffic Detector using the phase lock loop technique was presented. The reasons for the selection of the concept have been outlined and supported by practical results. The results of the project have demonstrated that the Detector is quite suitable for solving the defined problem. The two application areas discussed further illustrate the versatility of the Detector.

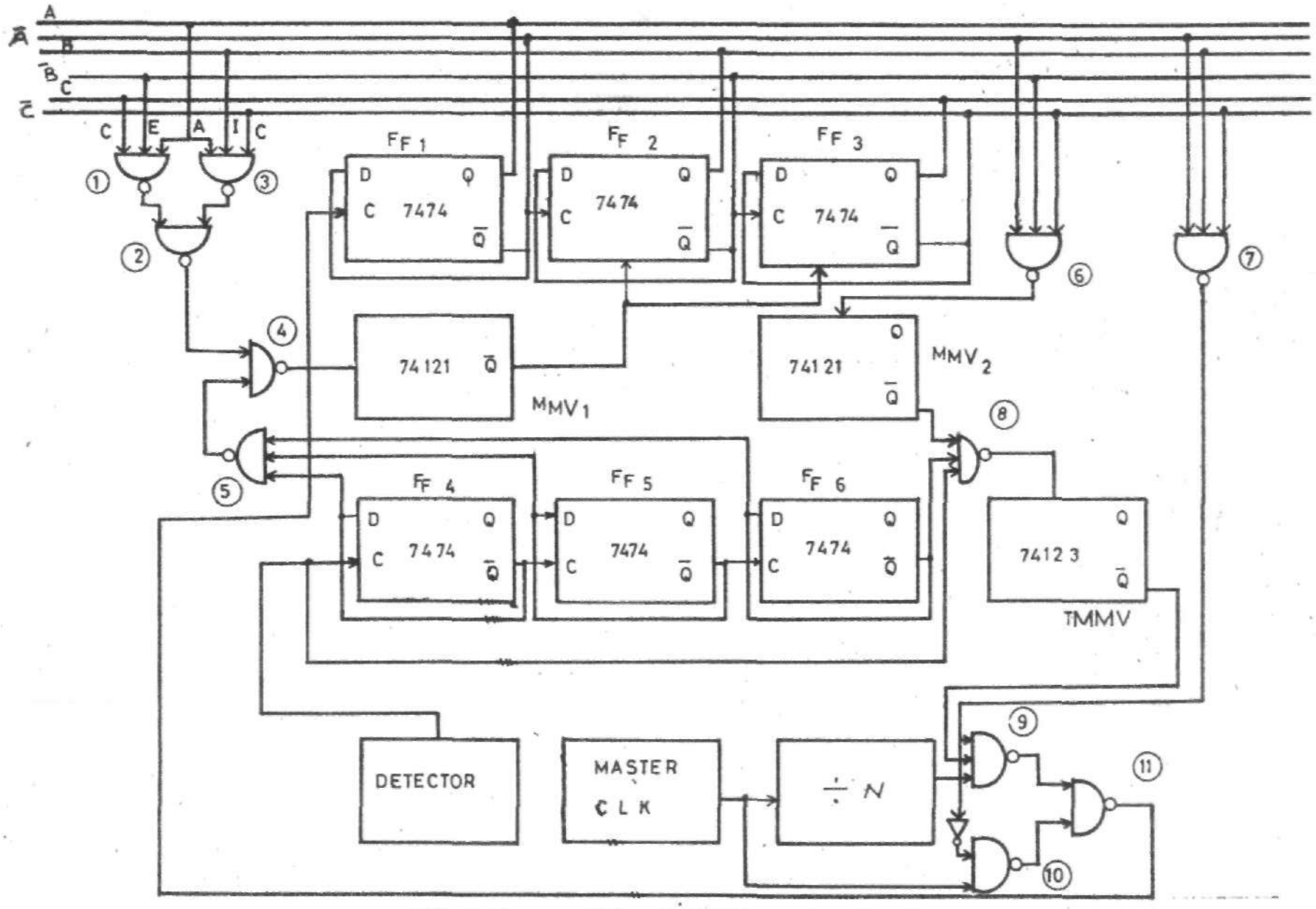


FIG 7.10 THE TRAFFIC CONTROLLER

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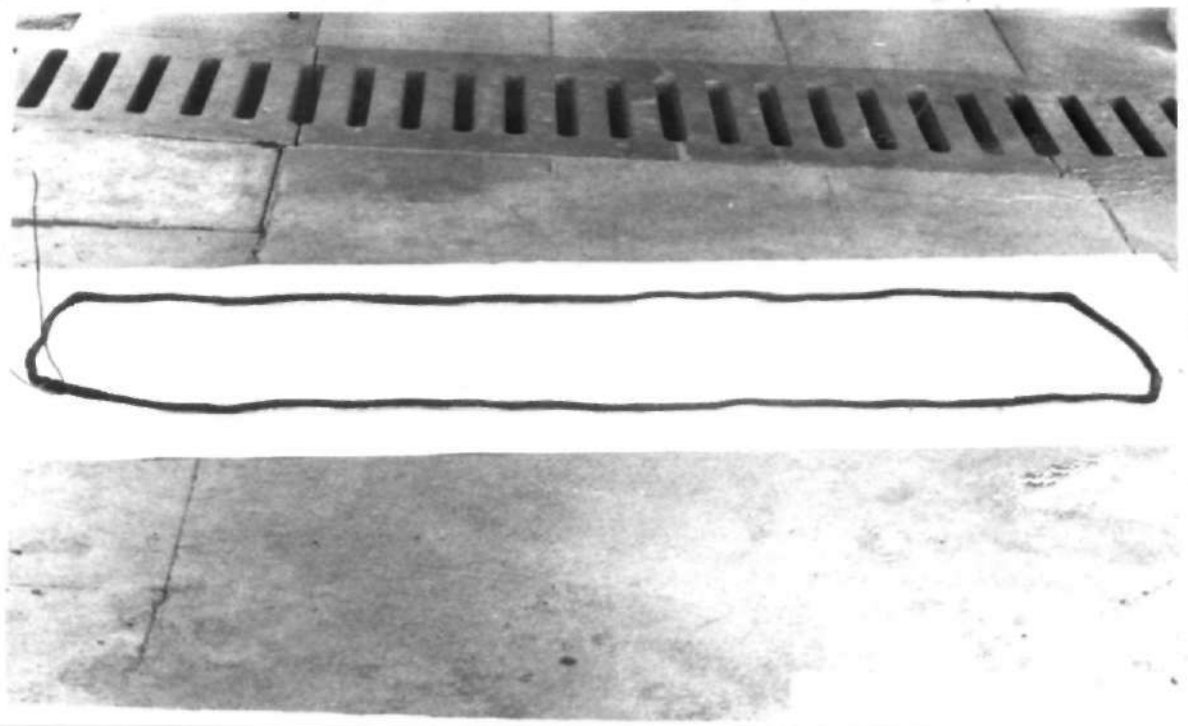


Plate No.1(b): The Coil

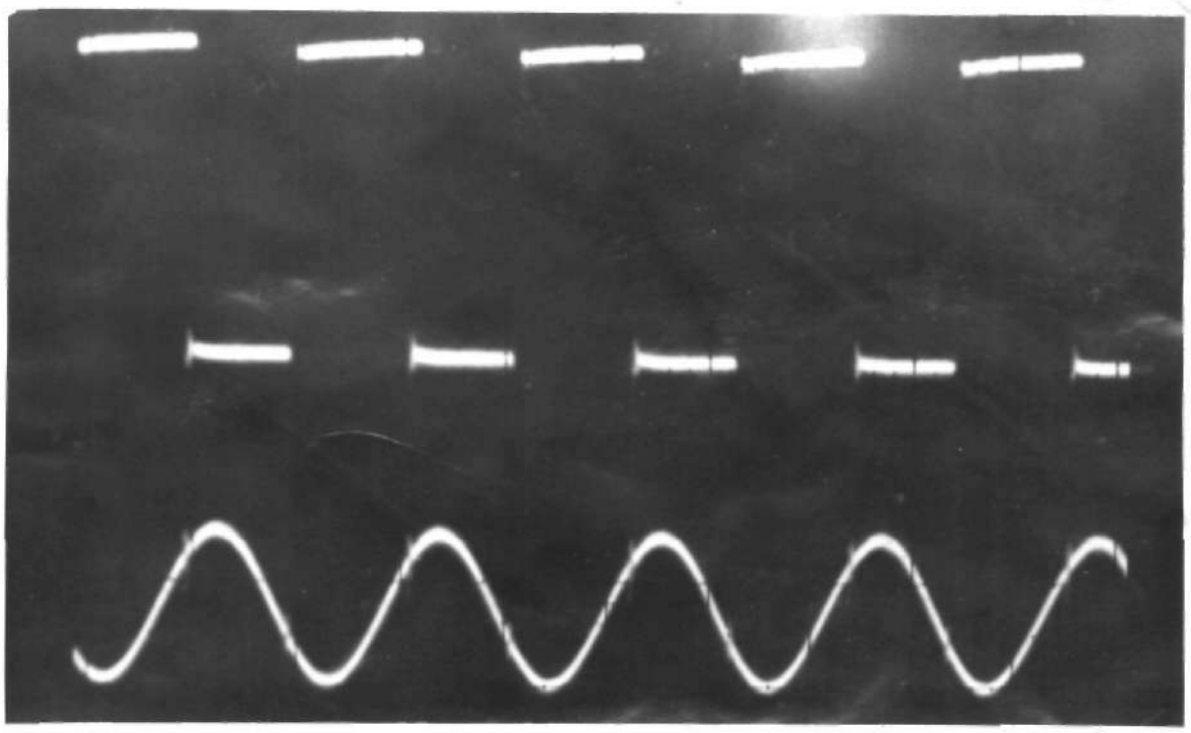


Plate No.1(b): Oscillator Using TTL 7414

Bottom: Output of Filter

Top : Squared by another gate

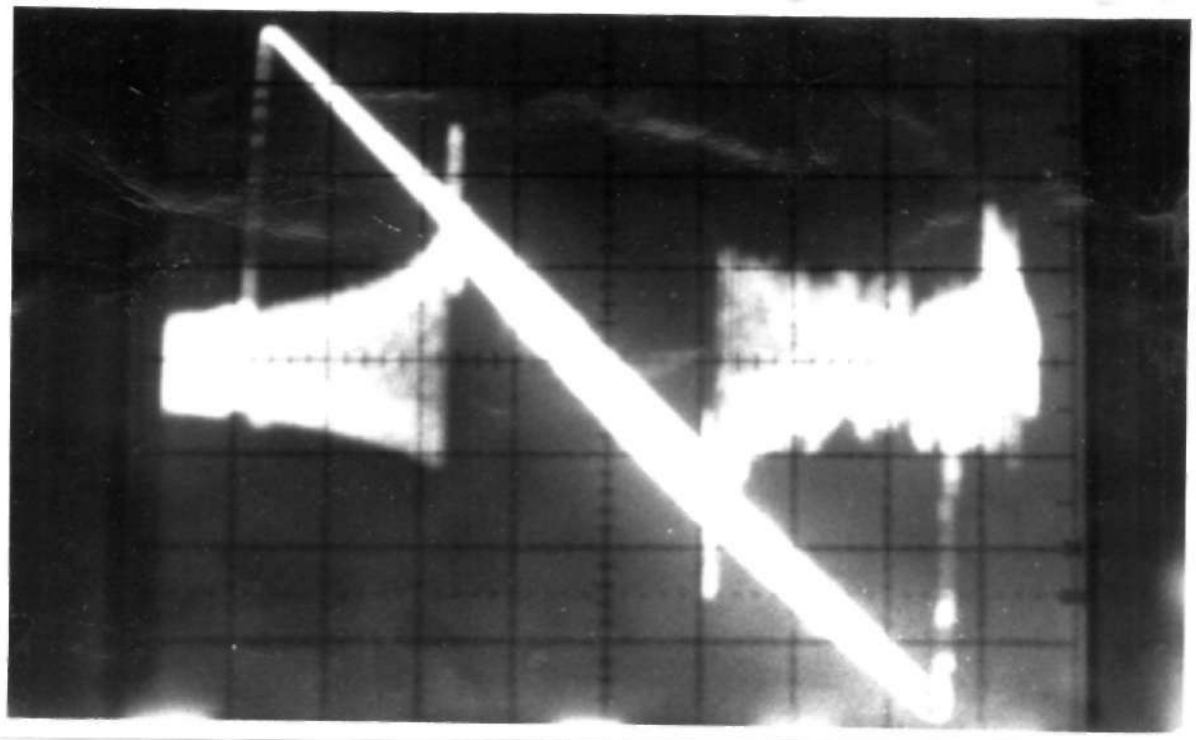


Plate No.2: Lock and Capture Range of PLL

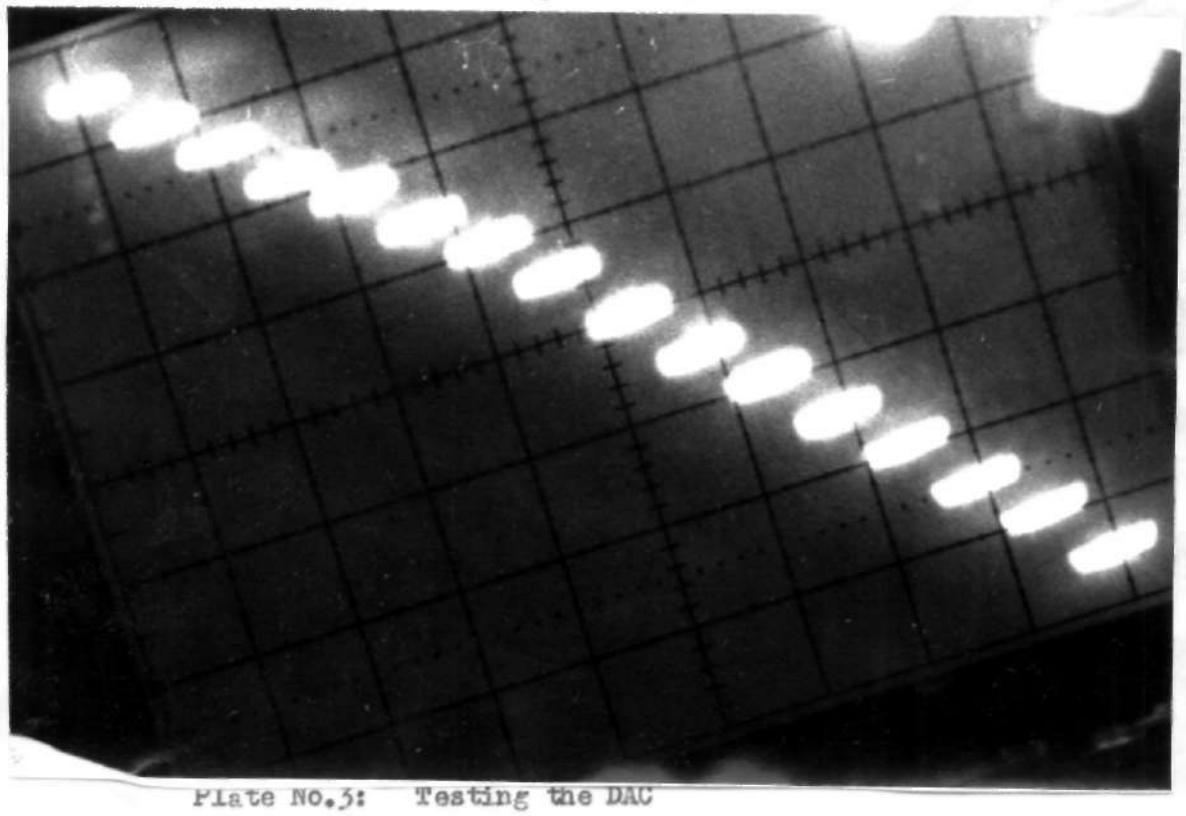


Plate No.3: Testing the DAC