A SYSTEM FOR RECORDING OF BIOLOGICAL SIGNALS

M. Tech. Dissertation

by

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ABSTRACT

The development of a recording system for biological signals, using a commercially available stereo cassette recorder has been the aim of this project. The system consists of a front-end module of a general purpose biosignal amplifier and frequency modulator-demodulator for faithful recording and retrieval of the low frequency biological signals. An attempt has also been made to have a precise compensation of flutter caused by random variation in tape speed.

CONTENTS

CHAPTER-I

INTRODUCTION

1.1 ACQUISITION OF BIOSIGNALS

In research, as well as in clinical situations, a need often arises to monitor and record a number of physiological signals. These signals can be used for diagnosis and also to understand physiological mechanisms. Physiological signals include bioelectric signals like Electrocardiogram (ECG), Electroencephalogram (EEG), Electro-occulogram (EOG), Electromyogram (EMG), etc as well as signals like lung pressure, blood flow, muscle force, etc. Research towards understanding of physiological mechanisms for years has made it possible to correlate these signals with the functioning of the various physiological systems. To extract diagnostic information from these signals, one needs to have a faithful representation of these signals.

Biosignals are often low in magnitude and have frequency range extending down to dc. Table 1.1 gives the frequency range and voltage levels for some clinically important signals: ECG, EEG, EMG, and EOG.

Table 1.1 Frequency range and voltage level for some of the biosignals [1].

Signal	Frequency Range (Hz)	Voltage Range (mV)	Voltage Typical Value
ECG	0.1 - 100	0.5 - 5	1 mV
EEG	0 - 100	0.01 - 0.03	50 μV
EMG	10 - 1000	0.02 - 5	50 μV
EOG	0.1 - 100	0.01 - 3	varies with eye position

Further, there are various other recordings like heart chamber pressure, blood flow, blood pressure, cardiac output, lung pressure, respiration rate, etc. Such parameters are estimated by various techniques like ultrasound reflections, impedance plethysmography, sensing by transducers like piezo-electric elements, strain gauges, etc. Almost all of them are converted into electric signals which happen to fall in the frequency range of dc to about a few hundred cycles per second.

Some of the major problems that one often encounters during acquisition of biological signals include offset and drift [2], noise, and artifact [3].

1.2 RECORDING AND RETRIEVAL OF BIOSIGNALS

In case of signals which are not recurring or aperiodic, the interpretation of the signals is greatly improved through further sophisticated signal processing techniques. This requires digitizing such signals. To derive diagnostic information from signals like EEG, EMG considerably long duration records are necessary.

For example, the EMG can be used to detect localized muscular fatigue. It has been shown that the median frequency of the EMG shifts to a lower value on fatigue. Such a shift also corresponds to reduction in muscular conduction velocity. To detect muscular disorders by this technique one needs to have long duration record of EMG [4]. Similarly, monitoring of brain and neural activities also requires a long duration record to extract the information of the organic brain disorders [2].

Signals like ECG and EEG are recorded in various lead configurations. To correlate various activities all the leads must be recorded simultaneously. In kinesiological studies, for example, EMG is also needed to be recorded on the same time scale along with the output of electrogoniometer which gives electric output of the position of certain body segment [5].

1.3 AIM OF THE PROJECT

Use of paper chart recorder for permanent record of biological signals is not always appropriate since for signal processing one needs subsequent digitization of signal which is very tedious once we have signal recorded on paper. Instead, it is more useful to have a magnetic tape record which makes it economical and convenient to playback the signal as often as required. It is a low cost alternative, especially when only a small part of a long record that is of most interest is to be further processed by other sophisticated instruments like a signal analyzer.

This project aims at developing a low cost recording system for biological signals. The work has two major components : (i) To build a low cost multi-channel FM recorder with precise flutter compensation, based on commercial stereo cassette player. (ii) To develop a front end module of general purpose instrumentation amplifier.

In commercial cassettes the tape width is used for recording four tracks. When each side of the tape is played, only one half of the tape width is used with the two channels providing stereo recording. However, commercial record-playback heads are also available with four channel facility (these heads with all four channels are used in auto-reverse type cassette recorders). These could be used to extend the facility to four channel recording, as shown in Fig. 1.1 [6].

Here it must be mentioned that a number of instrumentation tape recorders are commercially available. The objective of this project is to develop a low cost hardware that can be used to faithfully record and retrieve low frequency biosignals using portable cassette recorder.

Fig. 1.2 shows the general block diagram of the system. Electrodes or transducers are the sensing elements of the system. The amplifier is to bring the signal picked by them to the level necessary for recording. The low frequency biosignals are then frequency modulated by modulator circuit and stored on the tape recorder. On playback of the tape the retrieved signal is to be frequency demodulated in order to get back the original signal. This is done by a frequency demodulator.

1.4 OUTLINE OF THE REPORT

In Chapter 2 the characteristics of the biosignal amplifier will be discussed. Chapter 3 gives a brief account of the methods of recording on a magnetic tape. Scheme to implement frequency modulation and flutter compensation with multi-track recording is also discussed.

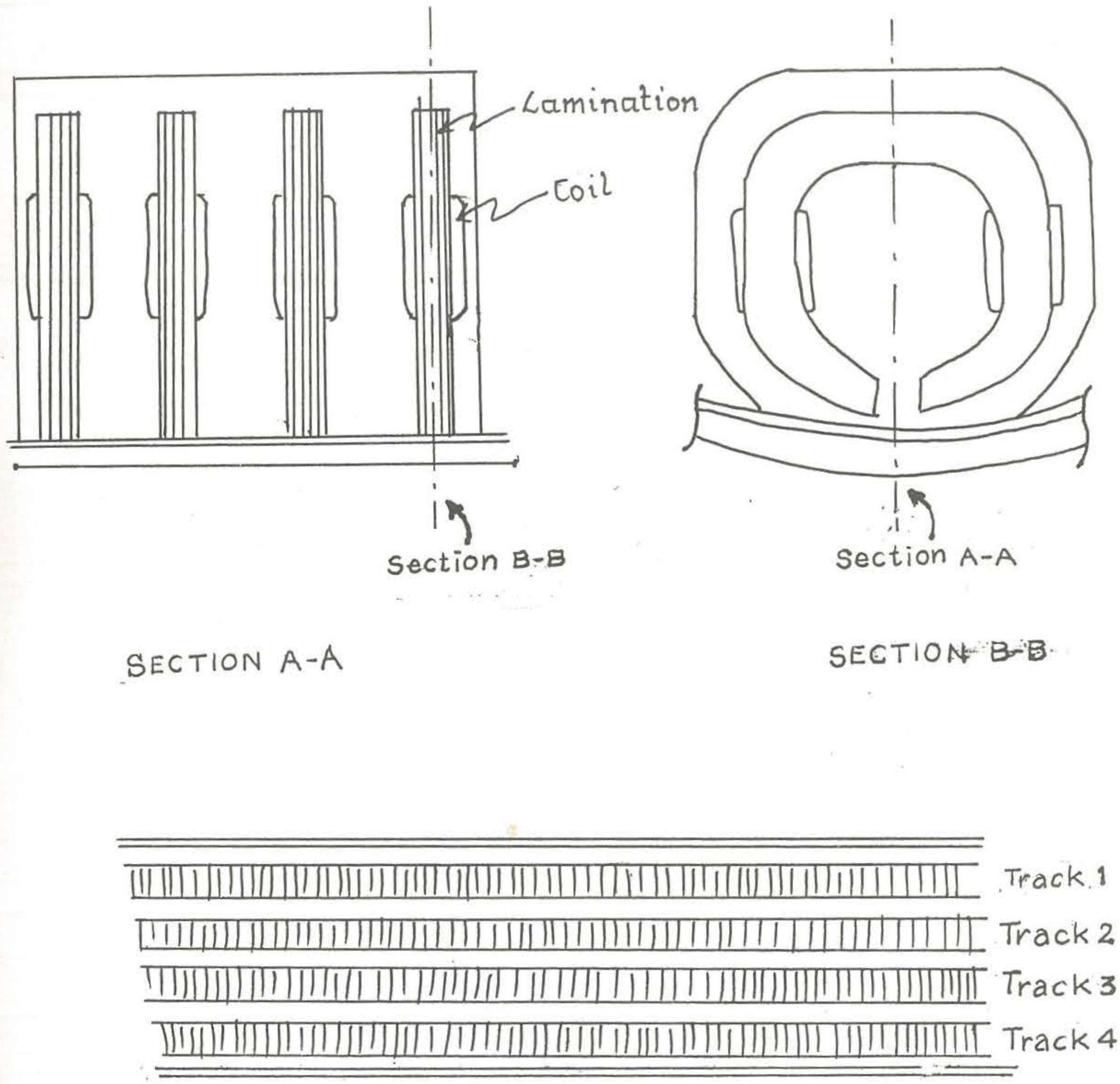
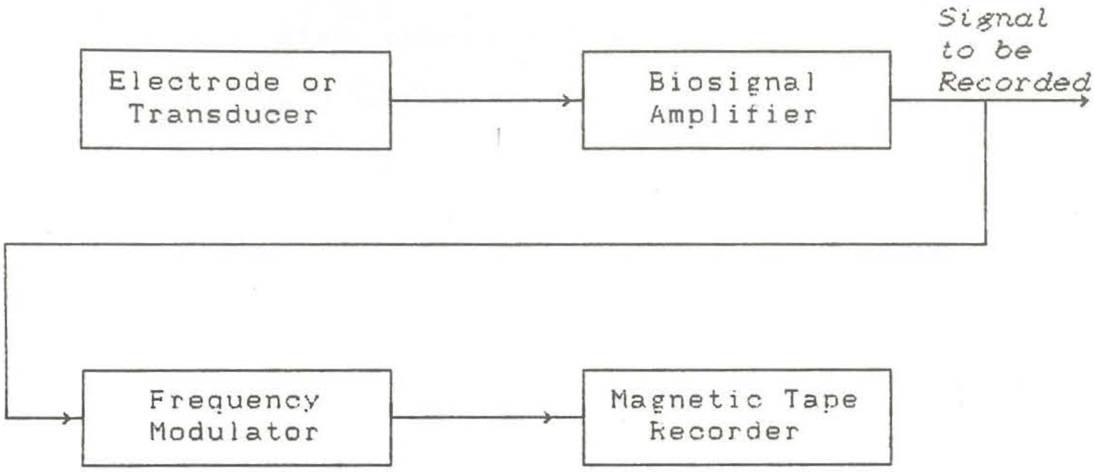
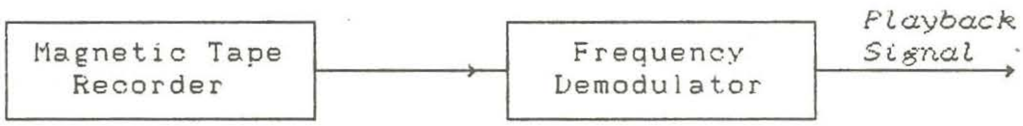


Fig 1.1 Head assembly for four-channel recording [6].



Recording Mode



Playback Mode

Fig 1.2 General block diagram of the system.

Chapter 4 explains about the design and testing of the hardware and also presents the results obtained. The last chapter of the report comprises of the conclusion and suggestions for further work.

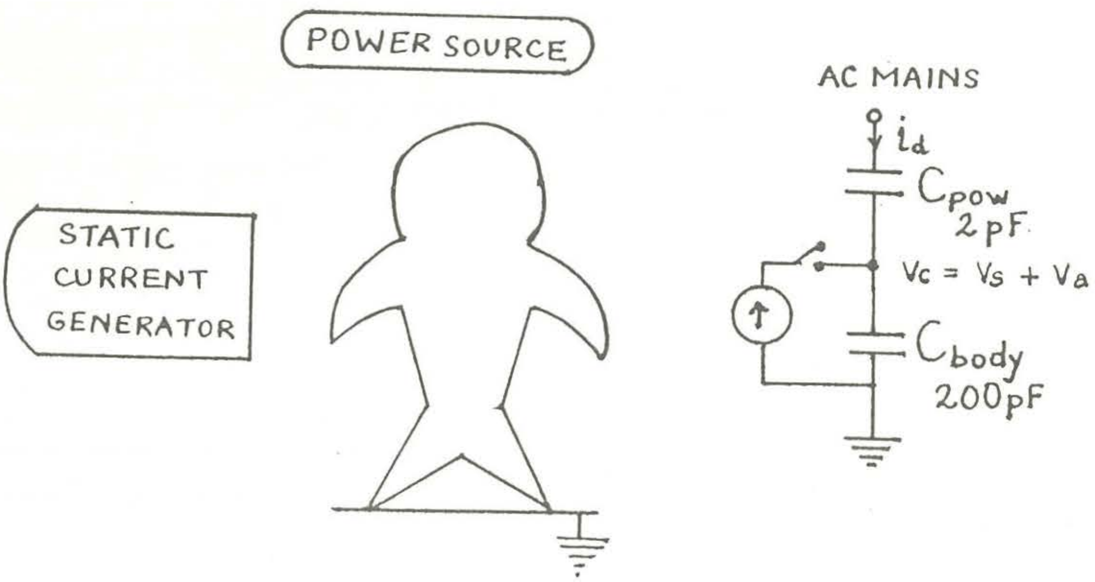
CHAPTER - II

BIOSIGNAL AMPLIFIER

2.1 INTRODUCTION

The general characteristics of the biosignal amplifier include high input impedance, high common mode rejection ratio (CMRR), and adjustable gain. High input impedance requirement arises due to a very high source impedance of biological signals. For example, for ECG recording, the electrode skin interface impedance is typically of the order of 100 KΩ for dry skin and 10 KΩ for a prepared skin [7]. The micro-electrodes of area of about 400 μm^2 that are used for applications like recording single-unit intracortical activity, manufactured by thin film technology offer impedance in the range of a 2 to 8 MΩ [8]. Tungsten microelectrodes having area 8 μm^2 present an impedance in the range of tens of MΩ [7]. Hence the prime requirement of biopotential amplifier is to have high input impedance typically in the range of tens of MΩ for ordinary electrodes and much higher for micro-electrode recordings. Low values of input impedance cause attenuation of the signal to be recorded which are already very weak.

Biosignals are usually recorded differentially with common mode voltage being usually larger than the differential signal. High CMRR is essential to get rid of these common mode voltages. For example, during surface recordings, the common mode voltage V_c is composed of a static voltage component V_s and power line induced ac common mode voltage V_a, caused by displacement current I_d which flows through stray capacitors as shown in the Fig 2.1. V_s is created by patient movements. To reject these voltages, one must have a high CMRR. Practical values of CMRR are in the range 60 - 120 dB [9].



V_s : Static component
V_a : Powerline ac component

Fig 2.1 Components of common mode voltage on body [9]

The typical frequency ranges of biosignals are given earlier in Table 1.1. The upper frequency cutoff of amplifier should just include the maximum frequency of interest in the signal so as to eliminate higher frequency noise. The lower cut off of amplifier is also important to avoid offsets and drifts. But as the signal frequencies extend down to dc. other methods for compensation of offsets and drifts may be required.

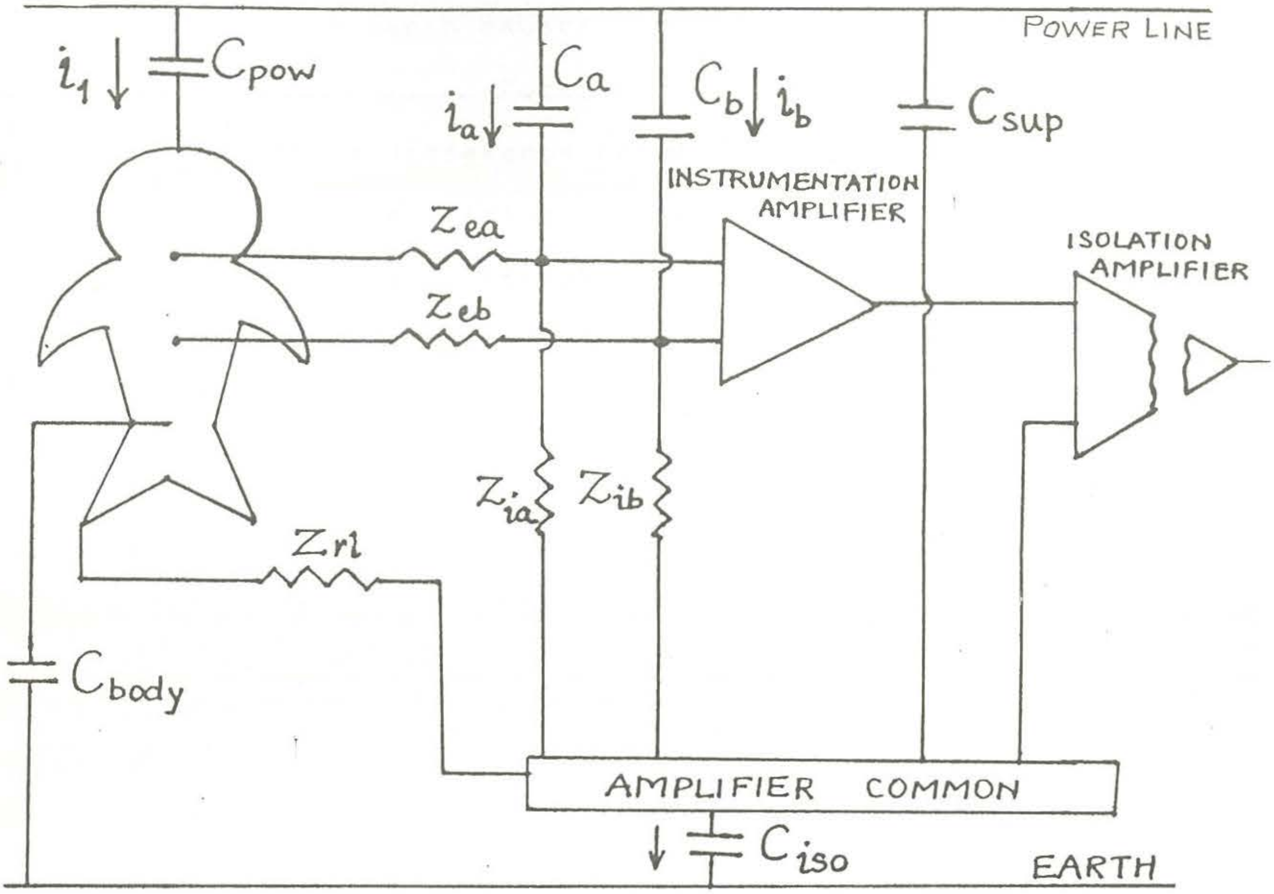
A general purpose biosignal amplifier should also have easily adjustable gain so as to amplify signals in all ranges mentioned in Table 1.1.

Here we will discuss some of the interference sources in biosignal recording and biosignal amplifier configuration.

2.2 INTERFERENCE SOURCES IN BIOSIGNAL AMPLIFIER

Biosignal recordings are almost always disturbed by an excessive level of interference. It is desirable that any disturbance should not exceed 1 % of the peak value of the signal. Fig 2.2 shows the different interference currents encountered in typical ECG recording which include interference currents through the body. that into amplifiers. and measurement cables. etc [10].

The capacitances between the patient and power line



C_{pow} = Coupling Capacitance between mains and body
C_{body} = Coupling capacitance between body and earth.
C_{iso} = Capacitance between amplifier common and earth
C_{sup} = Capacitance between supply and amplifier common
C_a, C_b = Coupling capacitors at opamp input terminals.
Z_{ea}, Z_{eb} = Electrode impedances
Z_{rl} = Impedance of reference electrode
Z_{ia}, Z_{ib} = Input impedances as seen from input terminals

Fig 2.2 Interference currents encountered in a typical ECG recording configuration [10].

and patient and earth causes a small interference current i_1 to flow through body. Portion of i_1 that flows through Z_{rl} causes potential difference between the average potential of the body and the amplifier common.

For isolated amplifiers. interference currents also flow between amplifier common node and mains as well as amplifier common node and earth through capacitors C_{sup} and C_{iso} as shown in Fig 2.2 [10].

Capacitive couplings C_a and C_b between cables with mains also induce interference currents I_a and I_b. They flow through the body via electrodes and from the body to earth via C_{body} and Z_{rl}. Since both the currents induced in both wires differ and the values of electrode impedances also differ from each other. a differential voltage may get presented to the amplifier.

Further. there could be magnetic interference which can be distinguished from other interferences because it varies with the area and orientation of the loop formed by cables.

2.3 BIOSIGNAL AMPLIFIER CONFIGURATION

A biosignal amplifier is essentially an instrumentation amplifier as shown in Fig 2.3 [11]. It is basically a differential amplifier with two voltage

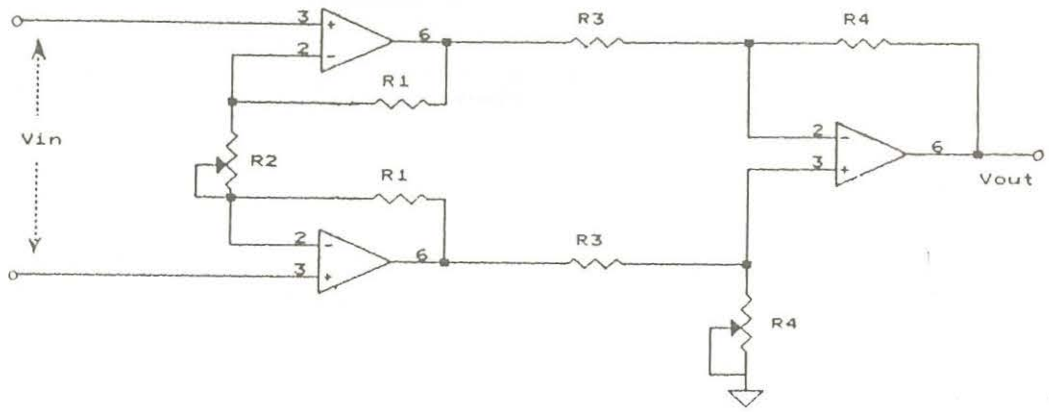


Fig 2.3 Instrumentation amplifier

followers at input stages. Use of special purpose operational amplifiers in this circuit configuration results in the following features that make it suitable for many bioengineering applications:

(i) Ultra-high input impedance: The input stages are generally built with JFET input opamps operating as voltage followers as shown in the figure. With such design the input impedance can be made as high as 100 GΩ.

(ii) High and variable gains: Gains typically range from 10 to 10,000. It can be easily set with an external potentiometer R_4.

(iii) Low input bias current : The input bias current to the amplifier is generally provided through the electrodes. This poses a requirement of the very low input bias current.

(iv) High CMRR : To achieve the best CMRR resistance R_2 can be made variable to compensate for the mismatch in the resistance ratio at inverting and non inverting inputs.

There are various modifications that are reported in literature in this basic circuit of instrumentation amplifier in order to improve its performance like reduction of common mode voltage by right leg circuit design [12]. two electrode bootstrapped common amplifier [9]. low power. low noise instrumentation amplifier [13]. etc.

CHAPTER - III

MAGNETIC TAPE RECORDING

3.1 INTRODUCTION

Instrumentation tape recorders are widely used by medical researchers, clinicians, and teachers who are interested in multi-channel collection, storage, and analysis of the physiological data. Instrumentation tape recorders are excellent means of storage and transfer of vast data from one tape to the other at very low cost. The need to transfer tapes and signals from recording equipment to different replay and analysis equipment in laboratory has created the need to develop the standards for tape format and speed. Inter Range Instrumentation Group (IRIG) has established standards for tape speed, reel sizes, and tape width for amplitude modulation (AM), frequency modulation (FM) as well as pulse duration modulation (PDM) [14,15]. There are commercially available tape recorders which offer one or more of FM, AM, PDM mode for recording analog signals. Analog signals after digitizing can be recorded on the tape by digital pulse recording technique. Here is a brief review of analog recording methods and we will see how FM proves to be advantageous.

3.2 TECHNIQUES USED IN DATA RECORDING

In recording of signal, for the purpose of later reproduction and analysis, it is desirable that overall degradation of the actual information in the signal due to the recording system should not be significant. Frequency selective characteristics of magnetic recording system, pose a constraint on frequency range which can be satisfactorily recovered, and makes it impossible to directly record and reproduce dc levels and very low frequency signals [16] which are very important especially in biological applications. Hence to recover such low frequency signals, one of the following carrier modulation techniques are used:

1. Amplitude modulation (AM)
2. Frequency modulation (FM)
3. Pulse duration modulation (PDM)

3.2.1. Amplitude Modulation (AM)

When a continuous frequency carrier wave is amplitude modulated, its amplitude is varied in a manner determined by signal voltage that is to be recorded. Thus, the variation imposed on the carrier amplitude contains the information of the signal to be recorded. The process of varying amplitude of a continuous carrier wave causes appearance of two frequencies in addition to the carrier frequency. If $e_c = A \cos 2\pi f_c t$ is a sinusoidal carrier wave and another sinusoidal waveform $e_s = \cos 2\pi f_s t$ has to modulate it, then 'A' varies as per the instantaneous value of the modulating waveform to be recorded.

$$A = A_o (1 + m \cos 2\pi f_s t)$$

where, m = modulation factor so that, the modulated waveform is given by,

$$\begin{aligned} e &= A_o (1 + m \cos 2\pi f_s t) \cos 2\pi f_c t \\ &= A_o \cos 2\pi f_c t + m A_o \cos 2\pi f_c t * \cos 2\pi f_c t \\ &= A_o \cos 2\pi f_c t + \\ &\quad m(A_o/2) [\cos(2\pi f_c + 2\pi f_s)t + \cos(2\pi f_c - 2\pi f_s)t] \end{aligned}$$

Thus, amplitude modulated signal contains a lower side band and an upper side band as shown in Fig 3.1 [16].

The disadvantage of AM is that with aging, the magnetization on the tape deteriorates non-uniformly and this results in non-uniform reduction in signal strength when replayed. These fluctuations in the amplitude of the carrier will result in corruption of the demodulated output.

3.3.2. Frequency Modulation (FM)

In this technique, the carrier amplitude is maintained constant and frequency is shifted back and forth about a central position as per the magnitude of the signal

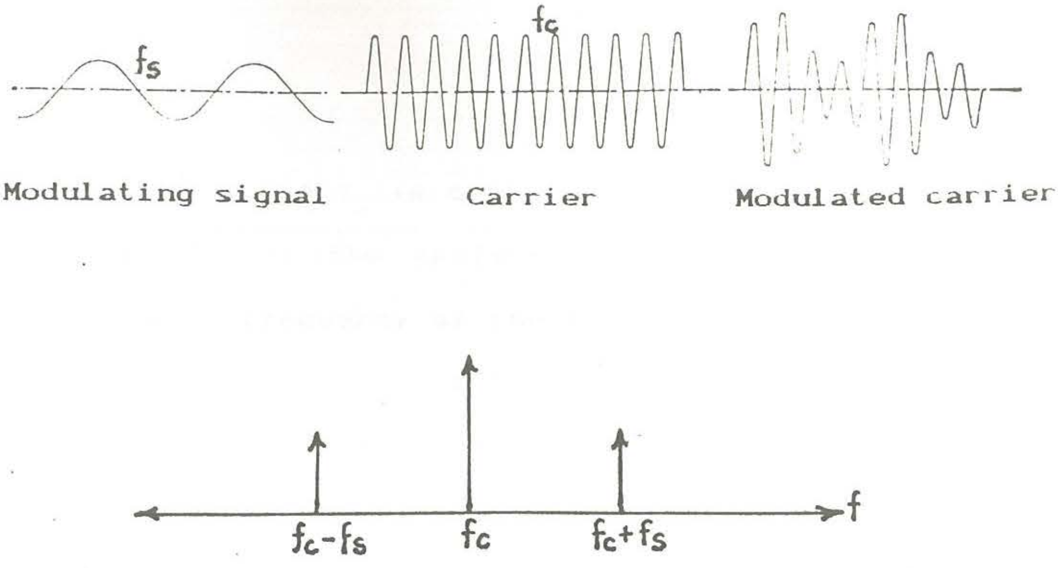


Fig 3.1 Amplitude modulation [16].

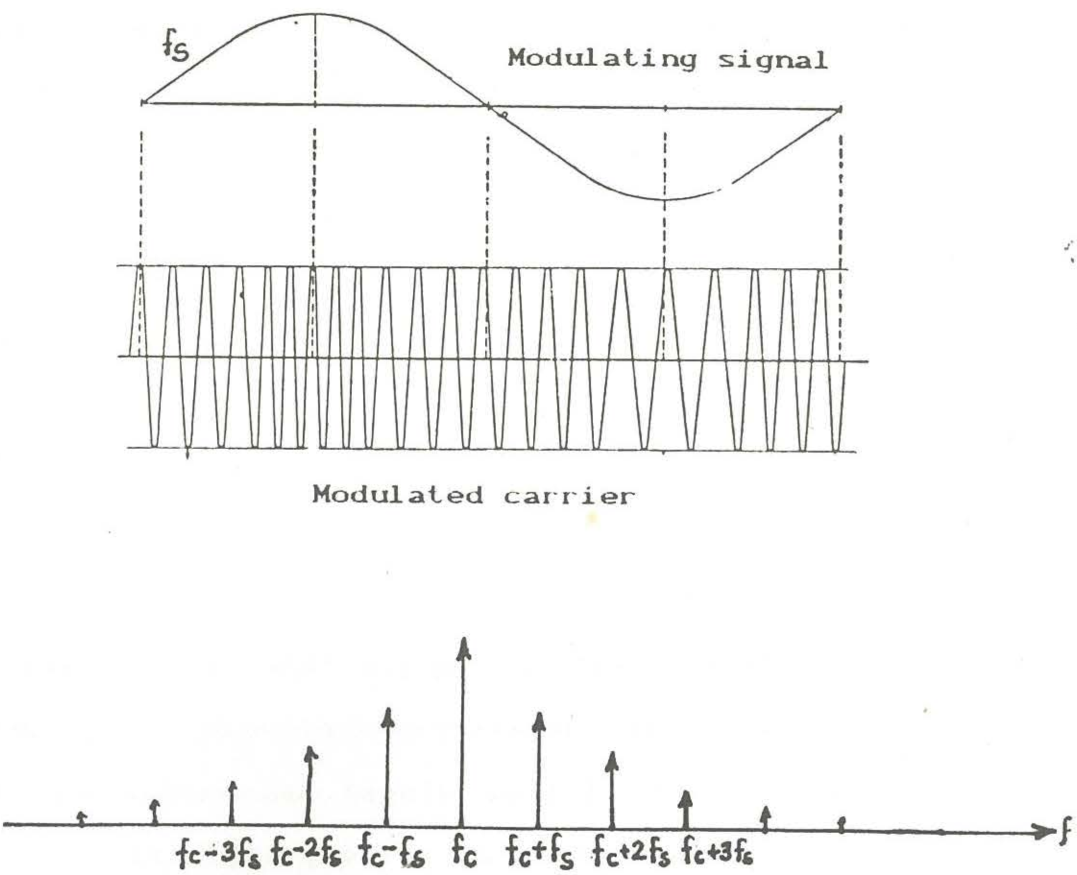


Fig 3.2 Frequency modulation [16].

to be recorded. If f_c is carrier center frequency, and f_s is the frequency of the maximum modulating signal, then the instantaneous frequency of the modulated waveform is,

$$f = f_c (1 + m \cos 2\pi f_s t)$$

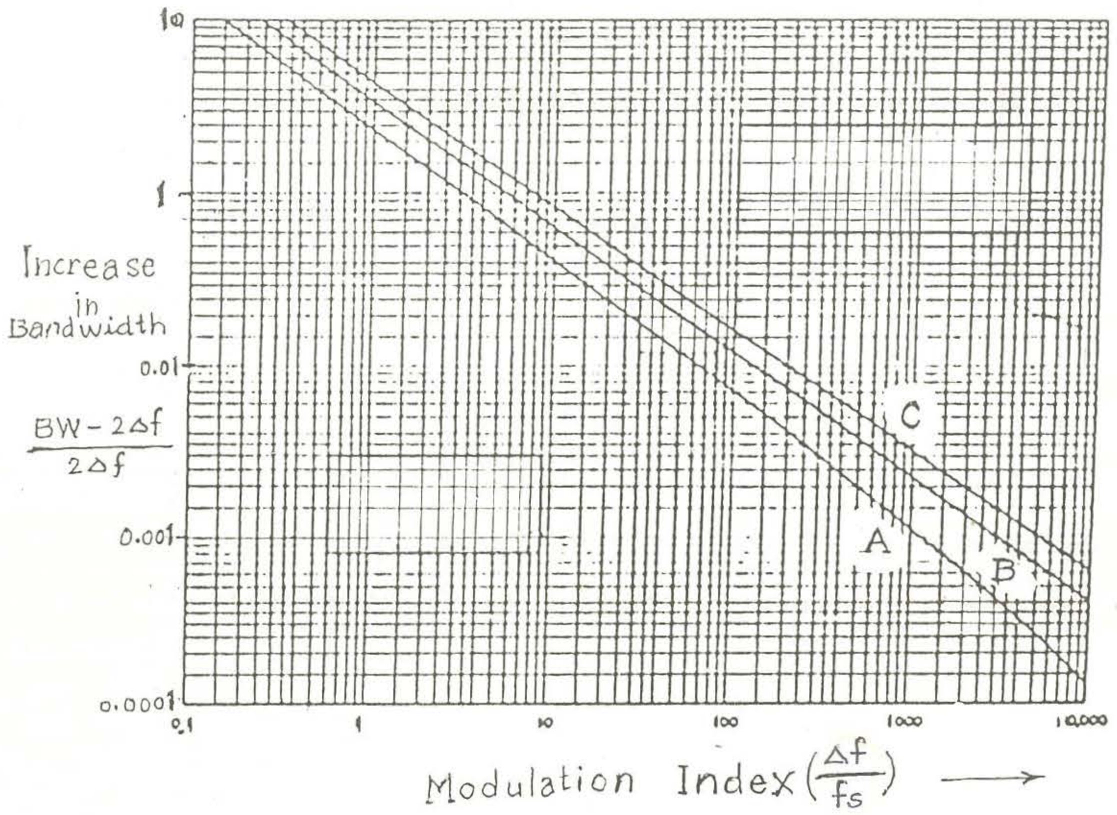
The modulated waveform is given as $e(t) = A \cos \phi$

where, $\phi = \int 2\pi f \, dt$

$$= \int (2\pi f_c + m \, 2\pi f_c \cos 2\pi f_s t) \, dt$$

$$= 2\pi f_c t + m(f_c / f_s) \sin 2\pi f_s t$$

The mathematical analysis using Bessel function shows the presence of different side bands apart from carrier frequency as shown in Fig 3.2. The side bands also exceed maximum and minimum points reached by the carrier, theoretically up to infinity. But, we can find a certain finite number of side bands generated which have sufficient power of value for the reconstruction of the signal. Modulation index is the quantity that decides the requirement of the bandwidth. It is defined as the ratio of $\Delta f / f_s$, where, Δf is the extreme swing of the carrier from the center and f_s is the maximum signal frequency. Lower the modulation index, more are the number of sidebands required for faithful reconstruction of the signal which in turn increases the bandwidth requirement of the system. Bandwidth that the system can handle should cover all the sidebands that have strength greater than 1% of the carrier amplitude obtained when signal is removed. Fig 3.3 shows the variation



Curves 'A', 'B' and 'C' give the bandwidth required to retain components upto 1%, 0.1%, and 0.01% strength of the carrier respectively.

Fig 3.3 Variation of bandwidth with modulation index of required bandwidth with modulation index for retaining significant components of carrier having strength above 1%, 0.1%, 0.01% [17]. We see that, the required bandwidth for retaining all frequency components with a certain significant strength increases as the modulation index decreases.

To obtain the bandwidth required to retain components above 1%, we refer to curve 'A' in Fig 3.3. Table 3.1 gives a simplified relation between signal bandwidth f_s and the required bandwidth BW as a fraction of the total frequency deviation $2\Delta f$ calculated from curve A. For example, if the maximum modulation frequency f_s is 500 Hz and the total swing of the carrier $2\Delta f$ is 5 kHz, then from the table, for the corresponding value of $f_s/2\Delta f$, the value of the required bandwidth as a fraction of total swing $2\Delta f$ is 1.7. Hence, in this particular case, the required bandwidth is 8.5 kHz. Thus, knowing the signal bandwidth and the swing of the carrier frequency, the requirement of the system bandwidth can be estimated.

Tape speed plays an important role in the selection of the carrier frequency and also decides signal bandwidth. Table 3.2 gives the carrier and maximum modulation frequencies at some common tape speeds reported in the operation manual of eight channel FM instrumentation tape recorder model 8868A by Hewlett Packard [18].

Table 3.1 Signal bandwidth f_s Vs required bandwidth BW as a fraction of total carrier swing $2\Delta f$

$f_s/2\Delta f$	$BW/2\Delta f$
0.05	1.4
0.1	1.7
0.2	2.3
0.5	3.5
1.0	5.0

Table 3.2 Carrier and Modulation Frequencies for common tape speeds [18].

Tape Speed (ips)	Carrier Frequency (kHz)	Passband (kHz)	Signal-to-noise ratio (dB)
15.0	27.0	dc - 5.0	46
7.5	13.5	dc - 2.5	46
3.75	6.75	dc - 1.25	44
1.875	3.38	dc - 0.625	40
0.938	1.69	dc - 0.312	40
0.469	0.85	dc - 0.156	40

FM technique, is immune to amplitude instability as against AM, but is troubled greatly by irregularities in the tape speed, which are equivalent to fluctuations in frequency [19]. These problems are discussed in the later sections.

3.2.3 Pulse Duration Modulation (PDM)

In this technique, a square wave is the basic carrier, when there is no modulating signal input. Modulating signal causes change in the duty cycle of the carrier wave as shown in Fig 3.4.

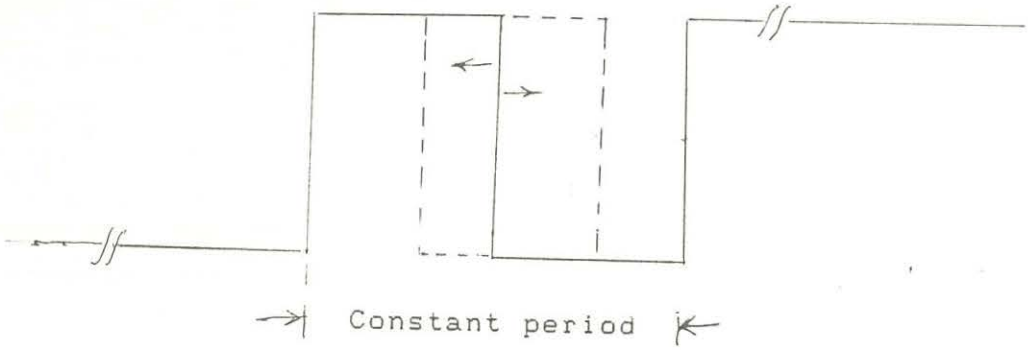


Fig 3.4 Pulse duration modulation

Compared to FM, PDM requires a larger system bandwidth. The bandwidth requirement is indicative from the frequency spectrum of PDM modulated signal. The basic square wave, with zero modulation input is comprised of a fundamental and a number of odd harmonics. Even harmonics appear in the frequency spectrum when the square wave is not symmetrical. Modulation of duty cycle produces some changes in the amplitude of fundamental and the odd harmonics. However, the major changes occur in the even harmonics. In other words, major information is stored in the even harmonics of the carrier frequency. Hence the reasonable number of even harmonics (about 5) must be preserved to accurately recover the crossover time of the carrier.

For example, consider a special case of wideband FM with 40% modulation ($\Delta f/f_c$) and modulation index ($\Delta f/f_s$) of 5, all the side bands after 8^{th} sideband are having strength less than 1% as shown by Bessel function theory [15]. Thus, the total bandwidth required is $16f_s$ or 1.3 times the center frequency. On the other hand, for PDM, in order to retain 5 even harmonics of the carrier, one needs to have a bandwidth of 10 times the center frequency. Thus, for the same ratio between the carrier and the maximum modulation frequency, FM system has almost 7 times usable frequency bandwidth of the PDM system at a given tape speed [15].

3.3 HARDWARE TO IMPLEMENT FM

A scheme for a single channel FM recording is shown in Fig 3.5. The basic idea is to vary the carrier frequency

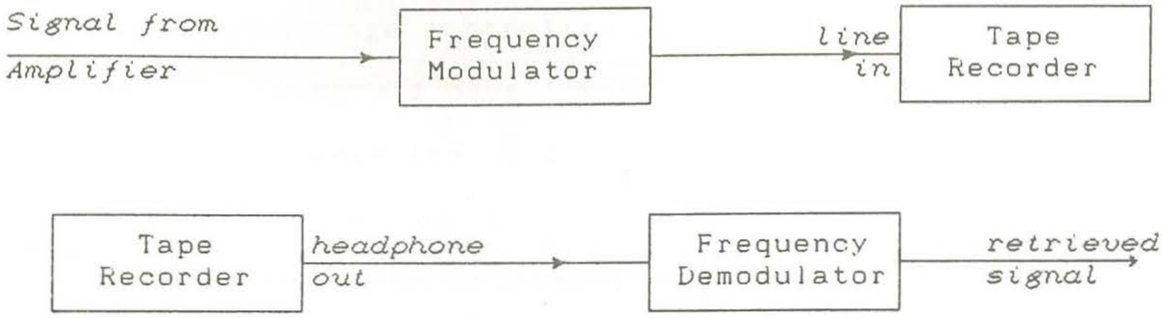
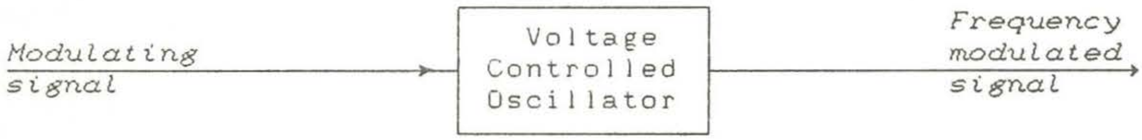


Fig 3.5 Single channel FM technique for FM tape recording



(a)

(b)

Fig 3.6 Schemes for modulation and demodulation

as per the amplitude of the signal to be recorded. This can be done by a voltage controlled oscillator as shown in Fig 3.6(a). The VCO should have linear operation.

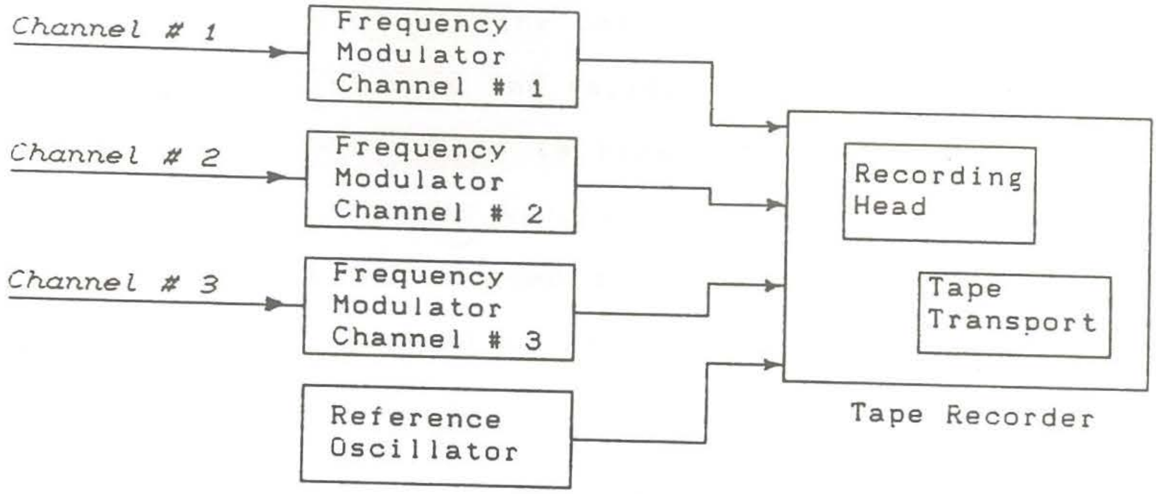
To retrieve the original signal after playback, the signal played back has to be passed through a frequency demodulator. There are a number of frequency demodulation schemes, such as those involving frequency discriminator (FM to AM followed by AM detector), phase locked loops, etc [20]. But these are generally suitable for low frequency deviations and high carrier frequencies. Since this application requires relatively small carrier frequencies but wideband FM, pulse averaging technique [11] is more appropriate.

The block diagram of a scheme to implement demodulation by this technique is shown in Fig 3.6(b). The modulated signal is fed to a monostable generating fixed duration pulses, the frequency of occurrence of the pulses being equal to the input frequency. The average value of the signal at the output of the monostable will be thus proportional to the input frequency. Hence one can recover the original modulating signal by passing this train of pulses through an averaging lowpass filter. The pulse width of the monostable should be smaller than minimum time period of the carrier frequency. Also the lowpass filter should be designed carefully to minimize the ripple to an acceptable level, and to still maintain all the frequency components in the frequency range of interest.

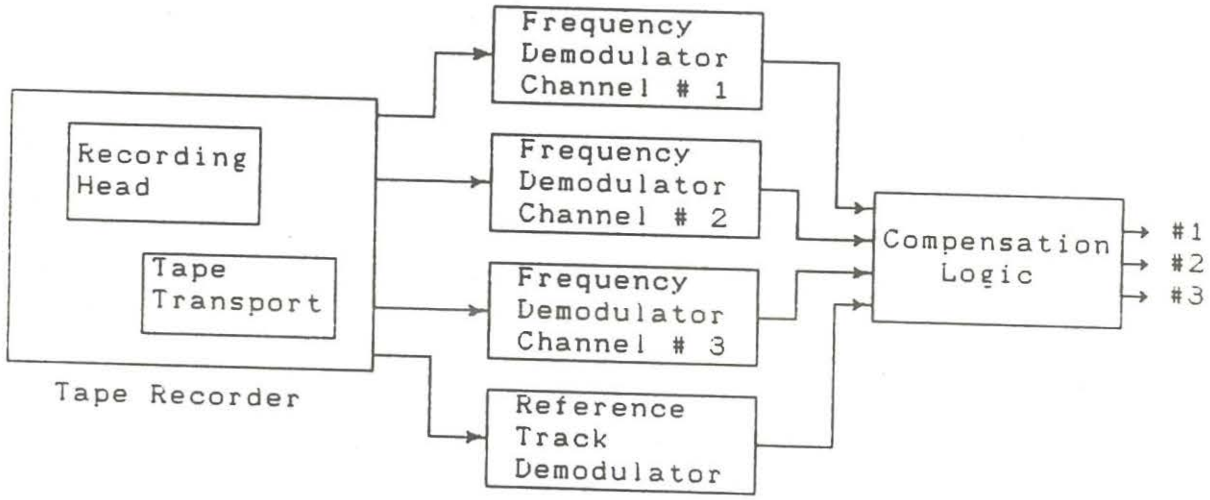
3.4 MULTI-CHANNEL RECORDING AND FLUTTER COMPENSATION

The fact that magnetization in a magnetic recording medium of high coercivity is greatly localized, is taken advantage of to produce a magnetic recording cassette consisting of a wide tape in which the number of tracks are laid down side by side relying simply on a slight separation between the tracks to avoid the cross-talk. On the commercially available cassette, four track record is possible using the cassette only in one direction. In this particular scheme as shown in Fig 3.7 three tracks can be used as recording tracks and the fourth one as a reference track.

Ideally, tape transport mechanism should move the tape across the head at a precisely known uniform velocity. But there are bound to be variations in the velocity of the tape as it moves across the head. The gross departures from the desired average velocity may be corrected by servo speed control during the playback. However the short term velocity variations can not be eliminated by such method. Short term velocity variations are of two kinds: those which are uniform across the tape so that the velocity at the given



Recording



Playback

Fig 3.7 **Block schematic of four channel recording and playback with flutter compensation.**

instance is same at all the points in the line across the tape and those in which the variations are not same across the tape. The former type is known as flutter while later are mentioned as random velocity variations [16]. Flutter arises from numerous sources in the tape drive mechanism, viz. loose belt, eccentricities in capstan, jamming of the rotary parts, excessive roller pressure, roughness of the tape, tension variation in the tape, etc [19]. Despite the advances in the design and assembly of tape mechanism, certain amount of flutter can still be anticipated.

One of the methods to reduce the effect of flutter has been suggested by Hirschberg [19] in which, input signal modulates the frequency of occurrence of the constant width pulses. The impulses corresponding to the leading and trailing edges of the pulses are recorded on tape, which in turn are used on playback to re-establish the original pulse with a timing that is governed by the tape speed itself. These pulses, then are passed through the averaging filter to recover the recorded signal. The average value of the pulses is given by,

$$V_{avg} = A\ f\ T_w \qquad (3.1)$$

where, A is the pulse height, f is the instantaneous pulse frequency, and T_w is the pulse width. On playback, if there is an increase in the tape speed, the pulse width (T_w) reduces and thus compensates for the increase in frequency (f). This method does not require a reference track. However, the available signal bandwidth for recording is small.

Another method that is commonly used in FM technique is to use one of the tracks to record a known dc level and to use the played back waveform from this track to compensate for the effect of flutter on the other track. This method provides a higher maximum frequency of the modulating signal and hence will be used for this system. This method is mathematically treated in the next paragraph.

The flutter causes small variation in the frequencies of the recorded waveforms as they are played back. In the system with FM recording technique, if 'f' is the frequency at the output of VCO, due to flutter, it will get recorded as $f(1+\delta_r)$ where, δ_r is the fractional (positive or negative) error in the frequency recorded on the tape. Similarly, due to the flutter during the playback, a further error occurs making the original recording frequency f to get played back as $f(1+\delta_r)(1+\delta_p)$ or $f(1+\delta)$.

Let x(t) be the signal to be recorded, i.e., the modulating signal. Then, the signal recorded after modulation is,

$$a = \cos\,[2\pi(f_c + k_1 x)t] \qquad (3.2)$$

where, f_c is the center frequency of carrier, k_1 is the frequency-to-voltage sensitivity of the VCO.

Due to flutter of the tape on playback, the signal at the output of the tape recorder is,

$$a' = \cos\,[2\pi(f_c + k_1 x)\,(1+\delta)\,t] \tag{3.3}$$

On demodulation of this waveform, one gets,

$$\begin{aligned} x' &= k_2\,[(f_c + k_1 x)\,(1+\delta) - f_c] \\ &= k_1 k_2 (1+\delta) x + k_2 f_c \delta \end{aligned} \tag{3.4}$$

where, k_2 is the voltage-to-frequency sensitivity of the demodulator circuit.

Since δ varies randomly with time, we see that flutter results in two types of errors:(a) random fluctuations in the amplitude of the retrieved waveform, and (b) random additive effect.

To compensate for this flutter, let x_o be a known dc level recorded on the reference track. By proceeding on similar lines, we get the retrieved reference waveform as,

$$x'_o = k_1 k_2 (1+\delta) x_o + k_2 f_c \delta \tag{3.5}$$

If the dc voltage recorded is 0 V, Eq 3.5 reduces to

$$x'_o = k_2 f_c \delta \tag{3.6}$$

On subtracting the reference waveform (Eq 3.5) from the recorded signal (Eq 3.4) we get,

$$x' - x'_o = k_1 k_2 (1+\delta)\,[x - x_o] \tag{3.7}$$

If $k_1 k_2$ product is adjusted to be equal to 1, and $x_o = 0$, then using Eq 3.7 the retrieved signal is given by,

$$x' - x'_o = (1+\delta)\,x \tag{3.8}$$

Thus, by subtracting the two signals using analog circuits, one can get rid of the random additive error caused by the flutter. If $\delta << 1$, subtraction should give an acceptable flutter compensation. For the precise compensation of the flutter value of δ can be obtained using Eq 3.5, δ being the only unknown quantity and can be used to eliminate the factor of $1+\delta$ from Eq 3.7. The most important premise under the above discussion is that the effect of variation in velocity is identical on both the tape tracks at a given instance of time. Therefore, it should be noted that the random velocity variations that cause independent effects on the two tracks can not be corrected by this method. Further, this analysis holds good only for small fluctuations in the tape transport speed, because the signal itself has been assumed to be constant during this variation. Therefore, this method is not applicable for correcting recorded signal when the playback speed is grossly different from the recording speed. After playback of the waveform, if it has been digitized at a particular sampling rate, errors due to mismatch in tape speed can be corrected by using sampling rate conversion, i.e., digital interpolation and decimation [22].

C H A P T E R - IV

DESIGN AND TESTING OF THE HARDWARE

4.1 INTRODUCTION

This chapter describes hardware development of the circuit. First, the frequency response of some of the commercially available tape recorders was studied. Next step was the development of the hardware circuit for modulator and demodulator. In order to have a display of the recorded data and also for precise flutter compensation, the data from two channels was digitized using a data acquisition card installed in an extension slot of a personal computer. The testing of the hardware was carried out using sinusoidal waves from function generator as well as with an ECG simulator and amplifier.

4.2 FREQUENCY CHARACTERISTICS OF TAPE RECORDER

In order to decide the frequency band of operation, the frequency response of the various tape recorders was studied. Fig 4.1 shows the frequency selective characteristics of some of the tape recorders that were available. The recorders used were the portable models PHILIPS DR628, SANYO MW700K, and a desk top model AIWA

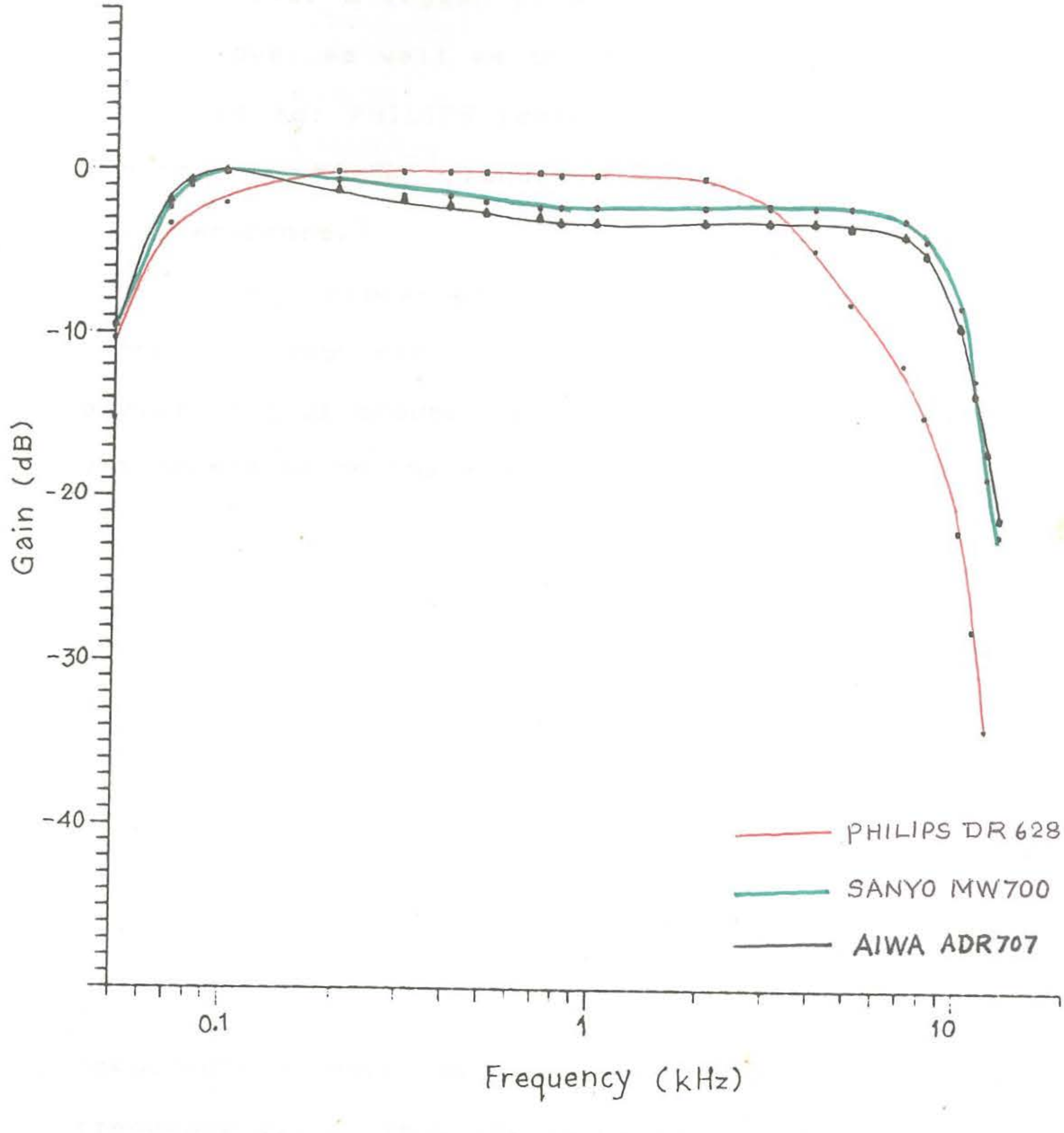


Fig 4.1 Frequency response of different cassette recorders on SONY cassette.

ADR707. The response of each recorder was obtained using the same SONY EF90 cassette. All the recorders show a flat response over a region of a few kHz and the response drops on the lower as well as the higher frequencies. The upper 3 dB cutoff for PHILIPS recorder occurs at 3 kHz whereas it occurs at a higher frequency of about 8 kHz for SANYO and AIWA recorders.

As explained earlier in Section 3.2.2, the bandwidth required for the FM signal with frequency deviation $\pm$ Δf around center frequency f_c can extend beyond $2\Delta f$ depending on the modulation index $\Delta f/f_s$, where f_s is the modulating signal frequency. While selecting the operation band of VCO, one must be sure that on playback. the side bands extending beyond $2\Delta f$ are also recovered. It was decided to use a center carrier frequency of 5.5 kHz, and the signal bandwidth of 0 to 300 Hz. For Δf =2.5 kHz, carrier frequency swing is from 3 to 8 kHz and modulation index is 8.33. By referring to the Fig 3.3 depicting the variation of the bandwidth with modulation index, it can be seen that the required bandwidth for retaining the carrier components of more than 1% strength is 1.5 times the carrier frequency swing. Thus the required bandwidth will be 1.75 to 9.25 kHz which is well within the operating band of the recorders studied.

4.3 FREQUENCY MODULATOR CIRCUIT

Modulator basically consists of a voltage controlled oscillator (VCO). The circuit for frequency modulator is shown in Fig 4.2. The signal conditioner preceding the VCO built by opamp A_{11} accepts the modulating signal of $\pm$ 1 V max, for recording without distortion. Since, VCO built using NE566 accepts only unipolar inputs, signal conditioner gives the dc shift to the input signal as well as scales the input. The relation between the input and output voltage of the signal conditioner is

$$V_{11} = -1.5\ V_s + 5.5 \qquad (4.1)$$

The voltage V_{11} modulates the frequency output of the VCO by following relation,

$$f_{out} = \frac{2.4\ (\ V^{+} - V_{11}\ )}{R_{106} C_{102} (\ V^{+} - V^{-}\ )} \qquad (4.2)$$

where V^{+} and V^{-} are the positive and negative supply voltages to the IC. Thus, the relation between the signal voltage at the input of the modulator and the frequency output in kHz of the modulator is given by following equation,

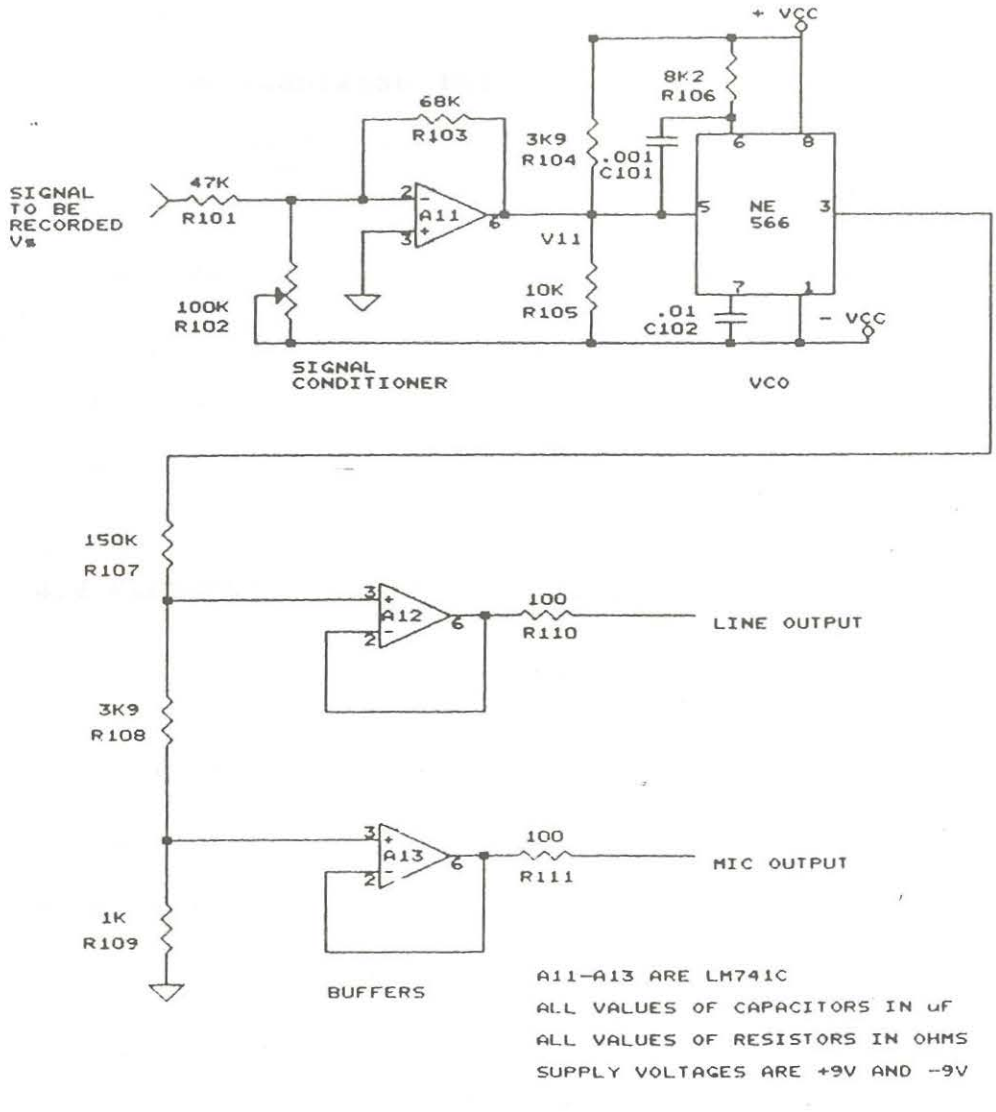


Fig 4.2 Frequency modulator circuit.

$$f_{out} = 2.5\ V_s + 5.5 \qquad (4.3)$$

Thus, the modulator has the sensitivity of 2.5 kHz/V and center frequency of 5.5 kHz.

The output of VCO is attenuated using a voltage divider comprising of R_{107}, R_{108}, and R_{109} to bring it to the level suitable to inputs of the tape recorder. Two buffered outputs of 50 mV and 250 mV are provided to feed "mic" and "line" inputs of the tape recorder respectively.

4.4 FREQUENCY DEMODULATOR CIRCUIT

A number of frequency-to-voltage conversion schemes can be used for demodulation. Since the current application has a relatively low carrier frequency but wideband FM, the pulse averaging technique [11] was selected.

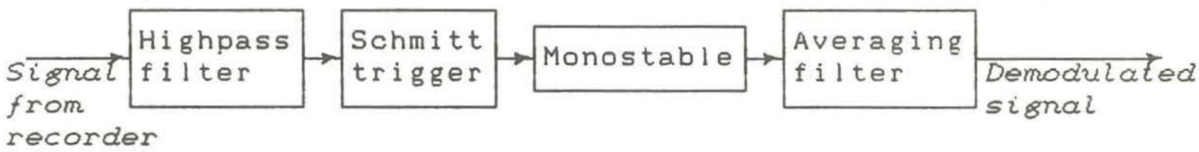


Fig 4.3 Block Schematic of frequency demodulator circuit

Fig 4.3 shows the block diagram of the scheme of demodulation. In order to remove the low frequency component from the playback signal, it is passed through a highpass filter. To regain the modulated square wave from the distorted playback waveform, it is passed through the Schmitt trigger. It is further passed through a monostable multivibrator which gives pulses of constant pulse width 'T_W' and constant amplitude 'A'. These pulses are passed through an averaging lowpass filter. If 'f' is the frequency of the pulse repetition, the average value of the pulses is proportional to the pulse frequency and is given by the equation,

$$V_{avg} = A\ T_W\ f \tag{4.4}$$

The important parameters of the demodulator are order of the filter and pulsewidth of the monostable. The lowpass filter should reduce the ripple due to carrier to a level of -48 dB (assuming an 8 bit ADC application). The carrier frequency swing is from $f_{min} = f_c - \Delta f$ to $f_{max} = f_c + \Delta f$. Thus, if we have a Butterworth filter, its order will be given by,

$$n = \frac{48}{20 \log (f_{min} / f_{clpf})} \tag{4.5}$$

where, f_{clpf} = cutoff frequency of the lowpass filter. For present application, where, f_{min} = 3 kHz and f_{clpf} = 300 Hz to have ripple rejection greater than the specified above, n=4.

Another important consideration while designing the monostable multivibrator is that, pulsewidth of the monostable must satisfy the following relation,

$$T_W < \frac{1}{f_{max}} \tag{4.6}$$

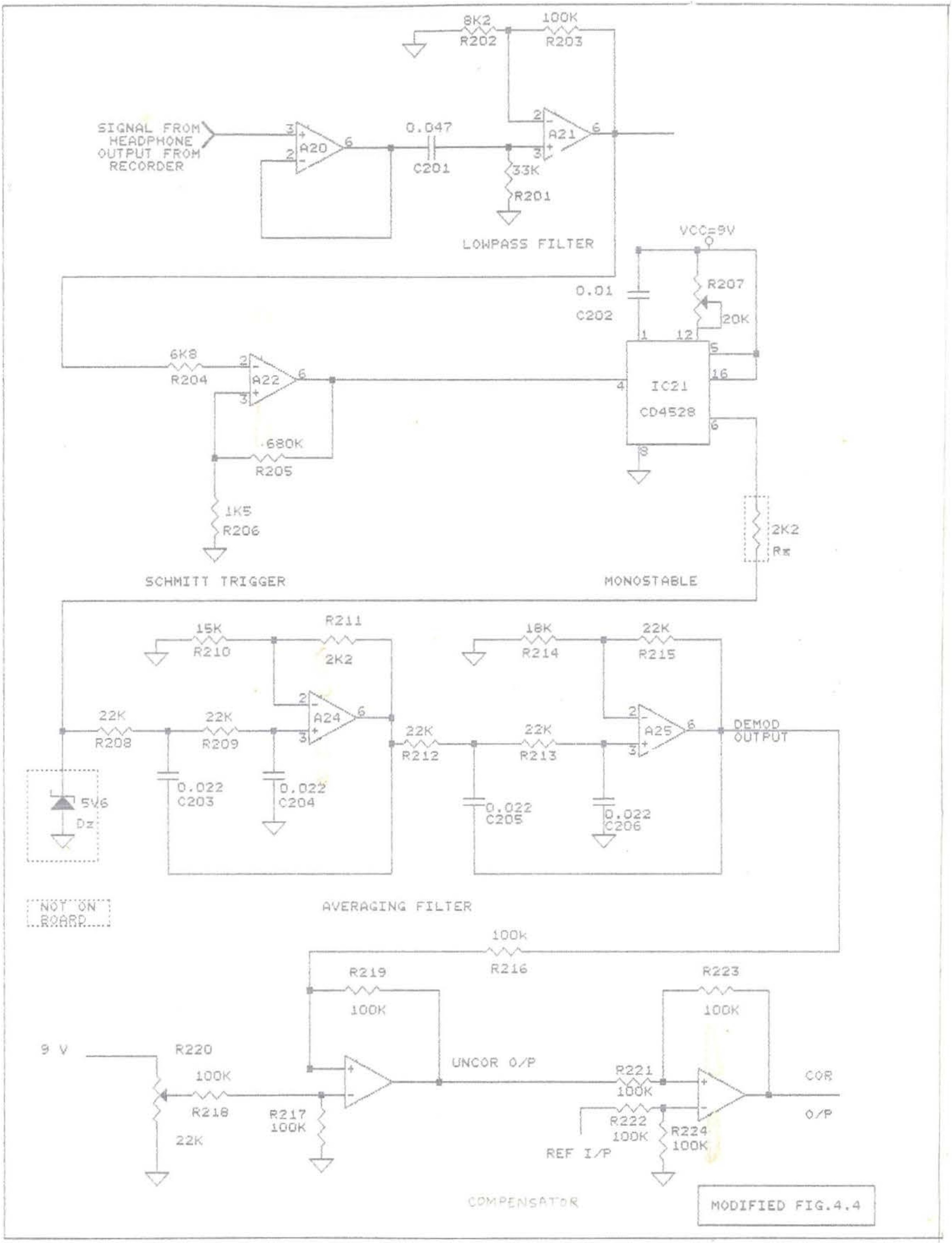
f_{max} = maximum swing of VCO from the center frequency on high frequency side, i.e., $f_c + \Delta f$

Fig 4.4 gives the detailed circuit of the demodulator. An active highpass filter built using opamp A_{21} eliminates the low frequency components in the playback signal from the tape recorder. The cutoff frequency of the filter given by,

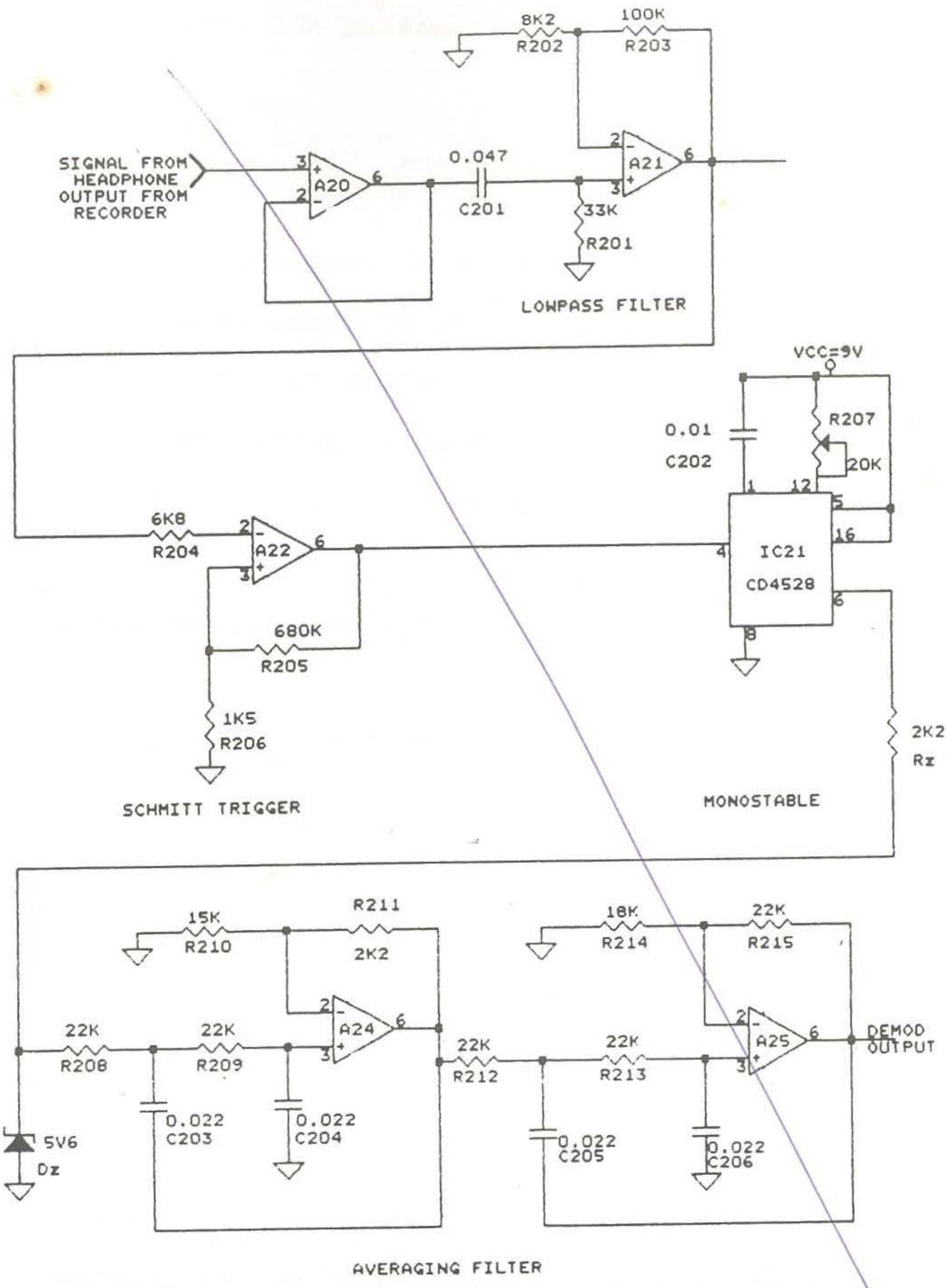
$$f_{chpf} = \frac{1}{2\,\pi\,R_{201}C_{201}} \tag{4.7}$$

is selected to be 100 Hz.

A Schmitt trigger has been realized by using opamp A_{22} and resistors R_{24}, R_{25}, and R_{26} which reshapes the playback signal to a perfect square wave. A CMOS IC CD4528 is used as a monostable to give the pulses of constant



MODIFIED FIG.4.4



A20–A25 ARE LM741C

All values of Capacitors in μF

All values of Resistors in Ohms

SUPPLY VOLTAGES ARE +9V AND -9V

Fig 4.4 Frequency demodulator circuit.

pulsewidth. The pulsewidth is given by the formula,

$$T_{w} = 0.2\ C_{202}\ R_{207}\ \ln\ (V^{+}) \qquad (4.8)$$

where, V^{+} = supply voltage of the IC. Here, the pulsewidth selected is about 70 μs.

To get the dc average value of the pulse train, it is passed through a fourth order highpass active Butterworth filter comprising of opamps A_{23} and A_{24} each forming a second order lowpass filter. The design of the filter is discussed in Appendix A.

4.5 TESTING OF MODULATOR AND DEMODULATOR CIRCUITS

The static transfer characteristics of the modulator and demodulator are shown in Fig 4.5. Both these characteristics show the linear operation of the circuits over the operating range of the inputs. The graph is drawn with a common frequency axis for the modulator and demodulator, input voltage to the modulator and output voltage of the demodulator being shown on y-axes. The sensitivity of the modulator is 2.5 kHz/V and that of the demodulator is 0.4 V/kHz. The input of the modulator is bipolar $\pm$ 1 V range. The output of the demodulator is unipolar and has to be given a dc shift in order to get bipolar output.

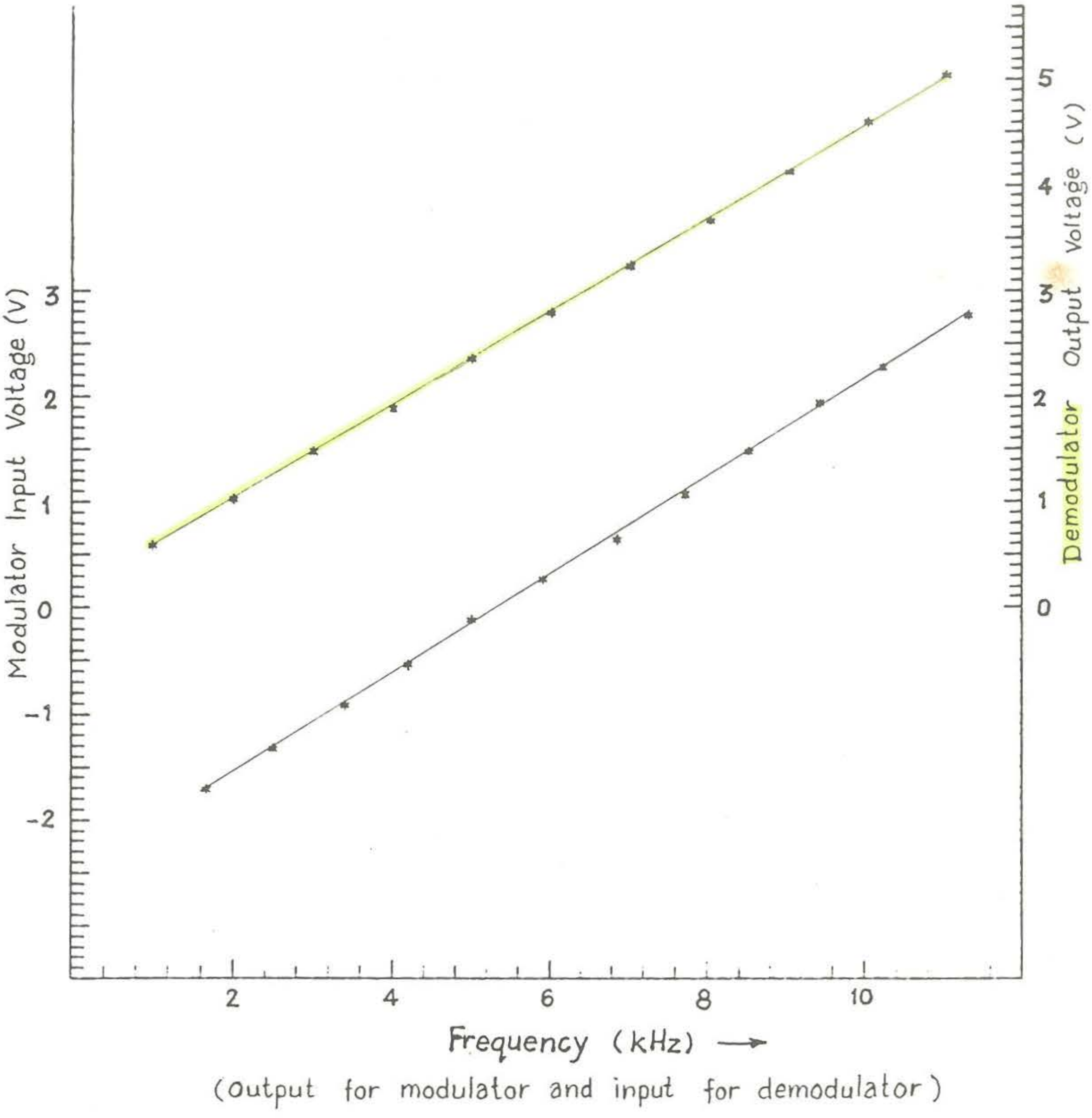


Fig 4.5 Static transfer characteristics of modulator and demodulator.

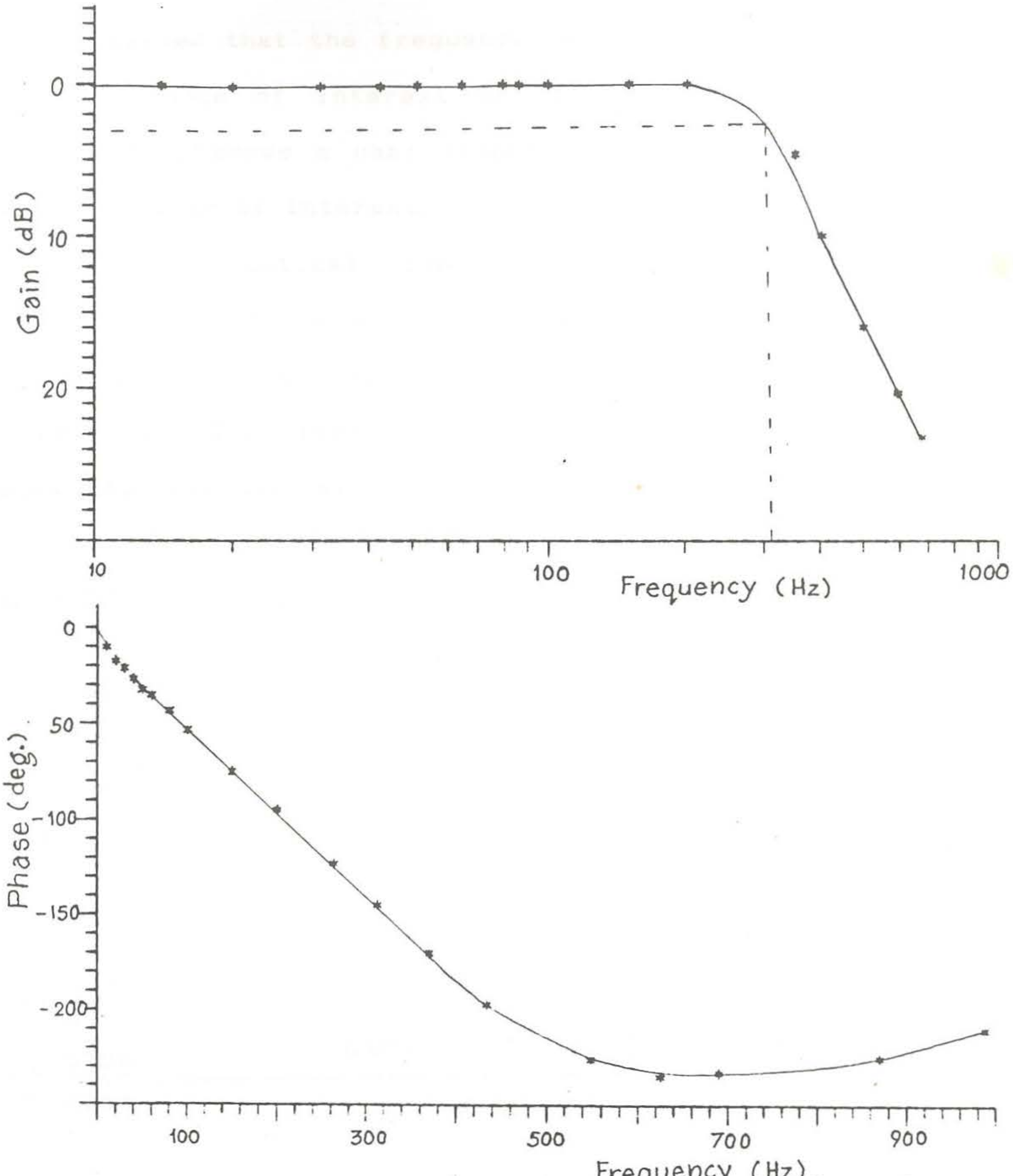


Fig 4.6 (a) Frequency and (b) Phase response of the modulator-demodulator circuit.

Fig 4.6 (a) and (b) show the frequency and phase response of the modulator and demodulator cascaded together. It is observed that the frequency response is flat over the recording range of interest of dc to 300 Hz. The phase response also shows a near linear characteristics over the frequency range of interest.

Two identical boards for two channels were fabricated each containing a modulator and a demodulator. The printed circuit board design of the board is shown in Appendix B. The specifications of the modulator and demodulator circuits are given in Appendix C.

4.6 BIOSIGNAL AMPLIFIER AND ECG SIMULATOR

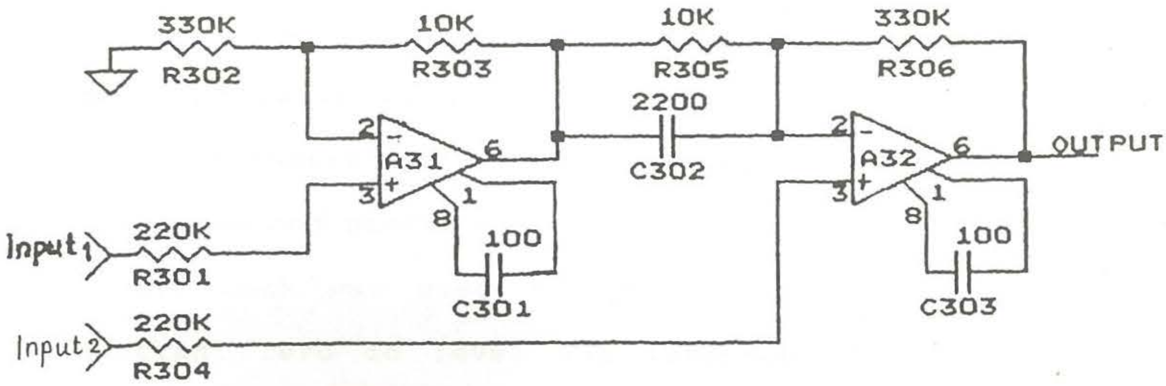


Fig 4.7 Biosignal amplifier

Fig 4.6 (a) and (b) show the frequency and phase response of the modulator and demodulator cascaded together. It is observed that the frequency response is flat over the recording range of interest of dc to 300 Hz. The phase response also shows a near linear characteristics over the frequency range of interest.

Two identical boards for two channels were fabricated each containing a modulator and a demodulator. The printed circuit board design of the board is shown in Appendix B. The specifications of the modulator and demodulator circuits are given in Appendix C.

4.6 BIOSIGNAL AMPLIFIER AND ECG SIMULATOR

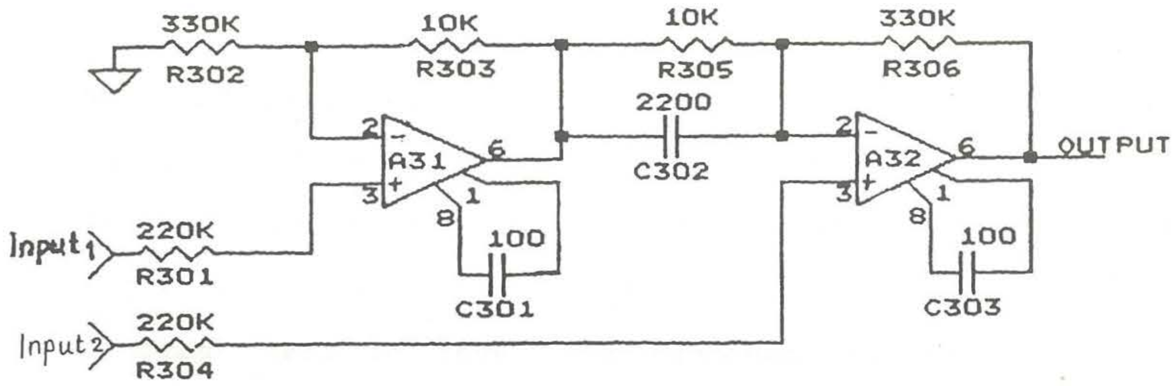


Fig 4.7 Biosignal amplifier

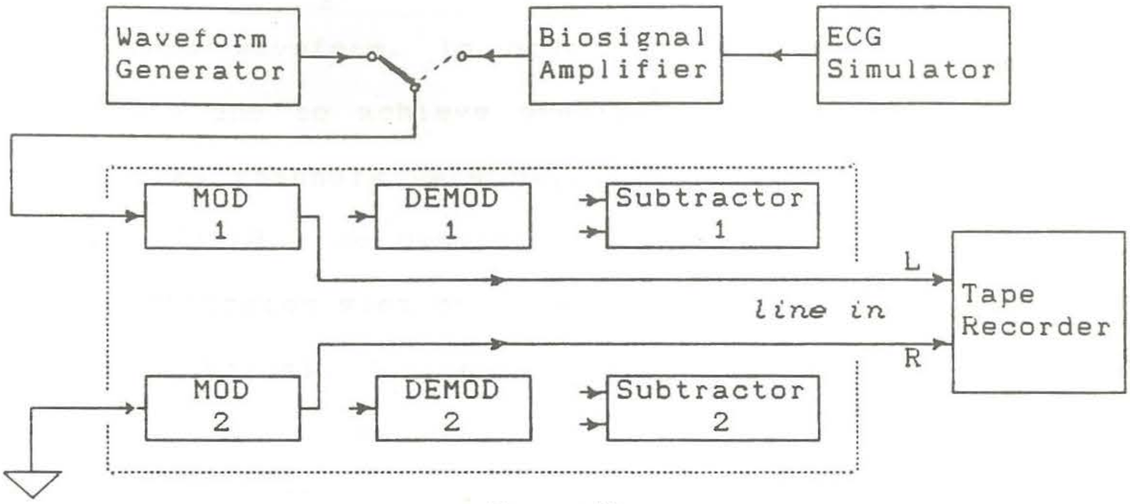
In order to test the hardware on the ordinary tape recorder, a simulated ECG was used in addition to the sinusoidal signals from the function generator. Functional block diagram and the circuit for the ECG simulator used are shown in the Appendix D. The simulated ECG was amplified by the biosignal amplifier module shown in Fig 4.7 [23] was used. The opamp A_{31} is used as a non-inverting amplifier which feeds opamp A_{32} working in inverting amplifier mode. The gain of the amplifier is given by the formula,

$$G = 1 + \frac{R_{302}}{R_{303}} \qquad (4.9)$$

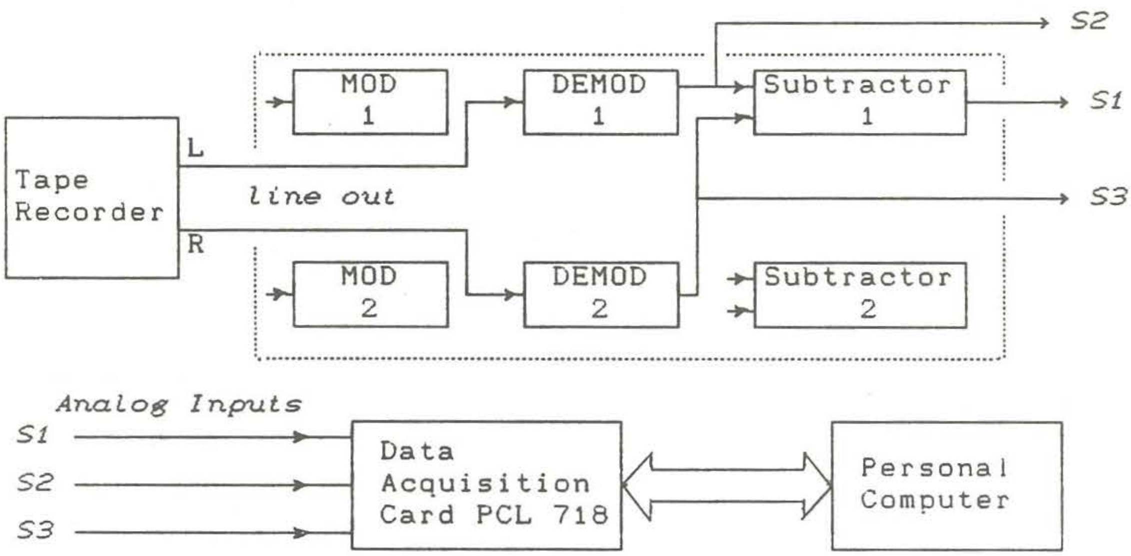
The specifications of the amplifier are given in Appendix I.

4.7 TESTING OF HARDWARE AND RESULTS

The schematic of the set-up for testing of the hardware is shown in Fig 4.8. The signal from either the function generator or from biosignal amplifier was frequency modulated and stored on left track of the tape recorder. The right track was used as the reference track on which a constant zero dc level was recorded. On playback of the signals, the retrieved waveform from right channel is subtracted from that on the left channel. Thus, we have three waveforms: S_1, the compensated signal by analog



Recording



S1 is compensated output by compensation C1,
S2 is uncompensated output,
S3 is reference waveform.

Playback

Fig 4.8 Block schematic of the experimental set-up

subtraction: S_2, the uncompensated signal; and S_3, the reference waveform. In order to display the played back waveform and to achieve precise compensation, the data of the three channels were digitized using a data acquisition card (PCL718 - by Dynalog Micro Systems, Bombay) installed in an extension slot of a personal computer.

Signal S_1 gives the compensation by analog subtraction of the waveforms from two tracks and will be hereafter referred as CA1. The waveform obtained after using digital subtraction of S_2 and S_3 for compensation waveform is referred as CD1. The second type of compensation as explained in section 3.4 can be done either after analog subtraction or after digital subtraction. Value of δ is obtained using Eq 3.6. x'_o in this equation is nothing but S_3. The value of δ thus obtained is substituted in Eq 3.8, in which, $x'-x'_o$ is nothing but S_1. This gives the compensated signal CA1D2. Subtraction $x'-x'_o$ is also calculated digitally using samples of S_2 and S_3 and then the further compensation can be applied in the similar way to get compensation CD1D2. Thus,

$$CA1 \rightarrow S_1(n),$$

$$CA1D2 \rightarrow S_1(n)/(1+\delta),$$

$$CD1 \rightarrow S_2(n)-S_3(n),$$

$$CD1D2 \rightarrow [S_2(n)-S_3(n)]/(1+\delta).$$

where, $\delta = S_3(n)/k_2\omega_c$. Recordings were made on three different tape recorders (SANYO MW700K. PHILIPS DR628, AIWA ADR707) using three different types of cassette tapes (TDK, MELTRACK, SONY).

The records were made with dc level, sinusoidal waveforms as well as simulated ECG waveform. Fig 4.9(a) to (e) show playback waveforms of dc record using different types of compensation of dc signals. The quantitative discussion of the results is given in the next section. There is a remarkable improvement in the signal using compensation CA1 and CD1 which is observed by comparing Fig 4.9(a) with 4.9(b) and 4.9(c). It is also observed that the improvement is more if we use CA1 than if CD1 is used as demonstrated by Fig 4.9(b) and (c). No significant differences are observed by incorporating second type of compensation as seen in Fig 4.9(d) and 4.9(e).

Fig 4.10 shows the records made using sinusoidal signals from function generator. Here also, we see the improvement in the signal on compensation using reference track. The extent of improvement however differs from recorder to recorder. For example, although the noise due to flutter is more while recording on SANYO than on AIWA, the former shows a greater improvement on compensation as demonstrated by Fig 4.10 (a) and (b).The PHILIPS recorder shows more noise due to flutter than the other two and also shows comparatively less improvement on compensation as seen

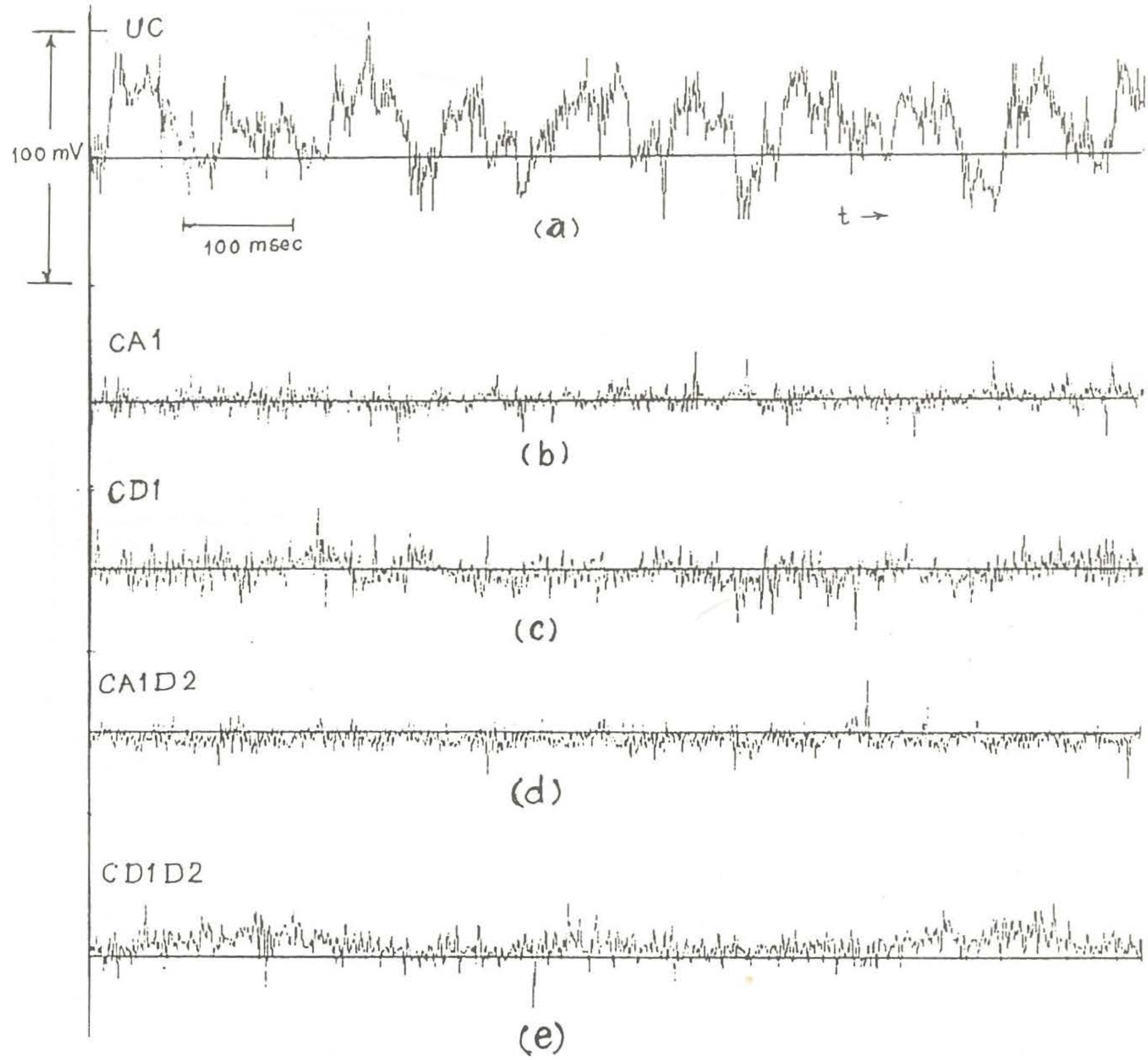


Fig 4.9 Playback waveform of a dc level record with different types of compensations.

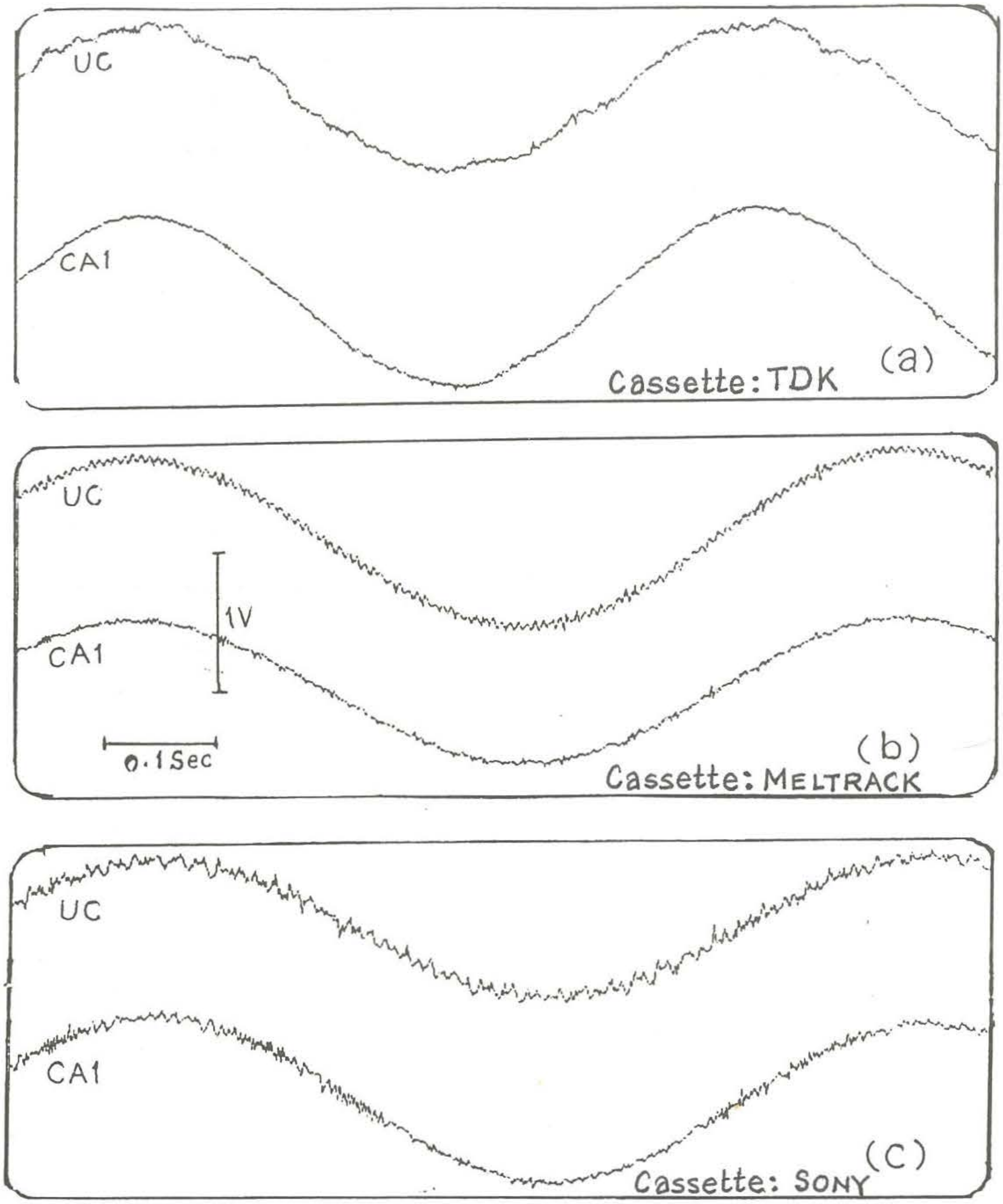


Fig 4.10 Uncompensated (UC) and compensated (CA1) waveforms using sinusoidal waveforms recorded on (a) SANYO (b) AIWA and (c) PHILIPS recorder.

from Fig 4.10(c). Fig 4.11 shows the simulated ECG waveforms before and after compensation. They also show the improvement in the waveform after applying AC1.

The exercise of recording on one recorder and playing back on the other was also carried out. In Fig 4.12 the first mentioned recorder is the one on which the waveform was recorded and the next mentioned is the one on which it was played back. The name of the cassette is also mentioned at the end. It is observed that the distortion due to flutter increases and the improvement in the signal on compensation reduces if the cassette is played back on a different recorder.

From the above mentioned observations, one can say that the method of recording a reference waveform on one track and using it for correction of the signal on the other track on playback shows an improvement in the recorded signal. Variations in the extent to which improvement is achieved could be due to the velocity variations across the tape, i.e., variations on one track that are independent of those on the other track . Such variations cannot be taken care by this compensation method. The recorders in which the irregularities in the tape speed fall under this category are negligible are expected to show a greater improvement.

Also, it is observed from the dc records that the digital subtraction cannot improve the waveform as well as

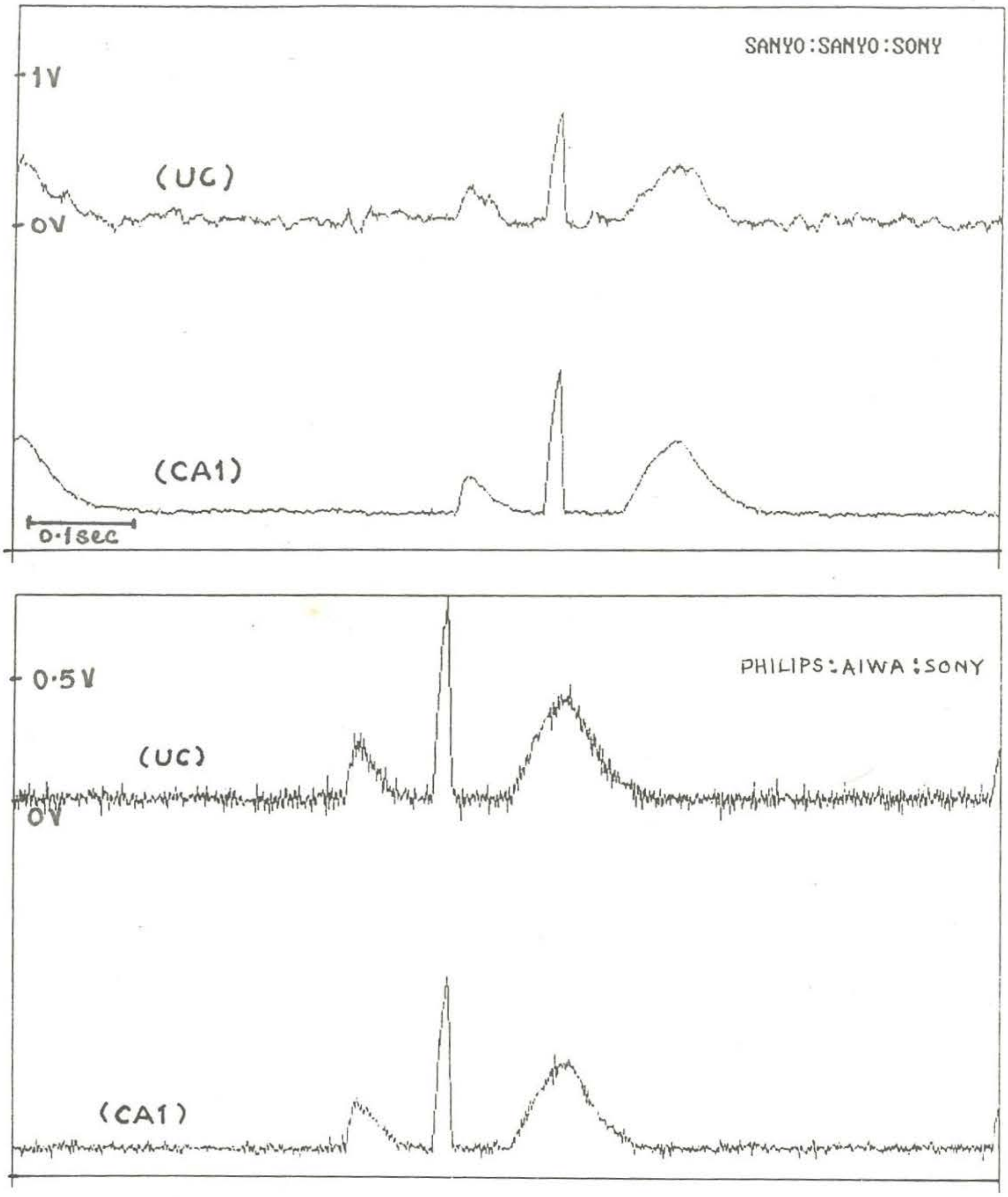


Fig 4.11 Uncompensated (UC) and compensated (CA1) ECG waveforms recorded using ECG simulator (a) SANYO for recording and playback, (b) PHILIPS for recording, AIWA for playback.

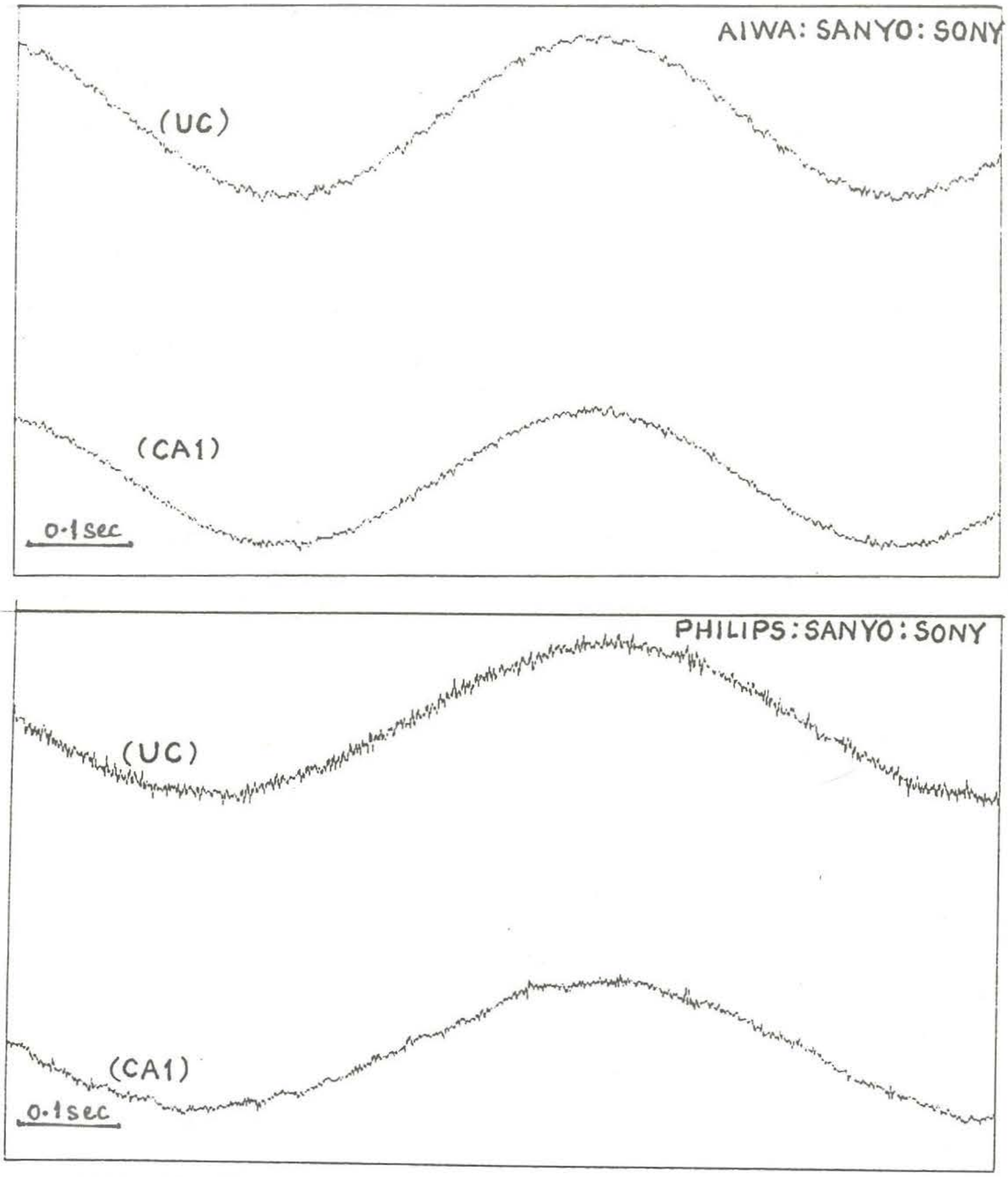


Fig 4.12 Uncompensated (UC) and compensated (CA1) sinusoidal waveforms recorded on one machine and played back on the other machine. (1 V_{p-p})

analog compensation because in analog subtraction, the subtraction of the reference and recording tracks is done at the same time instance. In digital subtraction, however, we subtract the value of the reference waveform that is occurring at a later time from the sample form recording track because ADC samples the two tracks sequentially and not simultaneously. The next section describes the quantitative results obtained using the dc records.

4.8 QUANTITATIVE ESTIMATION OF SIGNAL ERRORS

One of the major criterion for evaluation of the performance of an instrumentation tape recorder is signal-to-noise ratio. Because this figure varies considerably depending upon the test procedure [15], this section describes the detailed procedure by which the quantitative results of the noise introduced by flutter was estimated.

In order to find out the quantitative estimate of the noise occurring due to tape flutter before and after compensation, the digitized samples $x(n)$ of the dc records were used. In order to have an estimate of noise, rms value of ac error for a record length of N samples was obtained using the following steps,

$$X_{mean} = \frac{\Sigma\ x(n)}{N} \qquad (4.10)$$

$$X_{rms} = \sqrt{\frac{\Sigma\ [x(n)]^2}{N}} \qquad (4.11)$$

$$X_{ac} = \sqrt{X^2_{rms} - X^2_{mean}} \qquad (4.12)$$

AC error in dB with reference to full scale record level

$$X_{ac(dB)} = 20 \quad \log_{10} \frac{X_{ac}}{X_{max}} \qquad (4.13)$$

The results obtained by this procedure to get ac error in dB are given in tabular form in Table 4.1 and in graphical form in Fig 4.13.

Table 4.1 shows an improvement of about -9 dB over the uncompensated waveform on compensation by analog subtraction. The compensation by digital subtraction however gives less improvement of about -4 dB over the uncompensated waveform. This may be caused by the fact that the sampling is carried out by the ADC sequentially and not simultaneously. Thus, the sample used to correct the signal from recording track is actually occurring at a later instance of time. One must have a simultaneously sampling ADC to have an improvement as well as the analog subtraction. Hence, the exercise was repeated for a data sample sampled at 1 kHz and another at 5 kHz. The later sample was downsampled by considering only every fifth data

Table 4.1 AC error in dB with reference to full scale recording level with different types of compensation methods applied to records of dc levels. Sampling rate 3 kHz.
R - Waveform on reference track, UC - Signal retrieved without any compensation, CA1 - Signal after compensation by subtraction by analog circuit, CD1 - Signal after compensation by digital subtraction, CA1D2 - Compensation 1 by analog circuit, compensation 2 digitally, CD1D2 - Both types of compensations digitally.

Signal	AC error (dB)		
	V_s = 0 V	V_s = 0.5 V	V_s = 1 V
R	- 29.3	- 29.8	- 28.8
UC	- 30.3	- 28.7	- 26.6
CA1	- 39.8	- 37.7	- 36.8
CD1	- 34.2	- 32.9	- 33.3
CA1D2	- 38.9	- 37.9	- 37.3
CD1D2	- 35.6	- 32.2	- 33.4

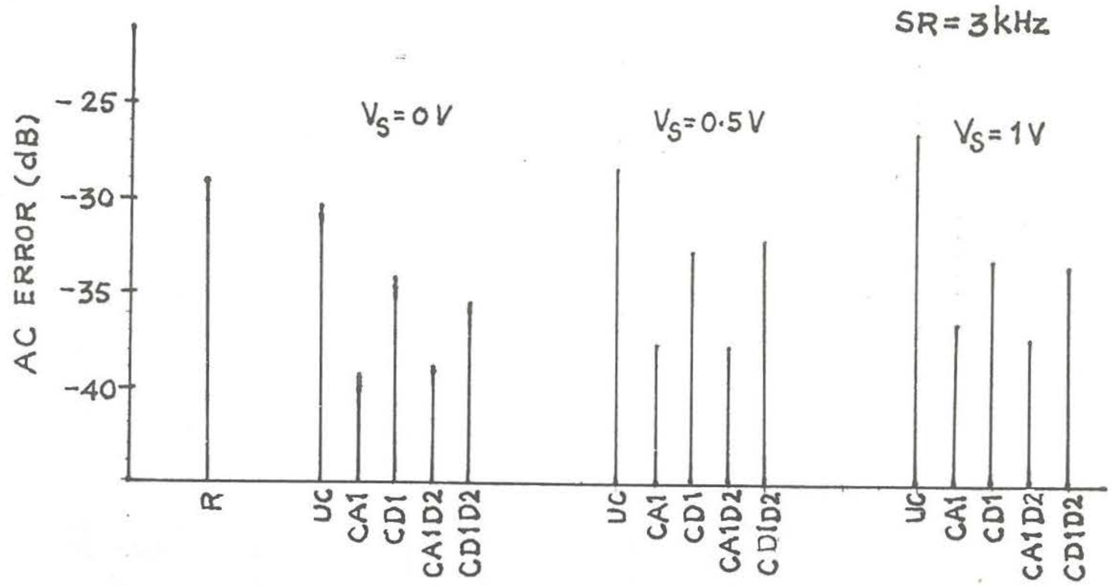


Fig 4.13 AC error in dB with reference to fullscale recording using different types of compensation at different dc voltages (Sampling rate SR = 3 kHz)

point for calculation. The record used was 2 sec. This has reduced the delay from 333 μs (for 1 kHz sampling rate) to 67 μs (for 5 kHz sampling rate) in sampling from the reference track to recording track. The results are show in Table 4.2 and are also presented graphically in Fig 4.14.

Table 4.2 AC error in dB with reference to full scale recording level for different types of compensation methods applied at two different sampling rates.

Signal	ac error (dB)					
	V_s = 0 V		V_s = 0.5 V		V_s = 1 V	
	SR= 1 KHz	SR= 5 KHz	SR= 1 KHz	SR= 5 KHz	SR= 1 KHz	SR= 5 KHz
R	- 31.6	- 30.2	- 29.7	- 32.9	- 30.1	- 30.8
UC	- 30.3	- 30.0	- 29.9	- 31.0	- 29.0	- 29.7
CA1	- 41.0	- 40.3	- 38.6	- 39.1	- 38.3	- 38.2
CD1	- 35.1	- 37.2	- 33.6	- 37.0	- 33.6	- 38.2
CA1D2	- 40.8	- 40.0	- 39.0	- 39.8	- 38.9	- 38.4
CD1D2	- 36.0	- 36.9	- 33.9	- 37.3	- 34.1	- 38.9

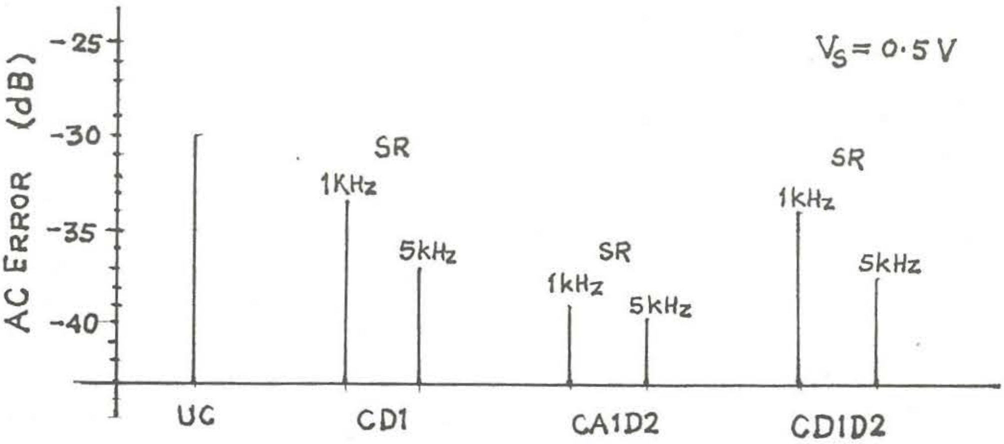


Fig 4.14 AC error in dB with reference to fullscale after digital compensation for two sampling rates (Recorded signal 0.5 V dc).

The table shows that there is an improvement in the dB error using digital subtraction CD1 when the sampling rate is increased. This shows that the increased sampling rate reduces the delay between the samples taken from the reference track and recording track which results in use of more correct value for compensating the signal. The improvement in the second type of compensation is however marginal. It is also observed that CA1 gives the best performance of all the methods suggesting that for precise compensation using digital methods, one needs to have a near simultaneous sampling ADC.

C H A P T E R - V

CONCLUSION AND SUGGESTIONS FOR FURTHER WORK

5.1 CONCLUSION

The aim of this project was to record low frequency biological signals on commercially available magnetic cassette recorder. To overcome the poor low frequency response of the cassette recorder, technique of frequency modulation was selected. An attempt was made to compensate for the errors in recording by using one of the tape tracks as a reference track.

The results obtained show that, low frequency biosignals can be recorded on the commercially available cassette recorder by the technique of frequency modulation. The major effect of tape flutter can be eliminated by using one of the tape tracks as reference track and recording a constant dc level which is equivalent to recording a stable frequency. There is remarkable improvement in the played back signal on subtraction of the reference track signal from the recording track signal. The improvement is not equally good in all the types of recorders because the tape movement irregularities are not identical in all of them. The method for compensation used by this system is based on the assumption that the effect of the tape motion irregularities are equal on any line perpendicular to the length of the tape. But it has been reported that there are also some irregularities that are not uniform across the tape width [16]. The effects of such irregularities are not eliminated by this method. The recorders in which the irregularities falling under later category are negligible are expected to show more improvements on compensation by this method.

The method of compensating for flutter by digital subtraction did not prove to be as successful as analog subtraction as seen from the quantitative analysis of the dc records. This can be because of the fact that the sample taken for compensation from reference track to compensate for a sample on record track actually occurs at a different instance in time because of the finite sampling rate of ADC.

5.2 SUGGESTIONS FOR FUTURE WORK

To have a general purpose system for recording various types of biological signals, the amplifier module should have some additional features like signal isolator, notch filter, offset removal facility etc. From the recorder point of view, it will be more useful to have a facility to playback at two different speeds so that the data of least importance can be viewed at a faster speed. It is also desirable to have certain time scale marking on one of the tracks as a time reference.

For precise compensation of flutter, to improve the compensation by digital method, one needs to sample the waveforms from the two tracks simultaneously to get the correct sample for correction. This can be achieved by using an ADC which could sample at a rate as high as 100 kHz or using sample and hold circuits.

APPENDIX-A

DESIGN OF FOURTH ORDER BUTTERWORTH FILTER

The transfer function of the fourth order Butterworth filter is given as,

$$H_4(s) = \frac{k}{(s^2+a_1s+b_1)\ (s^2+a_2s+b_2)}$$

Thus, a fourth order Butterworth filter can be realized by cascading two second order filter sections. A second order section is shown in Fig A.1.

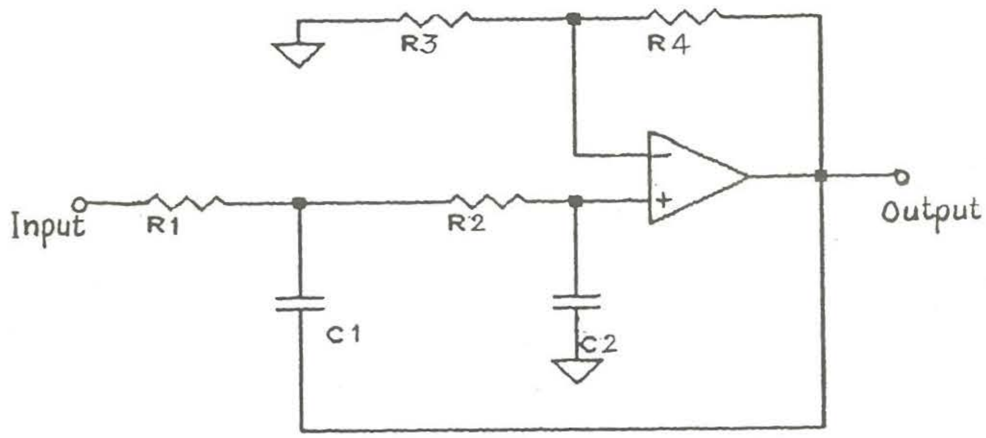


Fig A.1 Second order filter

The transfer function of this section is given by, [24]

$$\frac{V_o(s)}{V_i(s)} = \frac{\dfrac{A}{R_1C_1R_2C_2}}{s^2 + \dfrac{R_1C_1+R_2C_2+R_1C_2-AR_1C_1}{R_1C_1R_2C_2}\, s + \dfrac{1}{R_1C_1R_2C_2}}$$

where, $A = 1+\dfrac{R_4}{R_3}$

Selecting, $R_1=R_2=R$ and $C_1=C_2=C$, we get,

$$\frac{V_o(s)}{V_i(s)} = \frac{\dfrac{A}{R^2\ C^2}}{s^2 + \dfrac{3-A}{R\ C}\, s + \dfrac{1}{R^2C^2}}$$

By replacing $1/RC$ by ω_c, we get

$$\frac{V_o(s)}{V_i(s)} = \frac{A\ \omega_c^2}{s^2 + (3-A)\omega_c s + \omega_c^2}$$

Thus, the transfer function of the fourth order Butterworth filter after cascading two second order sections can be given by,

$$\frac{V_o(s)}{V_i(s)} = \frac{A_1A_2\omega_c^4}{[\ s^2 + (3-A_1)\omega_c s + \omega_c^2\ ][\ s^2 + (3-A_2)\omega_c s + \omega_c^2\ ]} \quad (A.1)$$

To realize the Butterworth response using two second order sections similar to shown in Fig A.1, consider the transfer function of the fourth order Butterworth filter having a cutoff frequency at ω_c.

$$B_4(s) = \frac{\omega_c^4}{(s^2 + 1.84776\ \omega_c s + \omega_c^2)(s^2 + 0.76536\ \omega_c s + \omega_c^2)} \qquad (A.2)$$

Comparing the denominators of Eq A.1 and A.2, we get,

$3 - A_1 = 1.84776$, $\quad \therefore A_1 = 1.15224$

$3 - A_2 = 0.76536$, $\quad \therefore A_2 = 2.23464$

Hence, for the first section, R_3=15 kΩ, and R_4 = 2.2 kΩ.
For the second section, R_3=18 kΩ, and R_4 = 22 kΩ.

To have a cutoff frequency at about 300 Hz, $1/RC = 2\pi f_c = 300$. Selecting $C = 0.022\mu F$, we get $R = 24.11$ kΩ. Selected value of R is 22 kΩ. The overall gain of the filter is 2.57. The following table gives the design values of the components used in the second order Butterworth filter.

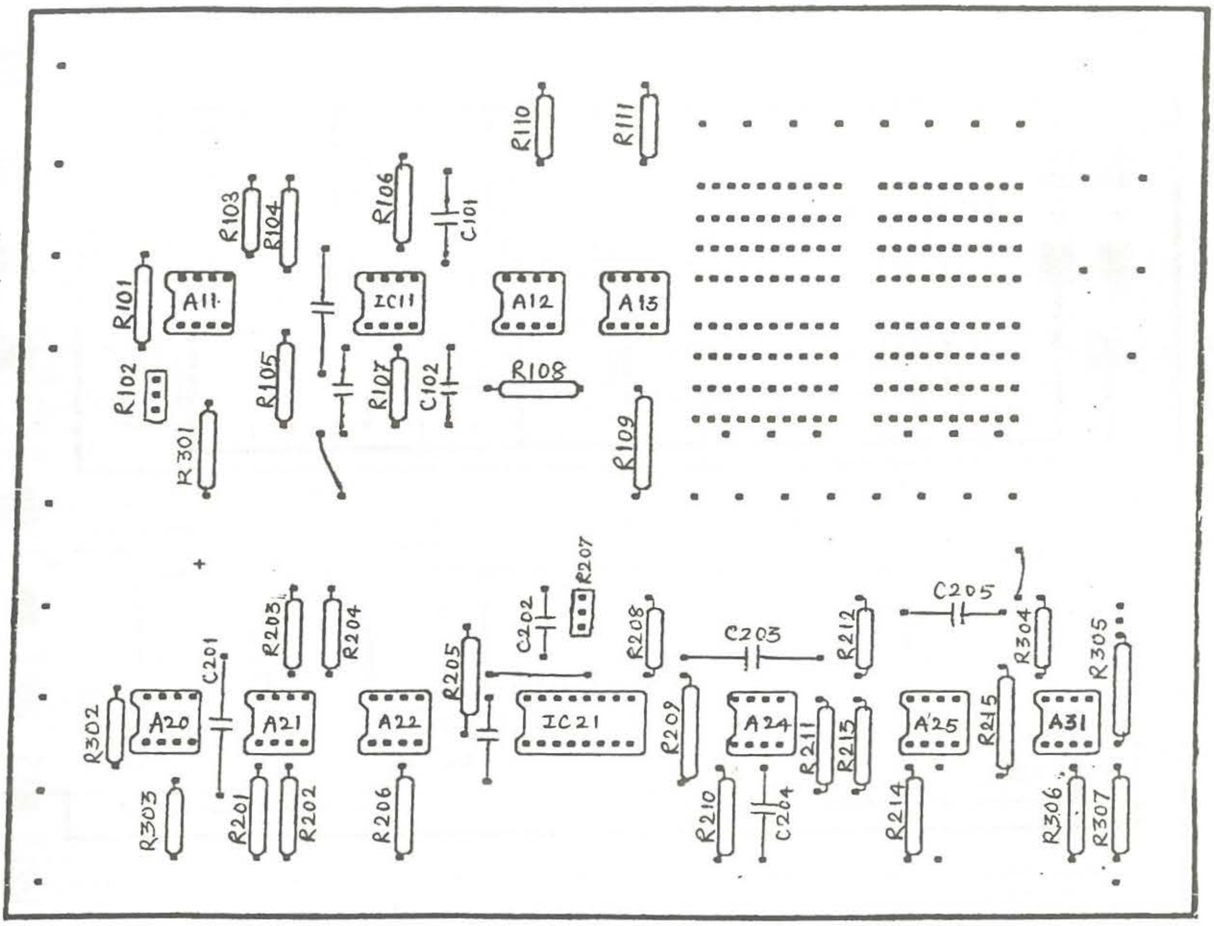
Table A.1 Design values of the component with reference to Fig 4.4.

Section	Component	Corresponding component in Fig 4.4	Value
Section I	R_1	R_{208}	22 kΩ
	R_2	R_{209}	22 kΩ
	R_3	R_{210}	15 kΩ
	R_4	R_{211}	2.2 kΩ
	C_1	C_{203}	0.022 μF
	C_2	C_{204}	0.022 μF
Section II	R_1	R_{212}	22 kΩ
	R_2	R_{213}	22 kΩ
	R_3	R_{214}	18 kΩ
	R_4	R_{215}	22 kΩ
	C_1	C_{205}	0.022 μF
	C_2	C_{206}	0.022 μF

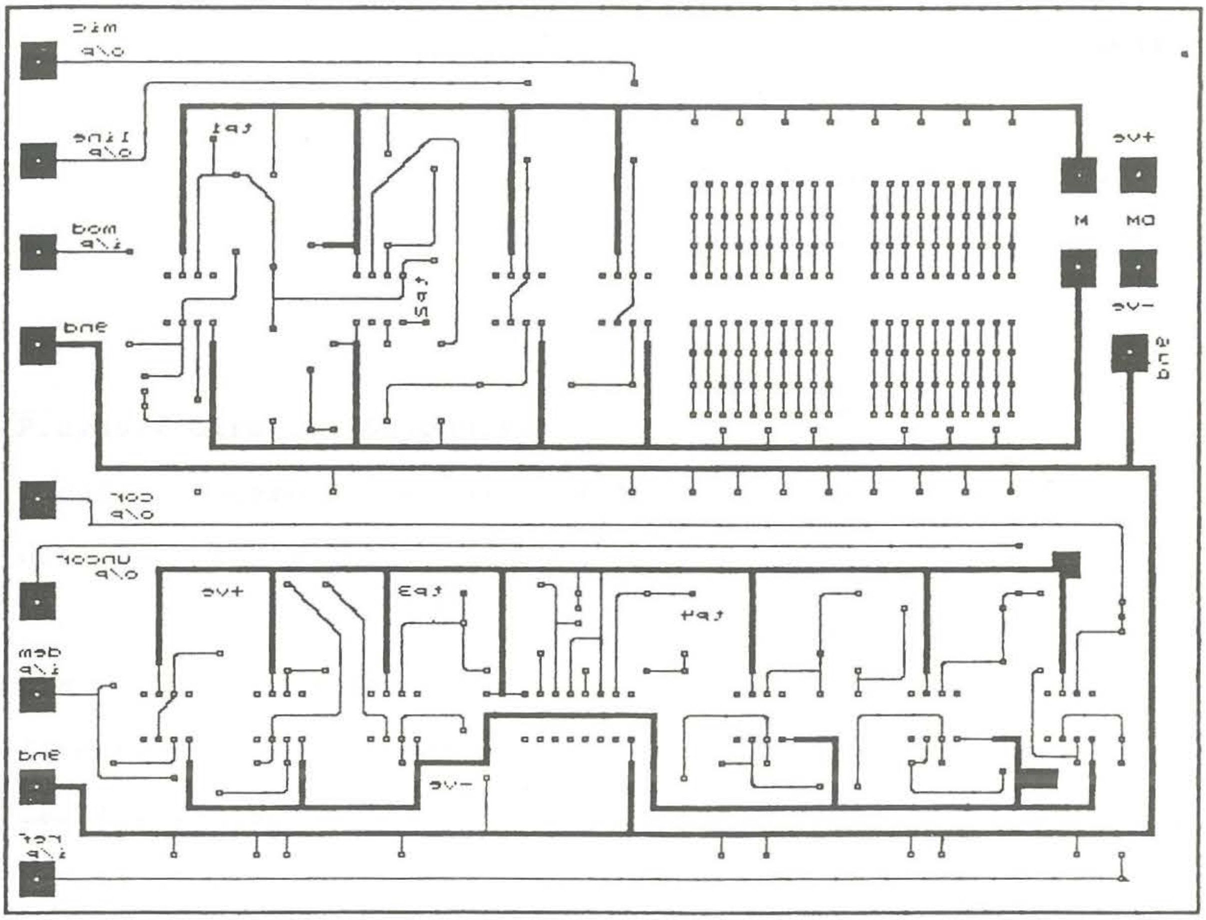
APPENDIX - B

PCB LAYOUT OF MODULATOR DEMODULATOR CIRCUIT

Component Layout



Solder Side Layout



APPENDIX - C

INSTRUMENT SPECIFICATIONS

Recording Circuit (Modulator)

Input : Single ended, impedance 50K, Maximum voltage level $\pm$ 1 V.

Output : 50 mV square wave for mic output, 250 mV square wave for line output. Peak recording current 20 mA, impedance 100 Ω.

Playback Circuit (Demodulator)

Input : Impedance 2MΩ, Minimum level required 50 m V_{p-p}.

Output : Single ended, Impedance 100 Ω.

System Specifications

Tape Speed : 1 $\frac{7}{8}$ inches per sec.

Carrier center frequency 5.5 kHz

Deviation from center frequency 2.5 kHz

Signal Bandwidth : 0-300 Hz

AC error after flutter compensation :- 40 dB with reference to fullscale

Amplifier Specifications

CMRR : 70 dB, Gain : 50, Input impedance: 5MΩ

Power Supply

Voltage $\pm$ 9 V.

Current drain (mA)

	+ve (mA)	-ve (mA)
Modulator	18	18
Demodulator	32	32
Amplifier	2	2
Total drain	52	52

A P P E N D I X - D

ECG SIMULATOR

The circuit for ECG simulator has been borrowed from a B.E. Project report by Deshpande [25]. The block diagram explaining the function of the ECG simulator is shown in Fig D.1. The circuit diagram of the same is given in Fig D.2.

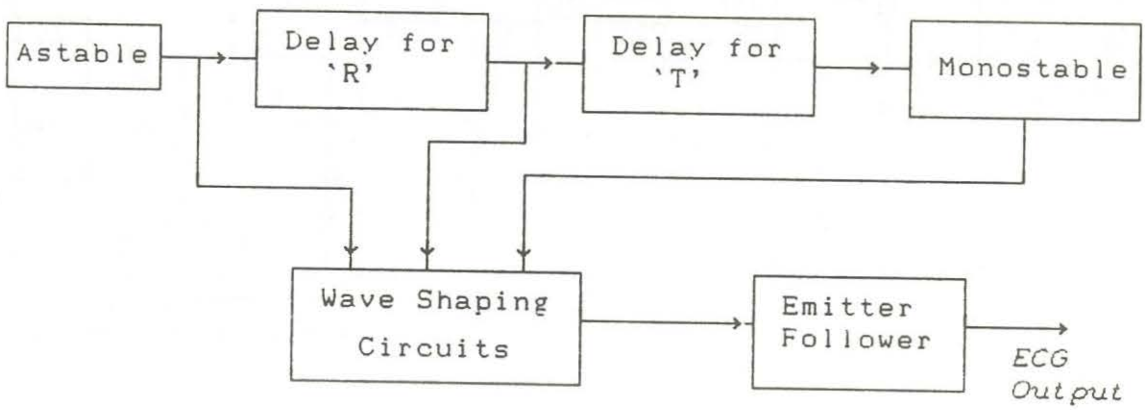


Fig D.1 Block schematic of ECG simulator

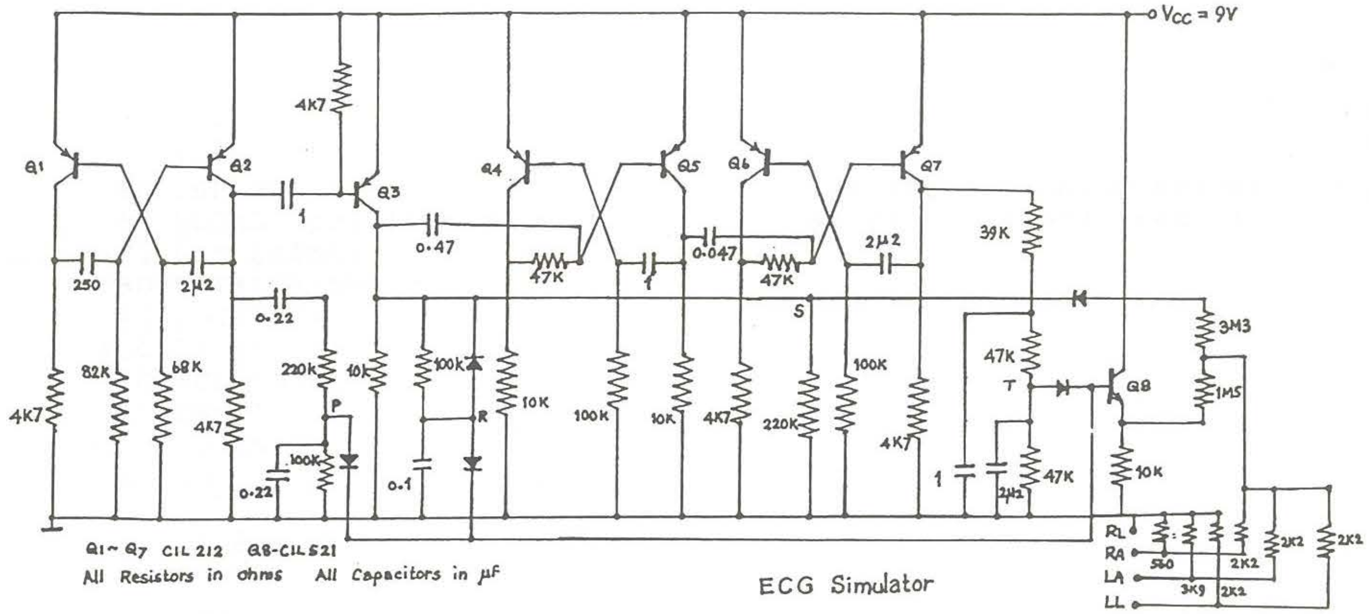


Fig D.2 Circuit diagram of ECG simulator

A P P E N D I X - E

PROGRAMME LISTING

/* This programme acquires data from three channels using data acquisition card PCL718. The waveform from reference track, recording track and waveform after their analog subtraction is acquired sequentially. The programme also calculates the ac error of the wavwform as explained in Chapter IV */

```
 10 SCREEN 10
 20 CLS
 30 I=0
 40 DIM DARR%(17999)
 50 DEF SEG=&H5000                       'Loading of PCL718 driver
 60 BLOAD "pcl718.bin",0                 'at given memory segment
 70 DIM DAT%(4)
 80 DAT%(0)=&H300
 90 DAT%(1)=4
100 DAT%(2)=1
110 PCL718=0
120 ER%=0
130 FUNC%=0
140 CALL PCL718(FUNC%,DAT%(0),ER%)       'Installation check
                                         'for PCL718 card
150 IF ER%<>0 THEN PRINT
    "INSTALLATION FAILED !"
160 PRINT "ENTER START CHANNEL NO :"     'Setup the sequence of
170 INPUT DAT%(0)                        'scanning channels
180 DAT%(1)=DAT%(0)+2
190 FUNC%=1
200 CALL PCL718(FUNC%,DAT%(0),ER%)
210 IF ER%<>0 THEN PRINT"SET SCAN
    CHANNEL FAILED"
220 PRINT "ENTER SAMPLING RATE IN Hz" 'Set the sampling rate
230 INPUT SR
240 IF SR<.15 OR SR>5000 GOTO 230
250 SR=SR*3
260 DAT%(0)=100
270 DAT%(1)=100000!/SR
280 FUNC%=17
290 CALL PCL718(FUNC%,DAT%(0),ER%)
300 IF ER%<>0 THEN PRINT
    "SET PACER FAILED"
310 FUNC%=5
320 DAT%(0)=18000
330 DAT%(1)=&H4000
```

```
340 DAT%(2)=1
350 DAT%(3)=0
360 CALL PCL718(FUNC%,DAT%(0),ER%)
370 IF ER%<>0 THEN PRINT"INTERRUPT        'Transfer of datablock
    MODE FAILED"                          'in interrupt mode
380 FUNC%=8
390 CALL PCL718(FUNC%,DAT%(0),ER%)
400 IF ER%<>0 THEN PRINT
    "READ STATUS FAILES"
410 IF DAT%(1)=1 GOTO 390
420 DAT%(0)=18000
430 DAT%(2)=0
440 DAT%(1)=&H4000
450 DAT%(3)=VARPTR(DARR%(0))              'Transfer the data to
                                          'an array

460 FUNC%=9
470 CALL PCL718(FUNC%,DAT%(0),ER%)
480 IF ER%<>0 THEN PRINT
    "TRANSFER DATA FAILED"
490 CLS
```

/* The calculation of ac error in dB is done using the Eq 4.10 to 4.13 */

```
500 LINE (0,0)-(639,0)
510 LINE (639,0)-(639,349)
520 LINE (0,349)-(639,349)
530 LINE (0,0)-(0,349)
540 FOR I=0 TO 5998
550 AVGR=AVGR+DARR%(3*I+1)
560 RMSR=RMSR+DARR%(3*I+1)*DARR%(3*I+1)
570 AVGU=AVGU+DARR%(3*I)
580 RMSU=RMSU+DARR%(3*I)*DARR%(3*I)
590 AVGA1=AVGA1+DARR%(3*I+2)
600 RMSA1=RMSA1+DARR%(3*I+2)*DARR%(3*I+2)
610 AVGD1=AVGD1+(DARR%(3*I)-DARR%(3*I+1))
620 RMSD1=RMSD1+((DARR%(3*I)-DARR%(3*I+1))
          *(DARR%(3*I)-DARR%(3*I+1)))
630 D=DARR%(3*I+1)/903
640 AVGA1D2=AVGA1D2+((DARR%(3*I+2)/(1+D))
650 RMSA1D2=RMSA1D2+((DARR%(3*I+2)/(1+D)
                   *(DARR%(3*I+2)/(1+D)))
660 AVGD1D2=AVGD1D2+((DARR%(3*I+1)-DARR%(3*I))/(1+D)
670 RMSD1D2=RMSD1D2+((DARR%(3*I+1)-DARR%(3*I))/(1+D)
                  *((DARR%(3*I+1)-DARR%(3*I))/(1+D)
680 NEXT I
690 PRINT "REFERENCE"
700 AVGR=AVGR/6000
710 RMSR=(RMSR/6000)^.5
720 XACR=(RMSR*RMSR - AVGR*AVGR)^.5
730 PRINT "dB=" 20*(LOG(XACR/420)/LOG(10))
740 PRINT "UNCOMPENSATED"
750 AVGU=AVGU/6000
```

```
760 RMSU=(RMSU/6000)^.5
770 XACU=(RMSU*RMSU - AVGU*AVGU)^.5
780 PRINT "dB=" 20*(LOG(XACU/420)/LOG(10))
790 PRINT "DIGITAL COMPENSATION"
800 AVGD1=AVGD1/6000
810 RMSD1 = (RMSD1/6000)^.5
820 XACD1=(RMSD1*RMSD1 - AVGD1*AVGD1)^.5
830 PRINT "dB= " 20*(LOG(XACD1/420)/LOG(10))
840 PRINT "ANALOG COMPENSATION"
850 AVGA1=AVGA1/6000
860 RMSA1=(RMSA1/6000)^.5
870 XACA1=(RMSA1*RMSA1 - AVGA1*AVGA1)^.5
880 PRINT " db= " 20*(LOG(XACA1/420)/LOG(10))
890 PRINT "A1D2"
900 AVGA1D2=AVGA1D2/6000
910 RMSA1D2=(RMSA1D2/6000)^.5
920 XCA1D2=(RMSA1D2*RMSA1D2 - AVGA1D2*AVGA1D2)^.5
930 PRINT "dB =" 20*(LOG(XACA1D2/420)/LOG(10))
940 PRINT "D1D2"
950 AVGD1D2=AVGD1D2/6000
960 RMSD1D2=(RMSD1D2/6000)^.5
970 XCD1D2=(RMSD1D2*RMSD1D2 - AVGD1D2*AVGD1D2)^.5
980 PRINT "dB =" 20*(LOG(XACD1D2/420)/LOG(10))
1000 FUNC%=7
1010 CALL PCL718(FUNC%,DAT%(0),ER%)
1020 IF ER%<>0 THEN PRINT "TERMINATION FAILURE"
1030 IF INKEY$="" GOTO 1030
1040 END
```

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