DEVELOPMENT AND CONSTRUCTION OF WEATHER DATA ACQUISITION SYSTEM WITH COMPUTER INTERFACING

BY

POPOOLA ADEWUMI ISAAC

A THESIS SUBMITTED TO THE DEPARTMENT OF PHYSICS, FEDERAL UNIVERSITY OF TECHNOLOGY, AKURE.

IN PARTIAL FULFILLMENT OF THE REQUIREMENT FOR THE AWARD OF THE DEGREE OF MASTER OF TECHNOLOGY IN ELECTRONIC INSTRUMENTATION AND MEASUREMENT

JANUARY 2004

CERTIFICATION

We certify that this research work and the result obtained were carried out by **Popoola Isaac** of the department of Physics Federal University of Technology Akure and has been approved as meeting the requirement for the award of M.Tech degree in physics of the Federal University of Technology Akure, Ondo State, Nigeria.

Dr. M. T. Babalola
B. Sc. Ph.D (Ibadan) – (Major supervisor)

2/3/2004
Date

Dr. O. S Ajayi
B.Sc. PhD - (Minor Supervisor)

09/03/04
Date

DEDICATION

This work is dedicated to Jehovah my Lord, who saw me through the successful completion of the master's study despite all odds. The thesis is also dedicated to my darling wife Margaret and daughter - Ini-oluwa.

ACKNOWLEDGEMENTS

I give all honour and adoration to my God for his care, love and directing spirit, who has been directing my path as to where I should walk in life.

I am very grateful to my supervisor Dr. M.T Babalola, for his kindness and contribution to the success of this work.

I very much appreciate the effort of Dr. M. O. Ajewole, the head of department of physics, F.U.T.A for his care and dynamic leadership.

I' am very grateful to Dr. O.S Ajayi for the co-supervision of this work and for proofreading the report.

I am also very grateful to Prof. and Dr. (Mrs.) Fuwape for their encouragement and trust in my ability to produce result.

I am most grateful to Prof. A.V. Popoola and family for their tutorship and support, which has created in me determination for success in life.

I very much appreciate the effort of Mr. Odo Ayodele who introduced me to the basic rudiments of computer interfacing and operation.

My sincere appreciation goes to my bosom friend "***Samuel Ayodele***" of the works department FUTA for his contribution to the successful and timely completion of this project. You will forever be a jewel in my hand.

I am grateful to Dr. and Mrs. A.B. Rabiu for their support, encouragement and prayers. May God continue to sustain us all. Amen. I also wish to appreciate Mr. Gbenga Olayanju of the department of geo-physics FUTA and his family for their kindness. God will continue to sustain you in life. Amen.

The kind gesture and benevolence of Mr. Ewetumo Theophilus towards the timely completion and realization of my objective in this work is highly appreciated.

To my loved one ***Margaret***, I say thank you a thousand times for being there for me despite all odds. Thank you for your love, care and understanding. God helping me, I shall forever love you.

Table of contents

page(s)

CHAPTER ONE: INTRODUCTION

CHAPTER TWO: LITERATURE REVIEW

CHAPTER THREE DEVELOPMENT AND CONSTRUCTION OF THE WEATHER DATA ACQUISITION SYSTEM

CHAPTER FOUR **RESULT EVALUATION, DISCUSSION AND RECOMMENDATION**

LIST OF FIGURES

LIST OF TABLES

ABSTRACT

A precise weather data acquisition system capable of measuring temperature, pressure and relative humidity has been developed. The system has three sensors AD590JH for temperature, HIH3610-001 for humidity and 24PCCF6DA for pressure. Each sensor drives a sensitive transducer amplifier with high stability and low thermal drift. For the purpose of data logging of the measured quantities, a computer-interface adapter was also developed. It includes a TC7109A 12bits microprocessor-compatible analogue-to-digital converter (ADC). The ADC was connected in a way to provide the necessary handshaking and interfacing required to input 12bits through 8bits data port available on the parallel port of the computer. The software for the control, data processing and presentation of the result on a visual display unit of the computer was written in Q-basic and Visual basic. The constructed weather data acquisition system demonstrates an excellent performance with accuracy's of ±0.3°C for temperature, 0.1mbar for pressure and ±0.5% for relative humidity

CHAPTER 1

INTRODUCTION

1.1 OVERVIEW

People's interest in the atmosphere is as old as the history of mankind. The impact of weather and climate on human life is so varied and all pervading that they naturally become the most important components of our environment. Information about weather parameters has helped man to improve in industrial and scientific exploration of his environment. The parameters often measured are many. They include; air temperature, pressure, humidity, precipitation, and amount of cloudiness. Many instruments have been used to monitor the atmosphere. The liquid-in-glass (mercury) thermometer is the standard equipment for measuring atmospheric temperature. Isaac et al., (1993), have described the development of a variety of precise temperature measuring circuits. It is however noticed that these circuits, especially those meant for high temperature measurement suffer from low accuracy and poor stability. Apart from the mercury thermometer, four other devices that have been developed and adapted for temperature measurement are listed out along with their properties in Table 1.1

Air pressure is measured most accurately with mercury barometer. Other pressure measuring devices are aneroid barometer, altimeter, and micro-barovariograph. Apart from the above-mentioned devices, many authors have described pressure sensors and their associated amplifying circuits. Franco, (1988) described many circuit requirements for effective pressure monitoring. He observed that since pressure-sensing materials are always arranged in a bridge, their associated amplifier must be a differential input type with a high input impedance (about 50 - 100 times larger than output impedance of the sensor). Also most of the materials used in manufacturing monolithic pressure sensors are temperature sensitive. Therefore, various techniques for temperature minimization/compensation such as constant current excitation must be employed. The variation of pressure with altitude is equally a well-documented fact. Pressure variation pattern shows that air pressure decreases by approximately 0.118mb/meter above sea level. This must be taken into consideration because a pressure sensor used at sea level will give values as recorded by meteorological reports or similar sources. If the sensor is used at heights greater than sea level, then the values it gives will be proportionately less than published data.

The measurement of humidity in the atmosphere becomes necessary because it is a useful index of dryness or dampness for determining evaporation, or absorption of moisture. Human comfort is dependent on relative humidity. On warm days, the weather is oppressive if the relative humidity is high but may be tolerable if the relative humidity is low. At low temperatures, comfort is not much affected by high relative humidity. However, very low relative humidity, which is common indoors during cold weather, can cause drying of skin or throat leading to discomfort or respiratory infections.

Hygrometer is the general name for humidity measuring equipment. A good example is the psychrometer, which consists of the wet and dry bulb thermometer. Another type of hygrometer, known as the *hair hygrometer* works on the basic principle that the length of hair changes with humidity. The electrical hygrometer is yet another instrument that uses a flat plate coated with a film of carbon to monitor humidity. When an electrical current is sent across the plate, the electrical resistance of the carbon coating changes as the moisture content of the air changes. Humidity sensors have hysteresis, and they require output linearization. Temperature compensation is required in some humidity sensors because of their high temperature coefficients.

Table 1.1 : Thermal sensors with their properties

Sensor	**Features**	**Typical useful temperature range**
Thermistors	Resistance falls with temperature It's output is non-linear. Its temperature/resistance relationship is not highly reproducible. Its sensitivity is high.	- 80 to 800°C
Thermocouple	Their voltage output rises with temperature. It has low sensitivity and high susceptibility to noise. There is requirement for cold junction compensation in many applications. These drawbacks impose limitations on its usage.	0 to 1000°C
Resistance Thermal Detectors(RTD's)	Their resistance rises with temperature. Their response is linear and stable. They are resistant to chemical effects	- 50 to 500°C
Semiconductor types	Their output voltage/current rises with increasing temperature. They have linear output. They are highly sensitive and cheap	- 55 to 150°C

1.2 OBJECTIVES AND SCOPE

This project is aimed at producing a weather data acquisition system that is completely tropicalljsed.

In addition to the general objective, the project aims to

(i) design a transducer amplifier of high stability and high precision.

(ii) device a means of faithfully transmitting the output signal from the transducer amplifiers to an Analogue-to-Digital converter located 10metres away from the point of measurement.

(iii) develop a 12bits interfacing adapter for a general purpose personal computer, using its printer's port as a channel for the control, acquisition and transmission of data and

(iv) develop software for the control of the attached interfacing circuitry. The software will also display the result on the visual display unit of the computer, while at the same time storing the acquired data against date and time in a data file for subsequent analysis/assessment.

1.3 BASIC HARDWARE AND SOFTWARE DESIGN FEATURES

Figure1.1 shows the block diagram of the development of weather data acquisition system. The system starts with three sensors, one each for temperature, pressure and relative humidity. Its respective amplifier amplifies the output from each sensor independently. The outputs from these amplifiers (which are d.c. voltages) are fed alternately through a multiplexer to an ADC. The conversion of the analogue signal from any of the connected amplifier to its digital form is done with a 12-bit A/D converter. The converter is a 12bits ADC IC (TC7109A) from Microchip Corporation. From the manufacturers data sheet, it is found that the device offers the following features: ***Low noise, zero integrator cycle for fast recovery from input overloads.*** Also, ***no zero adjustment is needed.***

The device can operate in two major modes: the **direct mode** and the **handshake mode**. In the direct mode, the device is allowed to make conversion continuously at a rate determined by the clock frequency (8192 clock periods per cycle). The handshake mode which is extensively used in this project, allows us to specify when the ADC should make conversion, and even to control the conversion time. This is done with the alternate selection of run/hold pin along with the byte enable control (see appendix D). The TC7109A is equipped with facilities for direct interfacing and multiplexing of data.

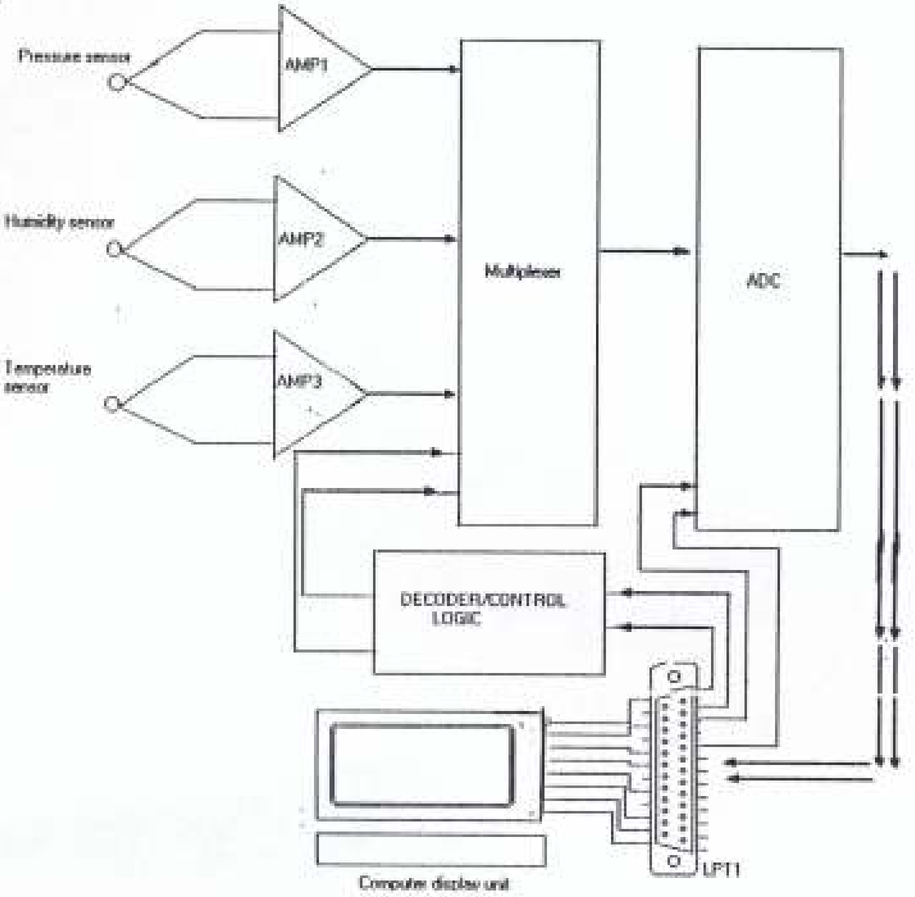


Fig. 1.1: The block diagram of the weather data acquisition system.

CHAPTER TWO

LITERATURE REVIEW

2.1 TEMPERATURE STANDARD AND SCALES

Temperature is a concept related to the flow of heat from one object or region of space to another. The term refers not only to the senses of hot or cold but also to numerical scales and thermometers as well. One major reason for devising a ***thermometer*** (literally, a meter for temperature) was that with it the composition of matter could be studied. Thermometers are systems with properties that change with temperature in a simple predictable, reproducible manner.

Over the years a number of different temperatures scale emerged, were tested and some survived while some others declined into uselessness. The most widely used scales today are the ***Fahrenheit***, the ***Celsius*** (or Centigrade) and the absolute or ***Kelvin*** scales. Experiments have shown that absolute zero corresponds to -273.15°C or -459°F. The centigrade scale has 0°C assigned to ice point and 100C (steam point) assigned to water boiling under the pressure of 1 atmosphere.

The absolute scale ***T***, the Celsius scale ***C*** and the Fahrenheit scale ***F*** are connected by the following formulae:

$$T = C + 273.15 \qquad (2.1)$$

$$\frac{C}{100} = \frac{F-32}{180} \qquad (2.2)$$

2.2 PRESSURE STANDARD

Pressure is expressed as the ratio of force to area. Its measurement is the determination of the magnitude of a fixed force applied to a unit area. Pressure measurements are generally classified as gauge pressure, absolute pressure or differential pressure. Pressures less than atmospheric are called vacuum.

In the laboratory, pressure is an important measurement, since pressure level has a significant effect on most physical, chemical and biological processes. In industry, particularly in the process industry, pressure is measured and controlled to maintain uniformity of product. It is also used to guide in safe plant operation, determine pumping head for fluid transfer and measure other variables such as weight, liquid level, temperature, flow and density of fluids and hydraulic forces indirectly. Pressure is variously measured in the following units: **Pascal, bar, psi** (per square inch)**, atmosphere, kg/cm^2, mmHg,** e.t.c. The standard atmospheric pressure is given as 760 mmHg. This is equivalent to 1013mbar. Other relationships in pressure units are; 1Pascal = $1N/m^2$, 1bar = $1.02*10^5N/m^2$, 1psi = $7.03*10^3$ N/m^2 and 1mmHg = $1.36*10^2$ N/m^2.

For pressures below $1.41*10^5N/m^2$ (138 kilopascals gauge), the universally accepted standard of measurement, both in the laboratory and in the industry, is the classic manometer using mercury or water. For higher pressures, the standard is the dead-weight tester. The principle is the balance of the force exerted by a precisely known weight on a piston of precisely measured area against a variable hydraulic pressure. Moreover, because of the advances made in device fabrication and technology, pressure measurements can now be taken with electrical pressure sensors up to about 2bar or so. All that we need to do is to rely on the manufacturer's calibration data sheet.

Pressure gauges generally fall in one of three categories, based on the principle of operation. These are liquid columns, expansible element gauges, and electrical pressure sensors. This project makes use of the electrical type of pressure sensor.

2.3 HUMIDITY STANDARD

Relative humidity is the ratio, in percent, of the moisture actually in the air to the moisture it could hold if it were saturated at the same temperature and pressure. A standard method of determining relative humidity is the wet and dry-bulb method, as in most hygrometers. Wet gauze is wrapped around a thermometer bulb, and a fast flow of air at room temperature is passed across it to cool it. The temperature drop (depression, typically 2 to 20C) is recorded along with the room temperature. The humidity is obtained by reference to a psychometric table.

2.4 TEMPERATURE SENSORS

Nearly all-electrical properties of a material or device vary as a function of temperature and could in principle be employed as a temperature sensor. The requirements of operation over a wide temperature range with high sensitivity, reproducibility and linearity greatly limit the possibilities, especially if cost, size and ease of use are also considered.

Since this project makes use of the solid-state thermal sensor, below is a run-down of the properties of the device.

They rely on the forward-voltage/temperature coefficient of a silicon diode, which is about -2mV/°C (Odo, 2002). Unfortunately the exact value of the effect varies considerably between samples, so calibration is necessary. However the effect is very linear and the sensors are extremely cheap. Examples of sensors in this category are the AD590, LM335 and LM35. They are usually two terminal integrated devices calibrated to produce a current of 1μA/°K (AD590) or a voltage of 10mV/°K (LM335).

They have a working range of about -50°C to 150°C over which the linearity is about 1%. Semiconductor monolithic temperature sensors can either have their output voltage/current proportional to absolute temperature (°K) or directly in degrees Celsius (°C).

Those sensors that produce an **O**utput **V**oltage that is **P**roportional **T**o **A**bsolute **T**emperature are referred to as _*VPTAT*_ sensor. While those with an **O**utput **C**urrents **P**roportional **T**o **A**bsolute **T**emperature are referred to as _*IPTAT*_ sensor.

2.4.1 VPTAT, IPTAT Sensors Versus Centigrade Sensors

In thermometer applications, it is preferable that the sensors output be calibrated in °C or in °F rather than in°K. For instance a VPTAT having a temperature coefficient of 10mV/°K, will produce an output of the type $V(T) = (10mVK)\ T$, T in °K. Also an IPTAT with a temperature coefficient of 1μA/°K yields $I(T) = (1\mu A/°K)T$, T in °K.

However, since the output of PTAT sensors extrapolate to zero at 0°K, it then becomes very easy to calibrate them in degree Celsius by simple slope adjustment, which is usually done at 25°C.

For instance, when expressed in °C, the output of LM335 VPTAT becomes
$V(T) = 2.732V + (10mV/°C)T$, T in °C. Likewise the current of the AD590 becomes
$I(T) = 273.2\mu A + (1\mu A/°C)T$, T in °C. Thus if we require a centigrade output, we must offset $V(T)$ by -2.732V and $I(T)$ by -273.2μA.

2.5 PRESSURE SENSORS

A Pressure sensor is an instrument component that detects a fluid pressure and produces an electrical, mechanical or pneumatic signal related to the pressure. Although pneumatic and mechanical transducers are commonly used, electrical measurement of pressure is often preferred because of a need for long distance transmission, higher accuracy requirements, more favourable cost and quicker response. For control applications, pneumatic pressure signal transmission may be desirable over electrical pressure sensor where flammable materials are present.

Electrical pressure sensors consist of an elastic element (in most cases a diaphragm), and an attached displacement sensor. When a force or pressure is applied to the elastic material, a small but reproducible strain or displacement of the diaphragm is sensed by an attached sensor (strain gage).

A strain gauge is a small resistor whose value changes when its length is changed. It may be made of thin wire, thin foil or semiconductor material

2.5.1 Diaphragm – Type Pressure Sensors

Most pressure sensors utilise a diaphragm as their elastic element, reason because they show linear and reproducible response and are relatively cheap to manufacture. Strain gauges are the most common means of measuring diaphragm deflections. In most of the arrangements, four gauges are employed in a bridge configuration to increase output sensitivity and, especially to reduce the temperature coefficients.

To use these sensors, bridge balancing is often required because the diaphragm fabrication and gauge mounting are difficult to control precisely. The balancing is done as shown in figure 2.1 (Franco, 1988).

In the absence of strain, each tap should be at $V_{ref}/2$. In practice, there will be deviation due to the initial tolerances of the four gauges. By varying R_2's wiper, an adjustable amount of current is passed through R_4. This will increase or decrease the voltage at the left tap (V1) until it becomes equal to that at the other (V2), thus balancing the bridge in terms of voltages.

2.5.2 Silicon Pressure Sensors

Solid-state pressure sensors employ a semiconductor both as a resistive strain gauge and as the elastic diaphragm. The gauge factor /sensitivity of the semiconductor-type strain gauges is much higher than that of metal gauges because the resistivity of the gauge is changed by any applied strain (not just the dimension alone). Temperature coefficient is also very high and temperature compensation is a major design problem

In silicon pressure sensors, sources of error include

- The gauges are not perfectly matched, thereby casing a zero shift
- Their gauge factor/sensitivity is temperature dependent

While constant current excitation is sometimes used to reduce the temperature coefficient of the full-scale output (by a factor of 2), better results are achieved by compensating the bridge voltage or amplifier gain. Zero shift compensation can be achieved by shunting the bridge element (which changes with temperature) by a high value trim resistor (which does not change with temperature).

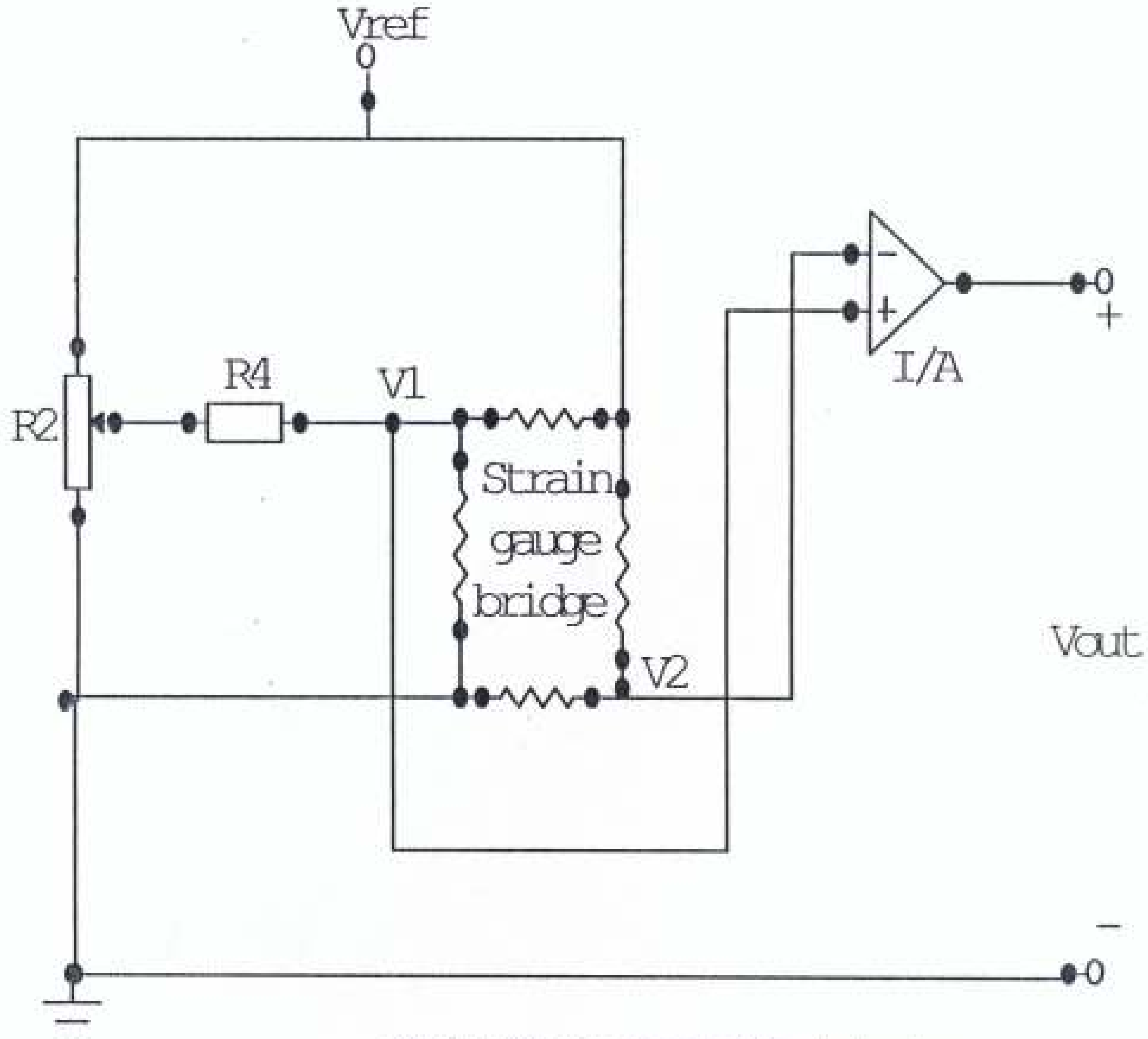


Fig 2.1: Strain gauge bridge balancing

2.6 HUMIDITY SENSORS

Humidity sensors measure the amount of water vapour/moisture content of air at any particular temperature and time. Humidity sensors are usually of two types. The two categories rely on the variation of some electrical property (e.g. resistance or capacitance) when water is absorbed on to their surfaces.

The first category of humidity sensor depends on the increase in conductance of a hygroscopic film. As the humidity increases, the absorbed water increases and thus the resistance between the electrodes decreases.

Older types utilize an ionic hydrate salt such as LiCl, but newer types are based on a hygroscopic polymer such as sulphonated polystyrene or a ceramic such as zirconium oxide. Generally the conductivity of these types of humidity sensor increases approximately exponentially with relative humidity (Darold, 1987). Hysteresis (different readings for rising and falling humidity) is rather high (2 to 10 percent). Temperature coefficients are also high (5 percent per degree Celsius) and temperature compensation is necessary. A further complication is that ac excitation is required to avoid electrode polarisation (a problem that is common with all ionic conductors).

Humidity sensors based on the capacitance increase that occurs when water is absorbed on gold surfaces or polymer show a reproducible response, but they are costly and not widely available.

2.7. SENSOR AMPLIFIER CIRCUITS

The data generated by sensors and some other instruments often require some level of processing and/or conditioning before they can give any meaningful information to the user.

Usually, since the electrical signals produced by most sensors are of very low voltage (of the order of a few microvolts), there is the need to amplify such signals. This will make them suitable for further processing, transmission and use.

A very important type of integrated circuit that is widely used in signal amplification is the operational amplifier. It has a differential input and a single ended output. Under ideal conditions, its output is always centered on zero volts. Since op-amp is extensively used in this work, it then becomes necessary to get us accustomed to the basic operation of the device.

Operational amplifiers are said to be ideal in their operations when they satisfy the following conditions:

1. Infinite open loop gain A_o , i.e $V_{out} = A_o(V_p - V_n)$ with $A_o = \infty$

2. Infinite input resistance R_{in}, so that almost any signal source can drive it without any overloading of the preceding stage.
3. Zero output resistance Ro, so that output can drive an infinite number of other devices
4. Output voltage is zero when both V_p and V_n are zero or have the same signal on them.
5. Infinite bandwidth –Any frequency can be amplified without any attenuation
6. Infinite common-mode rejection ratio – Any input, simultaneously applied to V_p and V_n will be rejected.
7. Infinite slew rate – Any change in output voltage occurs simultaneously with input voltage change.

However, under practical applications, op-amps deviate from the ideal conditions. Some of those non-idealities are discussed below.

2.7.1 Precision Design and Low Noise Techniques

An ideal op-amp would have an output of 0volt when its differential input voltage was zero. Real op-amps are likely to have a d.c output voltage even when the input voltage is zero. This is because it is impossible to make a differential amplifier with perfect symmetry. Because of this imperfection, there will always be a d.c voltage at the output of an Operational amplifier, when both inputs are grounded. This voltage is called an ***output offset voltage.*** Such an output voltage is an error voltage, which is undesirable in many applications.

In an op-amp, output offset voltage is caused by three main sources, namely (Odo, 2000);

1. Op-amp output offset voltage due to input circuitry
2. Op-amp output offset voltage due to the output circuitry
3. Output offset error due to the external components used to bias the op-amp.

Output offset voltage due to input circuitry

The input stage of most analogue amplifiers is usually a symmetrical arrangement of transistors. These transistors need a small input current to "bias" them for correct operation. If the transistor at each input is a bipolar transistor, this bias current must flow into each of the two input terminals.

However, since the two transistors at the input terminals of an op-amp cannot be perfectly matched, the currents that flow into the input terminals of the transistors are therefore not exactly equal. These unequal bias currents give rise to an offset in the output, and input offset current I_{IO} at the input.

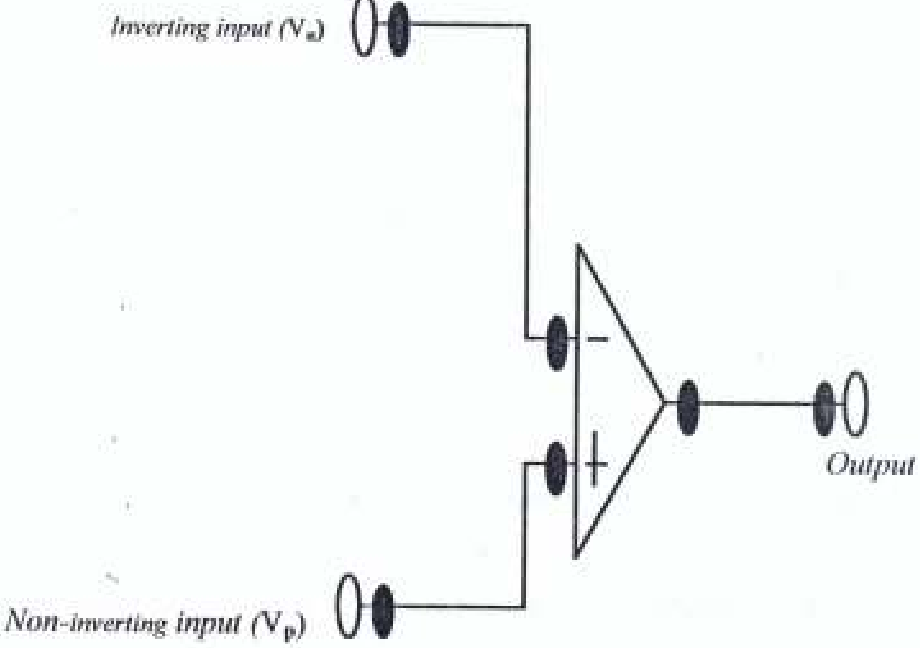


Fig 2.2: General symbolic representation of an op-amp.

A straightforward way of assessing the effect of nonzero input current upon circuit performance is to determine the output with zero input signals. We shall do this for the inverting and non-inverting amplifiers. As far as d.c conditions are concerned, fig 2.3 represents both non-inverting and inverting circuit configurations for an operational amplifier with feed back (Crecraft et.al,1993). R_3 represents the equivalent resistance of the bias current path from the non-inverting input to the 0v rail. In the non-inverting case, the signal is fed in here, so this bias path will include the source resistance if the signal is directly coupled.

Both R_F and R_1 provide paths for the bias current to the inverting input. In the inverting case, the signal is fed in between R_1 and 0V, and R_1 includes the source resistance if the signal is directly coupled. If the path from the non-inverting input to the 0V rail includes a series capacitor (in either inverting or non-inverting cases) then R_1 is infinite as a bias current path, and all the current must flow through R_F.

Because of the unavoidable mismatch at the input circuitry of an op-amp, I_p and I_n differ slightly. The input bias current I_B is defined as the average of the two:

$$I_B = (I_p + In)/2$$

While their difference is called input offset current (I_{IO}):

$$I_{IO} = I_n - I_p$$

The polarity of I_{IO} depends on the direction of mismatch of the input transistors. Usually I_B is about 5 to 10 times greater than input offset current I_{IO}.

Effect of input currents i_b and i_{io}

To evaluate the offset in output voltage caused by the input bias currents (I_B and I_{IO}), we make use of figure 2.3.

At node X,

$$I_2 = I_1 + (I_B + I_{IO}/2)$$

Substituting for I_1 and I_2

$$(V_0 - V_n)/R_F = V_n/R_1 + (I_B + I_{IO}/2)$$

Therefore,

$$V_0 = V_n(1 + R_F/R_1) + (I_B + I_{IO}/2)R_F \qquad (2.3)$$

Because of the very high d.c open-loop gain of the op-amp, $V_n \approx V_p$. Thus

$$V_n \approx V_p = -(I_B - I_{IO}/2)R_3$$

So substituting V_n in equation (2.3)

$$V_0 = -(I_B - I_{IO}/2)R_3(1 + R_F/R_1) + (I_B + I_{IO}/2)R_F$$

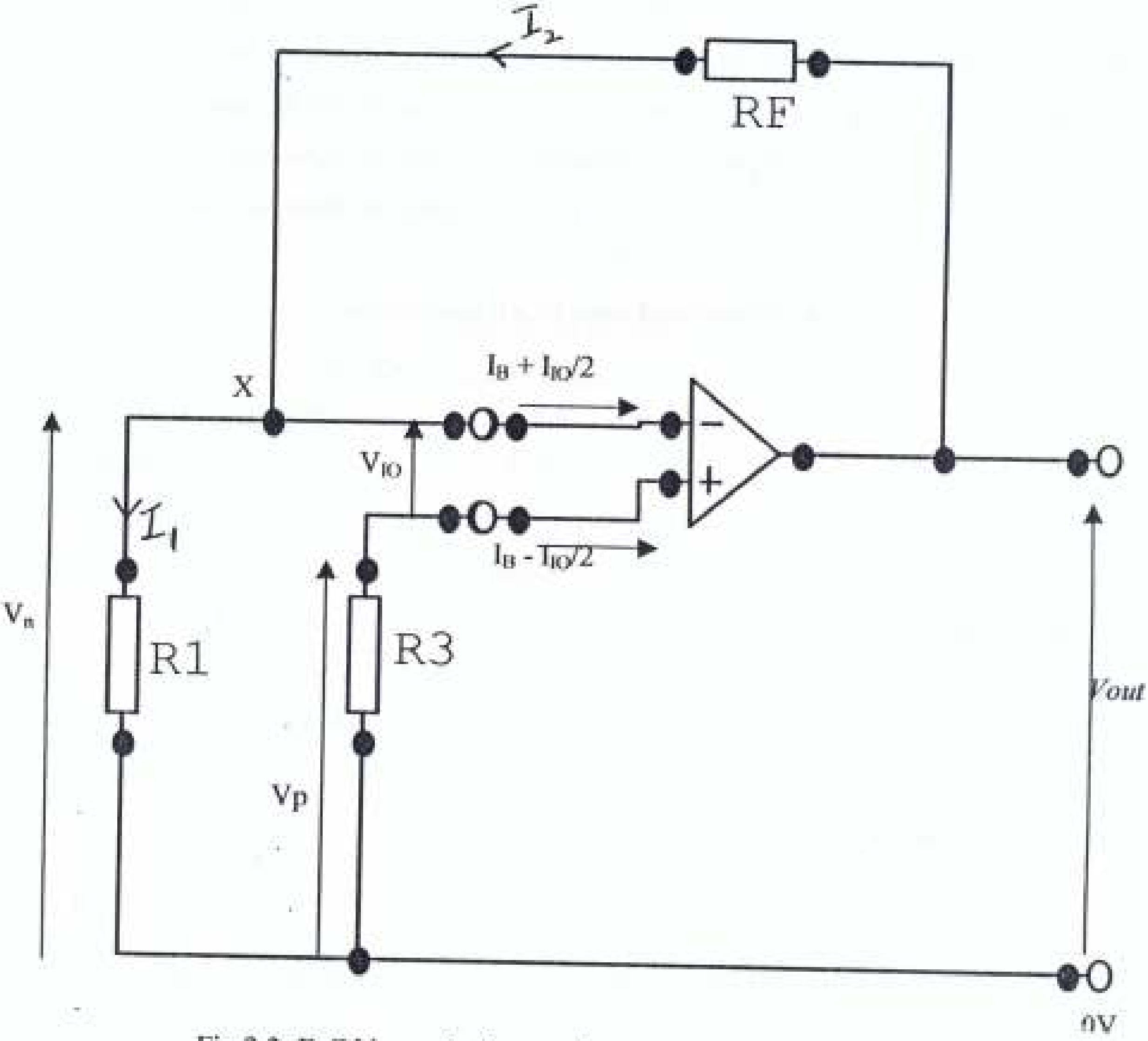


Fig 2.3: D.C bias paths for a real op-amp.

Collecting terms gives:

$$V_0 = I_B[R_F - R_3(1 + R_F/ R_1)] + I_{IO}\,[R_F + R_3\,(1 + R_F/ R_1)]/2 \qquad (2.4)$$

To make equation (2.4) independent of I_B, we make

$R_3(1 + R_F/ R_1) = R_F$. That is, if

$$R_3 = \frac{R_F}{(1 + R_F/ R_1)} = \frac{R_F\,R_1}{R_F + R_1} = R_1//R_F$$

When this is done, then $V_0 = I_{IO}R_F$ due to input currents alone.

Many technologies are used in fabricating the differential input stage of an op-amp. Examples are bipolar input stage, JFET input stage, MOSFET input stage. Op-amps are also available entirely in CMOS technology. In all these technologies, manufacturers have devised several ways to keep I_B and I_{IO} as small as possible.

Table 2.1 is a list of common op-amps and their I_B and I_{IO} characteristics.

Op-amp	Technology	I_B	I_{IO}	Manufacturer
LM308 & 312	BJT	1.5nA	0.2nA	National Semiconductor
LM312, OP-08	BJT	1nA	0.08nA	Precision Monolithic
LM 312 OP-08	BJT	50pA	10pA	National Semiconductor
LF385/6/7	JFET	30pA	3pA	National semiconductor
TL070/080	BiFET	30pA	5pA	Texas Instrument
AD549	JFET	≤100fA		Analogue Devices.
CA3130	BiMOS	2pA	0.1pA	Raytheon
CA080/1/2/3/4	BiMOS	15pA	5pA	-
ICL761x/2x/3x	CMOS	1pA	0.5pA	Intersil
TLC271/2/4/7	CMOS	1pA	1pA	Texas instrument

From the above data, we see that op-amps made of FET and CMOS technologies have lower I_B and I_{IO} than BJT input op-amps.

Input bias current variation with temperature

Fig 2.4 compares typical bias current and bias current variation with temperature for various technologies. It is noted that the FET input op-amps as compared with BJT-input types, namely, an approximately exponential increase of gate current with temperature.

A well-known rule of thumb states that gate current doubles with about every 10°C of temperature increase. Thus the advantage that FET-input op-amps hold over their bipolar counterparts at room temperature (the condition for which the ratings in fig:2.4 are given) disappear at sufficiently high temperatures.

Knowledge of the intended temperature range of operation is an important factor when selecting the optimal device.

2.7.2 Effect Of Input Offset Voltage, V_{io}

Input offset voltage V_{IO} is the voltage, which must be applied across the input terminals of an op-amp in order to bring its output voltage to zero. It is caused by manufacturing *tolerances. There exists an output offset voltage V_{OO} as a result of V_{IO}. To nullify this* error voltage, we need to have a current at the input of the op-amp that will give the flexibility of obtaining V_{IO} of proper amplitude and polarity. Any circuit that enables us to achieve this is known as Input-offset voltage compensation network.

As is often quoted in many electronics text, the error caused by V_{IO} is given as:

$$V_{OO} = (1 + R_F/R_1)\, V_{IO}$$

Manufacturer data sheets provide typical as well as maximum values of V_{IO} at room temperature. For the 741C, V_{IO} = 2mV (typ) and 6mV (max) ; for the 741E, V_{IO} = 0.8mV (typ) and 3mV(max). These figures are typical of many low cost op-amps, although types are available with much lower offset ratings.

Changes in v_{io} due to temperature drift

The V_{IO} ratings are usually given at room temperature, and, like other parameters of the op-amp, V_{IO} is temperature dependent. The temperature coefficient of V_{IO} is the change in V_{IO} brought about by a 1°C temperature change and is represented by $\frac{\Delta V_{IO}}{\Delta T}$.

LM 301A is rated at$\Delta V_{IO}/\Delta T$ = 6µV/°C (typ) and 30µV/°C (max). These ratings are typical of the BJT family. There are some op-amp types which are specifically designed for low input offset voltages and low temperature coefficient of V_{IO}. As an example, the OP-07 family of ultra low-offset –voltage op-amp comes in versions with V_{IO} = 10µV and $\Delta V_{IO}/\Delta T$ = 0.2µV/°C.

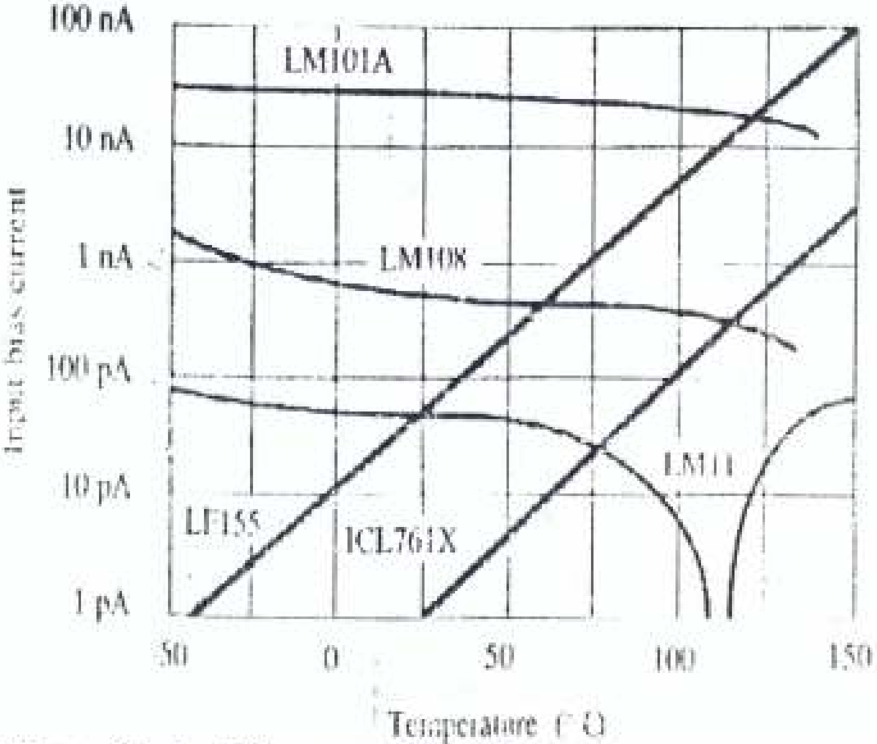


Fig 2.4: Comparison of typical bias currents for various op-amp types (Franco, 1988): LM101A (conventional bipolar), LM108 (bipolar superbeta), LM11 (bipolar darlington superbeta), LF155 (biFET), and ICL761X (CMOS).

On the basis of the average temperature coefficient one can estimate the value of V_{IO} at a temperature other than 25°C as;

$$V_{IO}(T) \approx V_{IO}(25^\circ C) + \Delta V_{IO}/\Delta T(T-25^\circ C)$$

FET input op-amps traditionally have been plagued by poorer input offset voltage and offset drift characteristics than their bipolar counterparts. However with advances in design and fabrication, this difference has become less pronounced.

2.7.3 Input Guarding

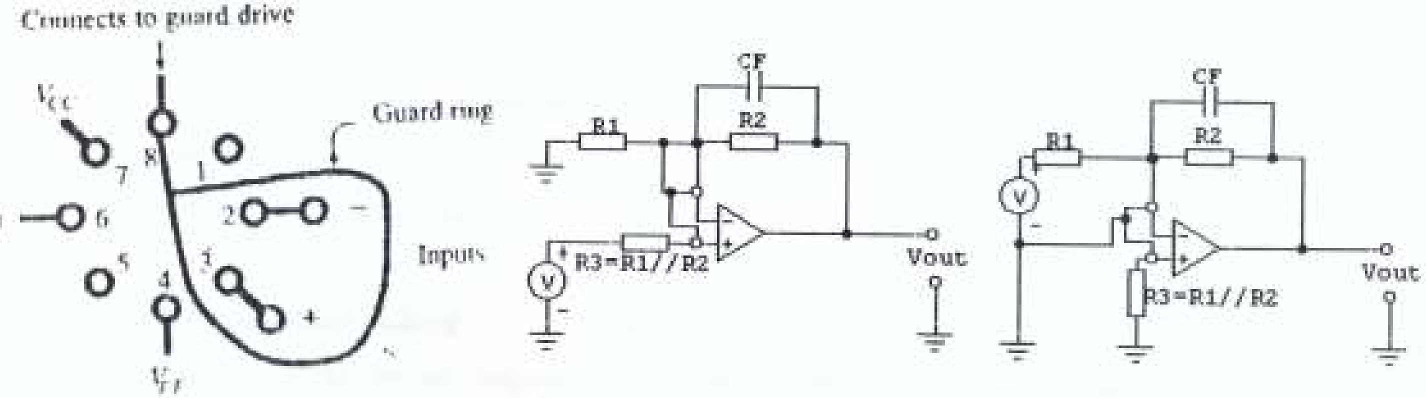
When an op-amp is wired on a printed circuit board, there is every possibility for current to leak across the circuit board. Therefore, an op-amp with very low input bias currents must be properly guarded to protect it from leakage currents. If this is not properly done, the leakage current can easily exceed the input bias current itself, thereby degrading circuit performance.

Leakage currents can be reduced in two ways, namely;

- Mounting the Op- amp on a Teflon IC socket
- Putting the input terminals in a guard ring if the op-amp must be soldered directly on the board (see fig 2.5).

A ***guard*** consists of a conductive ring surrounding the input terminals and driven at the same potential. The guard ring will absorb leakages from other points on the board and thus prevent leakage to the op-amp inputs. Secondly, the guard ring acts as a shield against noise pickup and reduces the common-mode input capacitance seen by the inputs (Franco, 1988).

To combat the stray capacitance of the ring itself, a small compensating capacitance (of a few pF) is connected between the output and the inverting input. When a sensor and its amplifier are far apart, a coaxially shielded cable with good insulation must be used in addition to the guard ring. The cable shield must be kept at the same potential as the guard ring.



(a) (b) (c)

Fig 2.5: Input guard layout and guard connections for the non-inverting (b), and inverting (c) configurations

2.8 OFFSET ERROR COMPENSATION

Manufacturers of precision op-amp employ various techniques to minimise sources of error in their products. One of such means is a provision made to compensate for all sources of error from I_B and V_{IO}.

Offset error compensation, also referred to as *offset nulling*, *offset trimming* and *offset balancing*, can be achieved in more than one way depending on the op-amp type as well as the configuration used.

Offset nulling techniques are classified as ***internal*** and ***external***.

2.8.1 Internal offset nulling

Fig 2.6 is the circuit diagram for internal offset nulling. The nulling procedure is straightforward. The potentiometer R_V is adjusted to drive V_O to zero. However, it should be noted that for internal offset nulling to be carried out, the op-amp must have provision for it. For example the popular 741op-amp has this facility accessible on its pins 1&5.

When a precision circuit involves more than one op-amp, experience has shown that introducing offset nulling into just one of the devices can null the overall error. However, for this technique to succeed, the offset voltage adjustment range must be wide enough to span the combined effect of all the other offsets. This is because the overall error is the result of the algebraic sum of all the individual offset errors. It is worth nothing that not all op-amps have provision for internal offset nulling. Even for those with this provision, the nulling scheme varies among types. For example the LF356 requires a 25KΩ pot with the wiper connected to Vcc. It is always advisable to consult the data sheets to find the recommended scheme for any device in use.

Though internal offset nulling is quite easy to implement, it affects other characteristics of the op-amp. For example in LM321 & OP-10, it reduces thermal drift. While in some other types, it degrades the thermal drift, the CMRR and the PSRR. Unfortunately, data sheets rarely provide adequate information in this regard. When in doubt, external offset nulling (which is being discussed next) offers a more predictable alternative.

2.8.2 External offset nulling

External nulling (shown in Fig 2.7) is based on the injection of an adjustable d.c voltage into the circuit to compensate for the existing offset error. External nulling does introduce additional imbalance in the input stage; hence there is no degradation in temperature drift, CMRR and PSRR.

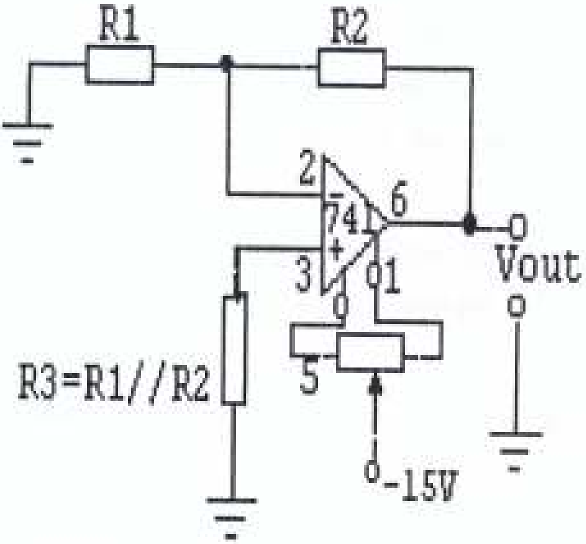


Fig 2.6: Resistor feedback configuration with internal offset nulling

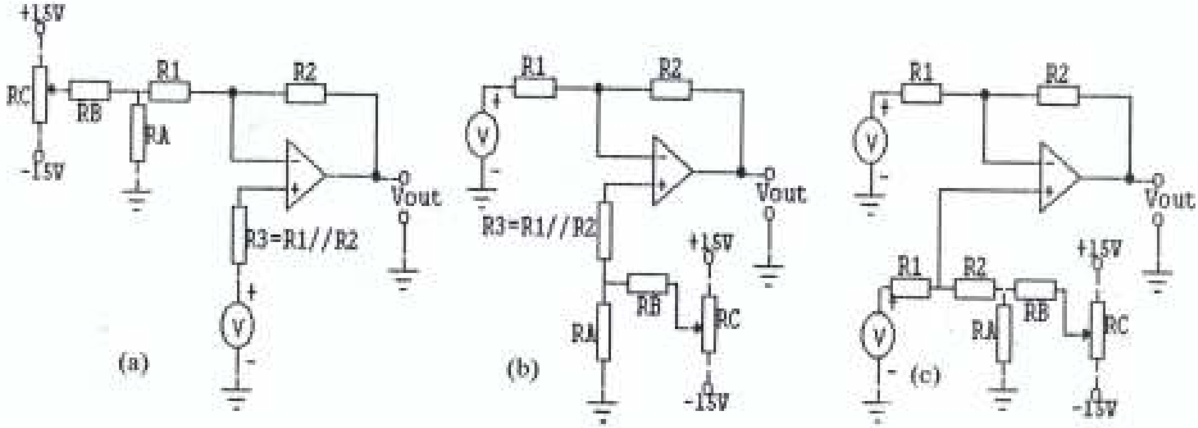


Fig 2.7: External offset nulling for (a) non-inverting amplifier, (b)inverting amplifier and (c)difference amplifier

2.8.3 Auto zero techniques

Offset nulling, whether internal or external, is done at one specific set of temperature, common-mode, and supply voltage conditions. As these conditions change so does the offset error. Manufacturers of precision op-amps have developed special types of op-amp that are capable of continuous offset correction, irrespective of the temperature and supply voltage condition.

These special types of op-amp are known as *chopper-stabilized op-amp*.

The main properties of the autozero chopper-stabilized op-amp are

1. They are built on CMOS technology.
2. Their V_{IO} and temperature coefficients are five times better than the best precision bipolar op-amp.
3. They deliver their full speed and bandwidth.
4. Being a CMOS device, most of them have a severely limited supply voltage (typically 15V total supply). Because of this, it cannot run from conventional ±15Vsupply. The only exceptions to this are the Maxim MAX430/2 and Teledyne TSC915 and TSC76HV52 which operate from ±15V supplies.
5. Many autozero op-amps have restricted common-mode input voltage. For example, the popular ICL7650 has a guaranteed common-mode input range of -5V to 1.5V when running from its usual ±5V supply. For the improved ICL7652, the range is from -4.3V to 3.5V; that is a wider range, but it does not include the negative rail, so it cannot be used as a single supply op-amp.

   The high voltage types are much better, for example, the MAX432 has a guaranteed common-mode range of -15V to 2V when running from ±15V supplies.
6. They have poor output current sourcing capability – sometimes as little as 1 - 2mA in the sourcing (positive output) direction.
7. They have disastrous saturation characteristics. Recovery is very low – up to a second. The "cure" is to sense when the output is approaching saturation, and clamp the input to prevent it. Most auto-zeroing op-amps provide a "clamp" output for this purpose, which can be tied back to the inverting input to prevent saturation.

   Saturation can be prevented in chopper op-amp without a "clamp" pin (and in ordinary op-amps as well) by bridging the feedback network with a bi-directional zener (two zeners in series) (Horrowitz et.al, 1995). This will clamp the output at the zener voltage, rather than letting it limit at the supply rail; this works best in the inverting configuration.

- Chopper stabilized op-amps are plagued with the problem of clock induced noise. It is caused by charge coupling from the MOS switches and can cause spikes at the output. In low frequency applications, we can (and should) RC – filter the output to a bandwidth of a few hundred hertz, which will make these spikes disappear (Horrowitz et.al, 1995). This spiky noise is of no importance in integrating applications (e.g. integrating A – D converters) or in applications where the output is low (e.g., a thermocouple circuit with a meter at the output). Infact, if we only want very low frequency (below 1Hz) output response, a chopper op-amp will actually have less noise than a conventional low noise op-amp (Horowitz, 1995).

2.9 OUTPUT OFFSET ERROR DUE TO OP-AMP OUTPUT CIRCUITRY

There are limits to the amount of voltage and current that an op-amp can deliver to a load, and there are limits as to how quickly this voltage can change. These limitations become sources of error when using op-amps in ac circuits.

Though most op-amp have over-current/short-circuit protection, it is a good design practise to always ensure that op-amp circuits are terminated with high impedance device so as to prevent excess current been drawn from the device (Odo, 2000). Otherwise, if the op-amp must drive a low impedance load, then, a unity buffer must come between the amplifier and the load.

2.10 ERROR DUE TO THE EXTERNAL NETWORK COMPONENT

Op-amps, like some other monolithic IC's needs to be biased alongside with some resistors, inductors, diodes and capacitors for proper operating condition. The degree of precision of circuits realised from such arrangements depends largely on the accuracy and thermal stability of the op-amp and indeed the entire discrete components used.

For example, the common-mode-rejection ratio of a differential amplifier is known to be greatly affected when the ratio of the two pairs of resistor used at the differential input do not match. In the same vane, the accuracy and linearity of the ramping action of integrators is also largely dependent on the properties and stability of the capacitors used in the circuit.

Following the above facts, it can be seen that the choice of component is vital in any circuit implementation.

Components are generally specified with an initial accuracy, as well as the change in value with time and temperature. Complete specifications also include the effects of temperature, cycling and soldering, shock and vibration, short-term overload and effect of moisture.

Table 2.2 gives the specification for the commonly used two resistor types. From this table, it is obvious that metal film resistors are better than carbon composition resistors for precision work.

Table 2.2: Showing variation of Resistor properties (Horowitz, 1995 pg. 1055).

Material	Temperature coefficient (tempco)	Soldering temperature/load cycle	Shock and vibration	Moisture
Metal film Resistors	50ppm/°C	-55°C to 175°C	0.1%	0.5%
5% Carbon composition Resistor	80ppm/°C	25°C to 85°C	2%	6%

2.11 DATA TRANSMISSION, MANIPULATION AND RECORDING

Very often, data needs to be transmitted to a place where it can be accessed or used. The main reason for data transmission in most cases is that the source of the data and the location where it is needed may be scattered or the source is remotely located..

In electronics, data is collected and transmitted in the form of an electrical signal. This electrical signal may come from sensors.

Basically there are three principal ways of getting electrical signal from one place to another. These are through:

- Wires (usually copper)
- Radio waves in free space
- Guided waves in wave guides, coaxial cables and optical fibres

2.11.1 Wire Transmission

Wire transmission is perhaps the most common way by which analogue and digital signals are being transmitted from one location to another. There is always a delay associated with the transmission of a signal through any medium. This delay arises because electromagnetic waves have limited maximum velocity at which they can be transmitted in a medium.

This inherent property of the medium cannot of course be eliminated. But a good knowledge of wire interconnections will save us from problems that could result from bad choice of wire network, which may lead to receiving distorted and attenuated signal, or any signal that has been corrupted by noise/interference.

Also a good knowledge of wire interconnectivity will help us to arrive at various cost effective ways of reducing these undesirable properties.

A transmission line simply comprises ***go*** and ***return*** wires in close and constant proximity to one another. Transmission lines can be classified into two main classes, namely; short transmission lines and long transmission lines.

Connections within local electronic circuits or between equipments are considered to fall under the short transmission line, while any transmission line longer than 1km is considered to be long. Short transmission lines are extensively used in this project. Therefore, further discussion will only be on short transmission lines.

The principal properties of short transmission lines are:

- The signal is transmitted with little loss of amplitude and without distortion
- There is a small propagation delay in the cable, depending on the length and characteristic of the line
- As viewed from the signal source, a transmission line has a characteristic impedance Z_0

There are two things that we must know about Z_0:

1. Z_0 depends on the dimension and construction of the line.
2. Z_0 is normally non-reactive, but a pure resistance.

Surprisingly Z_0 does not dissipate power in the way that a resistor does.

When the load impedance is equal to the characteristic impedance of the line, the two are said to be ***matched*** or the line is said to be "*properly terminated*". When this happened, source power will get to the load without any loss in the transmission line except for a little delay. Consequently, maximum power transfer is said to occur at impedance matching. If the load is not properly matched to the line, or the line is not properly terminated, the signal is partly reflected back down the line. To be absorbed in the internal resistance of the source, provided the source is matched to the line. If neither source nor load is matched to the line, the signal is reflected back and forth along the line. If the load is short-circuited or if it is an open circuit, the signal is wholly reflected, and the load of course receives none.

Interconnection methods

1. ***Open Wires***

A simple piece of wire is, of course, the basic way to connect two points together. It is perfectly adequate, provided the amplitude of the signal it carries are not too small and the wire is not too long, and there is also a return 'earth' wire. The length is important because open wires have series inductance, as well as shunt capacitance to other parts of the circuit and to earth. Together these act like a low pass-filter and attenuate high frequency signals. The longer the wire, the more the inductance and capacitance, and the more the attenuation of high frequency signals. In transmission lines, the inductance and capacitance between the 'go and return' wires are controlled and do not result in high frequency attenuation. Remarkably, ideal lines, with no ohmic resistance, do not attenuate signals of any frequency; they only delay them slightly.

In addition, open wires always pick up interference. The longer the wire, the more the interference; the smaller the signal, the more serious is the effect of the interference.

2. *Twisted pairs*

Twisting a pair of insulated wires together, as shown in fig 2.8 is the simplest form of transmission line. It provides a convenient way of carrying power to electrical equipment or for carrying signals between pieces of equipment. It is used widely for frequencies up to a few megahertz. Examples are digital systems, instrumentation systems and telephone systems.

As compared with a single wire, a twisted pair picks up much less interference, since the interference waveforms picked up in the two wires tend to cancel each other out. Typical characteristic impedances are 50Ω, 70Ω, 85Ω, and 110Ω, and the propagation velocities range from 130 to $200*10^6 ms^{-1}$.

3. *Twin feeders*

A typical form of twin feeder is shown in fig 2.9. The two wires have a greater separation than the twisted pair. Moulding the two wires into a strip of plastic makes the flexible twin feeder. A popular version has a separation of about 10mm, and is used for connection to VHF radio aerials. It has characteristic impedance Z_o of 300Ω, and a propagation velocity of $250 *10^6 ms^{-1}$. It is much cheaper than the coaxial cable, but is more susceptible to interference pick-up.

Twin feeders for connecting powerful radio transmitters to their aerials use rigid conductors and a series of insulating separators, so they are mainly air-spaced and have a propagation velocity of $300 *10^6 ms^{-1}$. They are commonly used at frequencies up to 30MHz.

Fig. 2.8: A twisted pair

Fig. 2.9: A flexible twin feeder

4. *Coaxial cables*

A coaxial cable consists of an inner conductor surrounded by a layer of insulation, and an outer conductor in the form of a cylindrical mesh, as shown in fig 2.10. The outer conductor is intended to screen the inner conductor from external interfering electric fields.

Coaxial cables – or *screen cables* as they are sometimes called – are widely used for signals at all frequencies, from the lowest-frequency instrumentation signals up to the highest microwave frequencies. Popular characteristic impedances are 50Ω and 75Ω. Professional radio frequency (RF) cables that contain polyethylene as insulator have propagation velocities of $200*10^6 ms^{-1}$. 75Ω cables for UHF television aerial down leads are part-polyethylene, part air-spaced, for low loss. Their propagation velocity is about $240*10^6 ms^{-1}$.

5. *Multi-way cables*

Fig 2.11 shows examples of 'multicore' cables. These use all of the types of interconnection that we have described so far, and combination of them.

The simple 'bundle of wires' is adequate for low-frequencies and large signals. At higher frequencies, interference between the wires, called '***crosstalk***', tends to become significant. If the wires are formed into twisted pairs the interference pick-up and cross-talk are reduced. An overall screen always helps to reduce pick-up. Individual screens around each pair help reduce cross-talk. A bundle of coaxial cables provides the best performance of all.

Fig 2.12 shows a **ribbon cable.** This is very cheap to produce, and provides a neat multiway interconnection method. It is especially useful in digital equipment for making parallel data connection. Plugs and sockets for ribbon cables have the so-called insulation-displacement connectors (IDC), so that all the wires in the cable can be joined simultaneously to the plug or socket terminals. For ribbon cable with PVC insulation, the characteristic impedance between adjacent conductors is about 105Ω and the velocity of propagation about $200*10^6 ms^{-1}$. With PTFE insulation, Zo is about 125Ω and the velocity about $125*10^6 ms^{-1}$ Ribbon cables are adequate for short-distance connections. At lengths over about 1meter crosstalk in digital systems can cause error in data transfer. The hybrid type of the cable, also shown in figure 2.12 combines the best of both worlds: the easy connection to plugs and sockets of the ribbon cable, and the better crosstalk rejection of twisted pairs. The twisted–pair sections are about 500mm long, separated by flat sections about 50mm long, suitable for IDC connectors. The characteristic impedance of each twisted pair is about the same as that in the flat cable, and the velocity of propagation is about the same too.

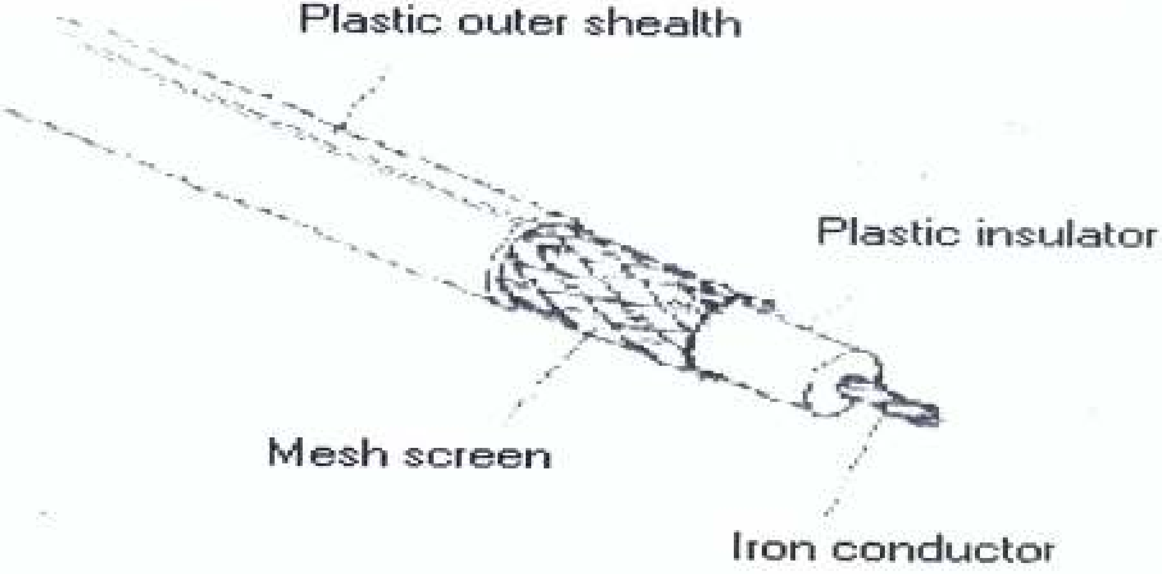
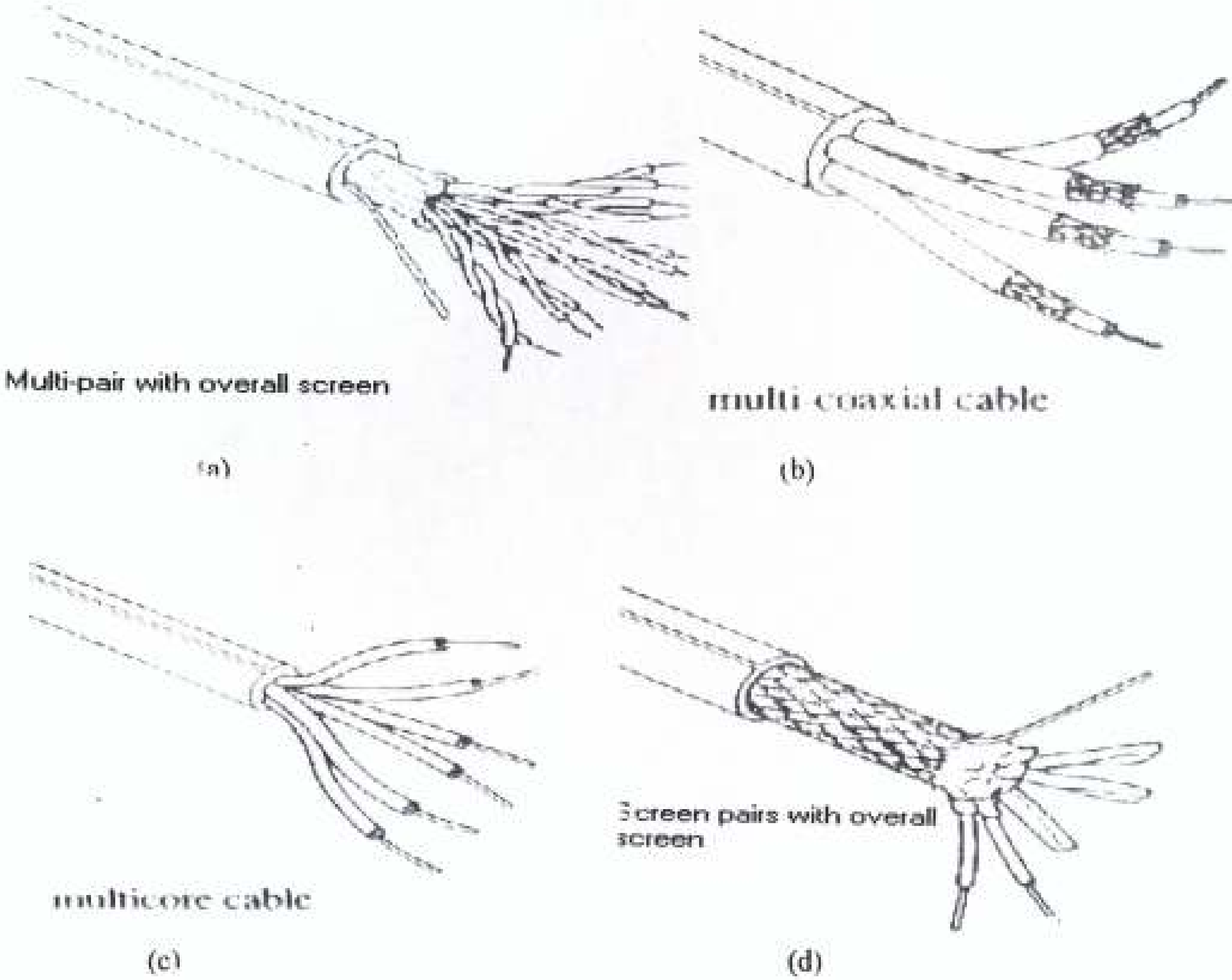


Fig. 2.10: A coaxial cable


Multi-pair with overall screen

(a)

multi coaxial cable

(b)

multicore cable

(c)

Screen pairs with overall screen

(d)

Fig. 2.11: Multi-way cables

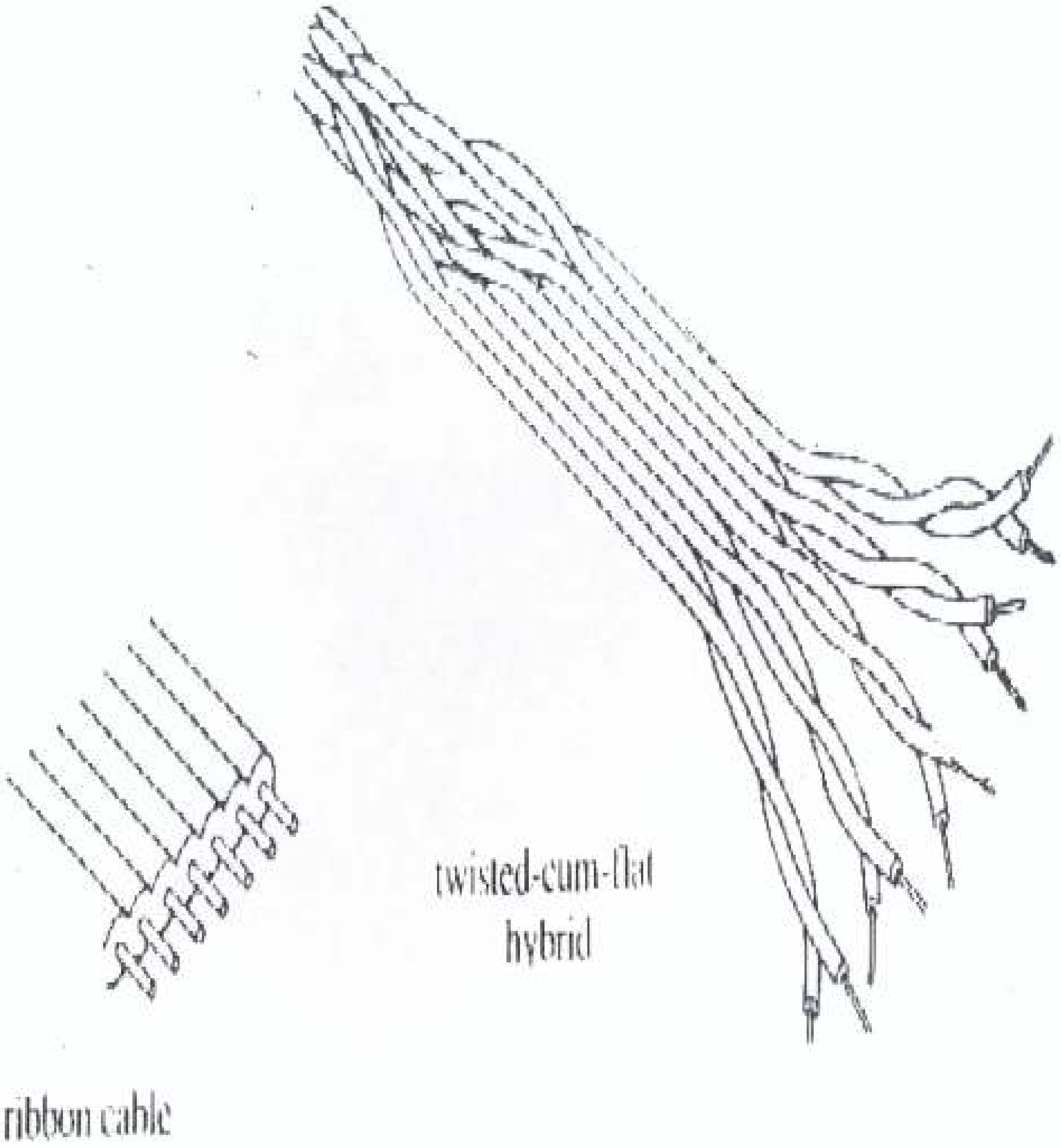


Fig. 2.12: Ribbon cable

Current-loop transmitters

Apart from the various wire interconnection techniques, current loop transmitter is yet another popular device employed for data transmission in the process industry. This device is a voltage-to-current converter. It converts sensors output voltage to a proportional current.

For the device, a zero input voltage produces a minimum current of 4mA and a full scale input produce 20mA, which is transmitted to the point of usage via copper wires. A major advantage with this type of device is that it is able to maintain a constant current over the entire transmission line for a given voltage input, thereby enhancing remote data acquisition, processing and control.

Another technique of getting uncorrupted data across transmission lines is to digitise the signal, frequency modulates it or frequency shift-keyed the signal on the line.

2.11.2 OPTICAL COUPLERS AND OPTICAL FIBRES

In an optical coupler, the input signal is converted to visible light or infrared, usually by a light-emitting diode (LED). The light is then detected by a photocell, or a photodiode or a phototransistor. The purpose of such a device is to provide electrical isolation between the input signal source and the output circuit. Electrical isolation may be needed for safety reasons if large voltage differences are involved. Data rates up to 15Mbits per second are possible. Devices with high isolation voltage rating are commonly called *opto-isolators*. They are available with ratings up to 10kV or so.

Optical fibres are used for transmitting signals in optical form. As in the case of the optical coupler, a Laser or an LED converts the input electrical signal to an optical signal. The optical signal is fed into a special glass fibre, which acts as optical wave guide. At the receiving end, the optical signal is detected by a photodiode.

Optical fibres were developed initially for telephone systems. Because they operate at optical frequencies, their bandwidth is far greater than that of coaxial cables, and their attenuation is lower too. For instrumentation systems and other narrow bandwidth systems, cheap emitters, receivers and fibres are available. Their advantages over coaxial cables are:

- Electrical isolation
- Freedom from capacitive and inductive pickup
- Freedom from cross talk
- Eliminations of sparkling and fire hazards
- No attenuation of signal strength

2.11.3 Data Transmission using Radio Waves

In many cases of data transmission, interconnection of wire from one point to another may become impossible. In such cases opto-telemetry or radio telemetry are often used.
The word **telemetry** simply means: measurement at a distance. In utilizing radio waves for data transmission, there is a considerable standard set to cope with the allocation of a frequency band within the radio frequency spectrum (Mie, 1997), (Odo, 2000).

The standard carrier frequency for telemetry is specified from the range 216 to 235MHz. The fig2.13 shows the widely utilised system of radio telemetry. The FM/FM refers to the fact that two frequency modulation processes are employed. In the first process, time varying d.c voltages are converted to proportional frequencies using a voltage-to-frequency converter. The standard FM/FM system allows for multiplexing of various inputs from different sensors and then transmitting the signals via a single carrier frequency as illustrated in the block diagram.
Digital form of radio telemetry is also in wide use. It is known as pulse code modulation, PCM.
Radio telemetry is very useful over short distances, especially when the relative motion of the measuring device and the readout equipment prevents a suitable direct connection. Examples of such situations are found in measurement on rotating machinery (e.g. a rotary furnace).

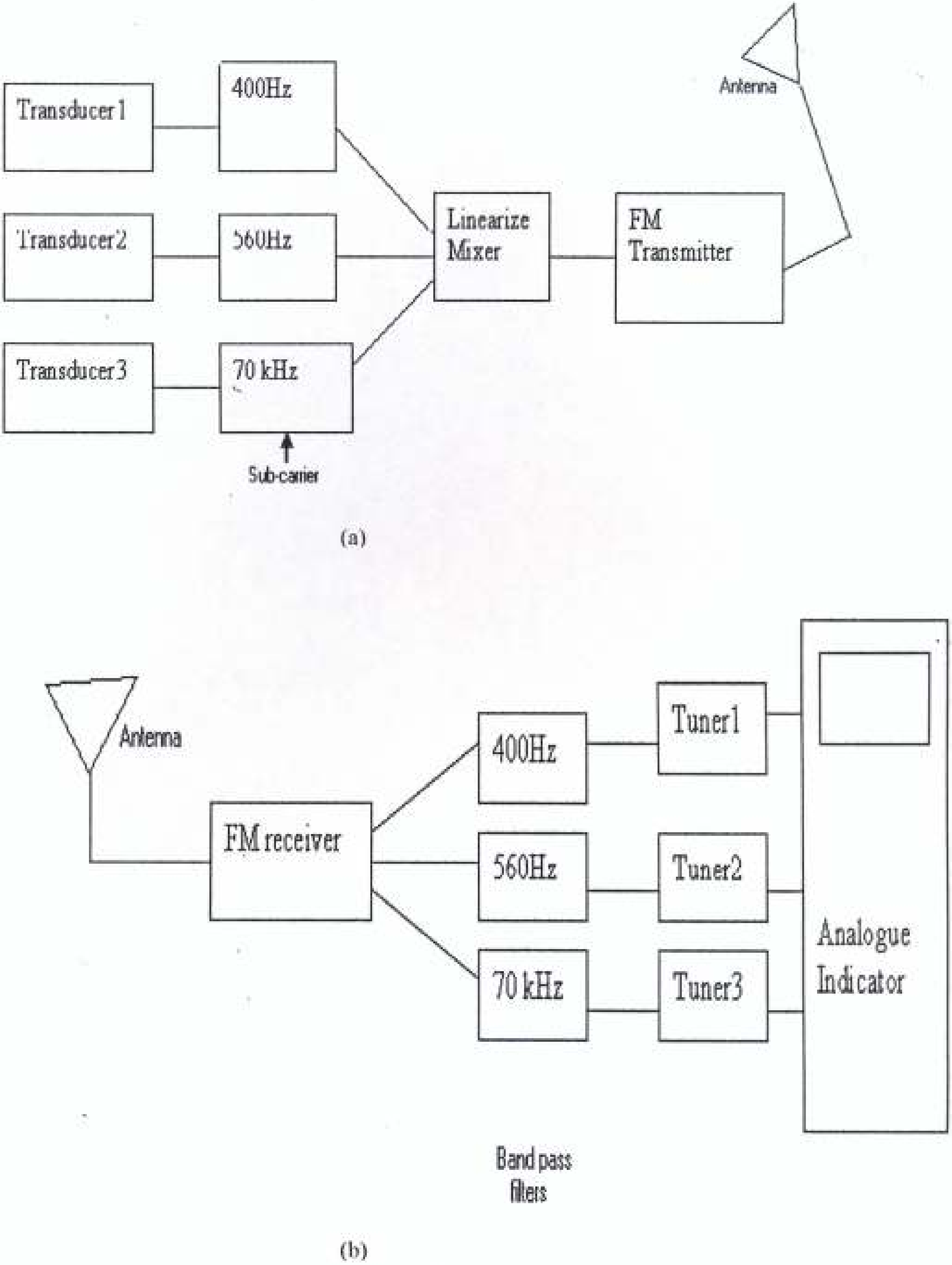


Fig 2.13: A typical FM/FM telemetry system used in multiplexed mode

2.12 ANALOGUE – DIGITAL CONVERSION

Information carrying variables such as currents, voltages and charges exist naturally in analogue form. But for effective processing, transmission and storage, it is often more convenient to express this variables in digital form. Also digital systems are increasingly becoming more popular due to their increasing efficiency, reliability and economical operating cost.

There are however many systems that incorporate both analogue and digital subsystems, particularly in the fields of measurement and control and in rapidly growing areas of communication technology. With the development of the microprocessor, data processing has become an integral part of many systems. Data processing involves the transfer of data *to* and *from* the microprocessor via input/output devices. Since a digital system uses a binary system of zeros and ones, the data input into the microcomputer have to be converted from analogue form to digital form. The device that performs this conversion is called an ***Analogue-to-Digital converter***. On the other hand, a ***Digital-to-Analogue converter*** is used whenever a binary output from a digital system must be converted to an analogue voltage or current.

Since an Analogue-to-Digital converter (ADC) is central to this work, detailed discussion on it is presented next

2.12.1 Analogue – to – Digital converters

This section examines only the techniques that are used in the realisation of Analogue-to-Digital converters.

1. Flash converters – The main components of this converter are comparators, a reference voltage for each of the discrete levels represented by the digital outputs. It also has an encoder.

Properties: Flash converters are the fastest of all the converters available. This is because all the n-bits are made available at the same time. However, they are expensive because the number of comparators needed is equal to the number of quantisation intervals. For example, 10bit converter contains 1023 comparators, together with control and encoding circuitry.

Flash converters resolution is limited to about 10bits due to their complex circuitry. This makes them unsuitable for high-resolution applications.

2 Counter-ramp converters – The main components of this converter are D-A converter, a binary counter, a clock, a single comparator and some control logic.

Properties: This converter is not as fast as the flash converter, but in its case, an n-bit device will require a conversion time that is proportional to 2^n. The major beauty of this device is that all the process of data conversion can be controlled. For example, when a conversion is required, a signal is sent to the converter from an external digital subsystem requesting a conversion. When the conversion is completed, an end of conversion signal is usually generated by the converter.

The time lapse between the start of conversion (**SOC**) and the generation of the end of conversion signal (**EOC**) is known as the ***conversion time*** of the converter. Both SOC and EOC can be software controlled. This makes the converter easily adaptable to any design that involves measurement and control.

The one major drawback of the counter-ramp A-D converter is its relatively long conversion time for large input signals (Gorham et al, 1993). Most commercial devices operate with maximum clock rates of 1MHz or less.

3. The Tracking converter – It has the same configuration as the counter-ramp converter (fig2.18). The only difference between them is that tracking converters are faster than the counter-ramp. This is so because tracking converters does not always reset the counter to zero. Instead, it only counts the difference between the present and previous samples.

4. The Successive approximation A-D converters – The major difference between this converter and the counter-ramp A-D converter is that a register has replaced the counter.

Properties: The total conversion time is equal to (n+1) clock cycles, one for each bit of the codeword and one to initialise the D-A converter output to zero. The converter is much faster than the counter-ramp type. An additional advantage is that the conversion time is independent of input amplitude, which is useful in systems that require a constant conversion rate.

5. Integrating (Dual-slope) A-D Converters – The main component of this type of converter are op-amp, with a resistor and a capacitor. All are arranged as an integrator.

Properties: Conversion time for the dual slope configuration is usually rather slower than the other methods of A-D conversion. This slow conversion time, is as a result of the long integration time. This longer integration time can be used to an advantage in that it can be used

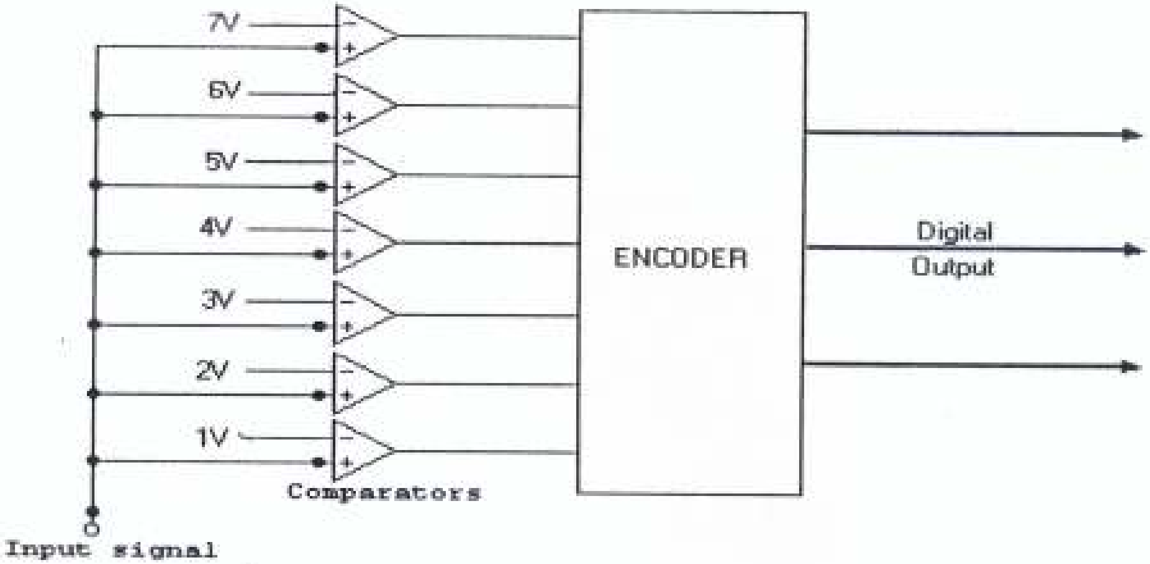


Fig 2.14: A Flash converter

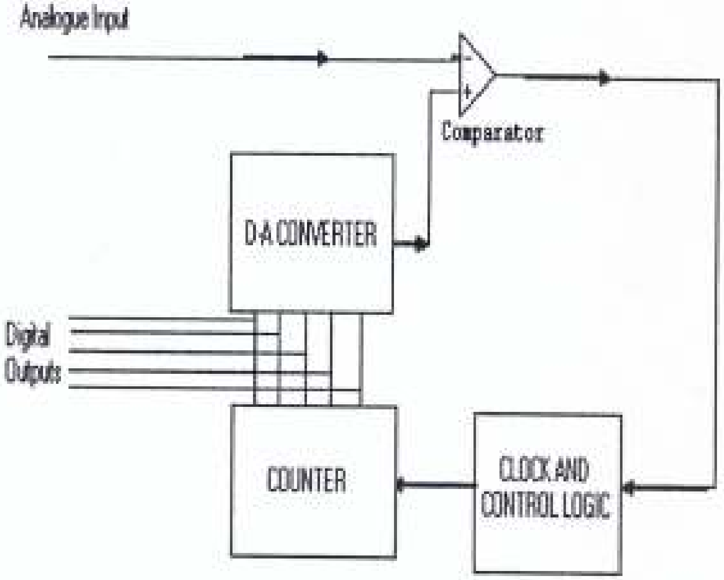


Fig 2.15 : Schematic diagram of a counter-ramp converter

to cancel out interference, by specifying the clock period so that 2^nT_{clk} is an exact integral multiple of any ac noise such as 60Hz power line pickup or 120Hz Bridge rectifier pickup.

The conversion accuracy of the dual slope ADC is independent of R, C, T_{clk} and the op-amp offsets. Dual-slope converters are particularly suited to highly accurate measurement of slowly varying signals.

Having observed the various techniques used in the realisation of an ADC and their respective merits and demerits, next is the presentation of the general specification that needs to be considered when selecting an ADC.

2.12.2 Analogue-to-Digital Converter Specifications

- *Saturation Error* – The most obvious limitation of an ADC is its defined upper and lower limits of voltage response. Typical full-scale ranges are 0 – 10V and –10V to +10V. If the input signal exceeds the upper limit of response, the converter saturates and the recorded signal does not vary with the input. Saturation can be prevented by appropriate signal conditioning such as amplitude attenuation or dc offset removal.
- *Resolution and Quantisation Error* – Each output code of an ADC corresponds to a whole range of input values. Fig 2.16 shows the normalised transfer characteristics of 3bits ADC. It is noted here that any input in the range between 1/8 and 2/8 yields the same code, namely 001. This inability of the converter to distinguish different levels within this band is known as **quantization error.**

Quantization error is an irreducible feature of the A-D conversion process. The output code can be in error by as much as ±1/2 LSB. Quantization error can only be improved by increasing ADC resolution (S. Franco, 1988).

The quantization error, also known as quantization noise, has a root mean square value E_n, which is related to the resolution of the system as:

$$E_n = V_{fs}/2^n\sqrt{12} \qquad (2.5)$$

Where V_{fs} is the full-scale voltage range of the converter, and n is the number of bits of the A-D converter.

If we observe equation (2.5) carefully, it can be deduced that for each additional bit of resolution, E_n is cut in halve.

- *Conversion Error* – An ADC may also suffer from non linearity, offset error, gain error and stability. If the differential non-linearity exceeds 1LSB, some digital codes may be missing at the output. Missing codes are intolerable, especially in control applications, where it may lead to instability.

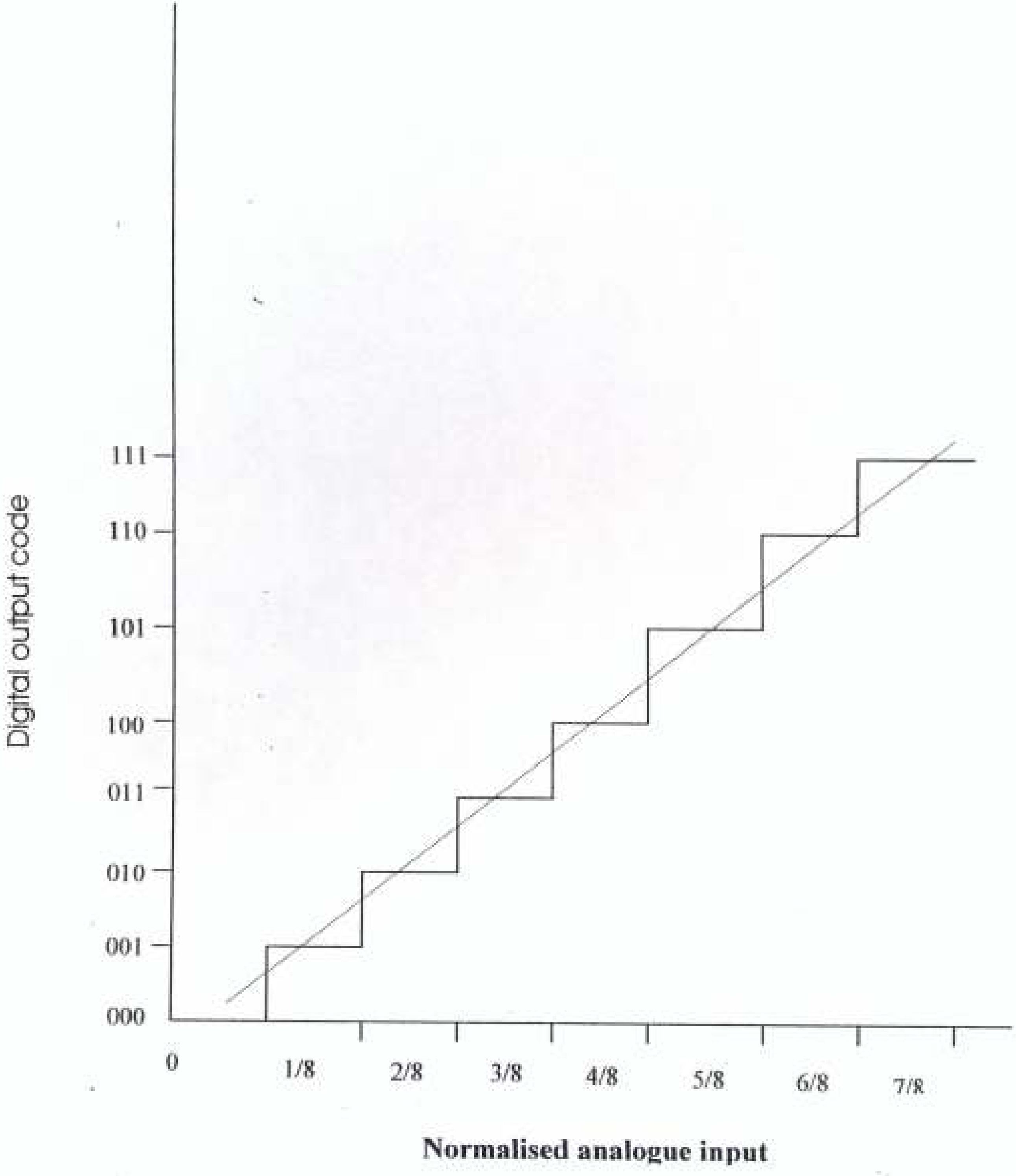


Fig 2.16: The transfer characteristics of 3 bits analogue-to-digital converter

Normally ADC manufacturers will provide specification on the potential size of the different types of conversion error.

- *Sample rate* – Usually all A-D conversion processes require a certain amount of time to produce the desired output code following the arrival of the START command. This amount of time, known as ***conversion time/sample rate*** is determined by the A-D conversion technique, resolutions and technology. The conversion time can range from as little as 10ns to as much as 100ms.

Software often allows us to specify any sample rate up to the maximum value.

2.12.3 Signal Conditioning for A-D Conversion

To make the best use of an ADC, conditioning of the input signal may at times be required. The conversion of dc signals is straightforward, but for ac signals, we must sample the input. The rate at which we sample is determined by the sample rule, which requires that a signal of bandwidth *fHz* be sampled at a rate in excess of $2f$ samples per second.

If the A-D converter cannot maintain this rate of conversion, it is essential that the input signal bandwidth be restricted by using a low-pass (otherwise known as anti-aliasing) filter to remove signals of frequencies greater than $f_{sample}/2$.

Please note that with the above-mentioned arrangement, important information may be lost.

To prevent any error arising from the variations in the input signal during conversion, a sample-and-hold circuit usually precedes some ADC. This is done in other to present a stable dc level prior to conversion cycle of the ADC.

2.13 MULTIPLEXERS

Multiplexers are devices with many inputs, but only one output. They are like switches, with the capability of selecting one input out of its many inputs at a time and connect it to the output. There are analogue multiplexers as well as digital ones. Figure 2.17 is a schematic diagram of a 4-input analogue multiplexer. Multiplexers are necessary in systems where several analogue signals are needed to be converted one after the other to binary codeword. Instead of using a separate and possibly expensive A-D converter for each channel, a single multiplexer is adequate to adapt all the channels to be used with only one ADC. Multiplexers contain field-effect transistors (FETs), so that when they are connected, they provide low resistance (~100Ω) signal path, which can be switched on or off by means off logic signals.

Multiplexers are available in many different configurations, and are sometimes incorporated into some A-D converter IC's.

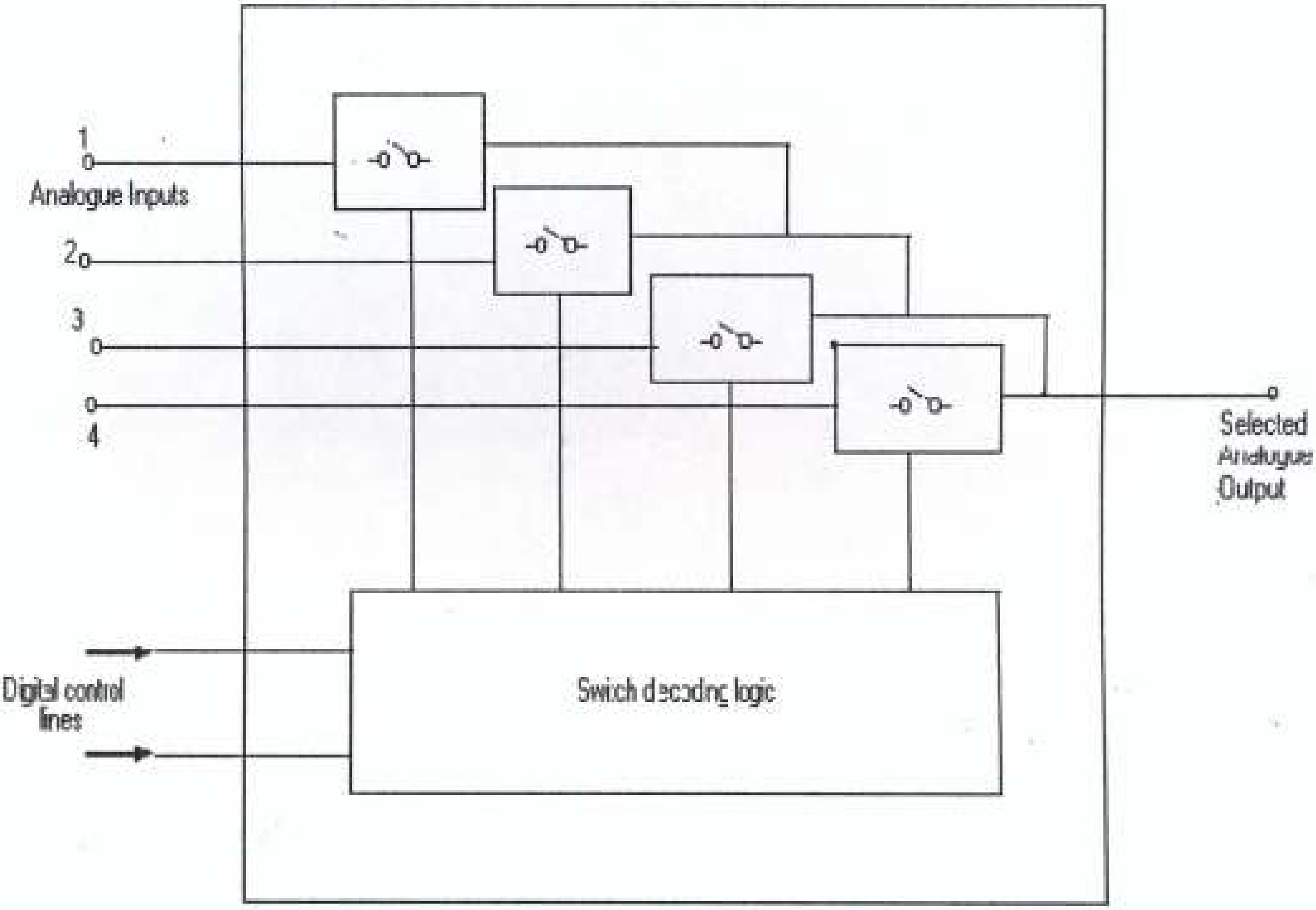


Fig 2.17: Schematic of a 4-input multiplexer

2.14 DATA INDICATION AND RECORDING DEVICES

The majority of signals in measurement systems ultimately appear as voltages. Since a voltage cannot be seen, it then becomes imperative to change the measured voltage to forms that are intelligible to human observer. The forms in which data are presented generally include a pointer moving over a scale as in potentiometers and D'arsonval meter. Others include a light beam writing on a photosensitive paper, an electron beam writing on a cathode- ray tube, the visual display of a set of ordered digits and the printout of digital data by a printer.

Digital computers, either dedicated or general purpose is widely becoming popular in the field of measurement and control. This is as a result of their programmability. The computer can interact with the outside through its ports, which are generally bi-directional. That is, they may either receive or output data. This process must however be done in well-defined and orderly manner. The computer must be informed through a software programme whether a particular port is to handle incoming or outgoing information. Software assignment of port function is referred to as ***configuring*** the port. Data transmission between the computer and its outside world may be handled in either serial or parallel form (Barney, 1988).

2.14. THE MICROCOMPUTER

This is a machine having two functional parts; the ***hardware*** and the ***software,*** working together to provide various forms of data communication and control.

For a better understanding of this versatile machine, an illustration of what each of its components can do and how they can be applied to practical problems is hereby presented next.

2.14.1 The microprocessor

The heart of a microcomputer is the central processing unit (CPU), otherwise known as the microprocessor unit (MPU). The microprocessor is a device that has evolved to provide means of handling the storage, retrieval and manipulation of data in systems with large memory facility. The microprocessor is built from fundamental logic devices like – registers, gates, ROMs and buses. The microprocessor is a sequential logic component, whose behaviour is determined by a list of ***instructions*** stored in the memory. These lists of instructions are called ***programs***. Normally digital memories can only hold binary words, so the instructions are represented by codes, which the microprocessor is designed to interpret. Each instruction directs the microprocessor to carry out an ***operation*** such as, adding two numbers or storing an item of data in the memory. Also the instruction must either contain the ***operand*** (the numbers to be added or the data to be stored) or it must specify the address at which the operands are stored.

It is the program stored in the memory that determines the behaviour of the microprocessor, and a particular microprocessor can be used in variety of applications, depending on the application and level of complexity. Microprocessors are available in 4bits, its, 16bits, 32bits and so on.

2.14.2 Structure and operation of microprocessors

As it has been mentioned, a program of instruction stored in an addressable memory determines the behaviour of a microprocessor. Each instruction indicates either implicitly or explicitly, where the next instruction is stored. So, once a computer begins to run a program, it automatically runs through.

Instructions (e.g., load, store, add & jump) always occupy a fixed number of memory locations, and these instructions are usually carried out in times of less than one or two nanoseconds. The actual set of instructions that control microprocessor operations are written in ***machine codes.*** The machine codes make use of the binary system, in which instructions are only represented in zeros and ones.

However, working out instructions in machine codes is time consuming and very difficult. Therefore in practice, programs are written in some other forms of code, otherwise known as ***high-level language.*** In this mode, instructions are represented by words or abbreviations, while a translation program that is running on the microprocessor itself then generates the actual machine code that can be directly executed by the processor. The way the microprocessor is organised, the function of the main registers and the overall structure of internal connections are called the ***architecture.***

The basic infrastructures that determine the performance of microprocessors are:

(a) Instruction set – This gives the range of instructions that a particular microprocessor can carry out.

If we consider the progress made so far in microprocessor technology, we will note that the first microprocessor was the '80 series' introduced in 1974 by Intel Corporation. This was a 4bit microprocessor, meaning those internal registers and data buses are 4-bits wide. The 8-bits devices, which followed on, were the 8008 and the 8080 in 1974. The 8080 design progressed to the 16-bits 8086 and 8088 (used in the IBM PC and similar models) and

80286 (used in the IBM AT and others). The 32-bits 80386, 80486 and 80586 are progressively more powerful developments that have led to 286, 486 and 586 computers.

(b) Clock – The clock sets the regular signal that paces the processor's activity, thereby determining the time for the execution of an instruction.

(b) Clock – The clock sets the regular signal that paces the processor's activity, thereby determining the time for the execution of an instruction.

The faster the clock, the shorter is the overall execution time. The maximum possible clock speed is therefore another important indication of overall performance.

2.14.3 Memories

The storage and recall of information is a fundamental property of sequential logic circuits. The basic circuit elements that perform storage operations are: latches, flip-flops and registers.

Flip-flops/Latches are logic elements that are capable of holding their outputs until they are told to change it. These stable outputs can either be 1(set state) or 0 (reset state). A latch/flip-flop can only store one bit of information. To have a device that is capable of storing larger bits of information, several latches/flip-flops are stacked together to form registers. Many registers are stacked together to form memories.

Thus an electronic memory consists of a large number (typically between a few hundred and a few million) of independent storage locations. Each location holds a binary word. Presented next is the various memory facilities available to our use on the general-purpose computer.

Random-access Memory (RAM)

RAM chips represent by far the largest share of electronic memory components. They form the basis of computer memory and other large digital systems. RAM chips are either static or dynamic. In a static RAM (SRAM) each bit is stored by a flip-flop consisting of several transistors, which remain in a fixed state (as long as the power supply is connected to the chip) until it is deliberately changed. However in a dynamic RAM (DRAM) each information bit is stored as a certain amount of charge on a small capacitor that is connected to a single transistor.

The major advantage of dynamic RAMs lies in their simple structure, and they can store more bits per unit area because the basic memory cell occupies less space. SRAMs are faster than DRAMs. On any given chip, the size of binary word that can be stored in each location is the same and is predetermined during manufacture. Common sizes for the words stored on RAM chips are 1bit, 4bits or 8bits (1 byte). Hence a memory chip that holds 256K bits could be organised internally as 32K locations with addresses running from 0 to 32767 (= 32K – 1) and each location holding 1 byte. The major drawback with RAMs is that they are ***volatile.*** This means that when the power supply to it is switched off, all the data contained in it is lost.

Read-only Memory (ROM)

Read-only memories (ROMs) provide non-volatile storage of data which is then available as soon as the device is connected to the appropriate power supply. They are structured in the same way as RAM, with each data location being specified by an address. However, in the use, only a read operation can be carried out on them. The process of fixing ROM contents is called programming the device, and different types of ROM are available which require this to be done either by the chip manufacturer (as in a mask-programmed ROM) or by the user (as in the various forms of programmable read-only memory, or PROM).

There are different versions of the PROM. In one form of the device, internal connections are either fused or left intact to create the data pattern. These are often referred to as FPROMs (standing for fusible link PROM or field-programmable ROM). Another more common form of the PROM is the one-time programmable ROM (OTPROM). In the device, packets of electrical charge that remain fixed represent data even when the device has no power connection. In either case, programming is a once-and-for-all process, and the memory contents, once programmed, cannot be altered.

Erasable PROM is another version of ROMs whose content can always be re-fixed. These devices are known by several names and acronyms: **electrically erasable and programmable ROM, (EEPROMs or E^2PROMS), or electrically alterable ROM (EAROM)** are the main ones. In these devices electrical signals are often used to erase the individual cells before a reprogram can be carried out. A more common method of erasure is achieved by illuminating the whole chip (through a transparent window in the package) with fairly high power ultra-violent light for several minutes. This causes the stored charge to leak away. The difference between an electrically erasable PROM and ultra-violent EPROM is that each cell can be erased and fixed independently in electrically erasable PROM, while all the content of the chip will be erased in the ultra-violent erasable PROM once it is exposed to UV-light.

Programming of these devices normally requires a PROM programmer, a special-purpose piece of equipment, which is connected to a computer. The computer supplies the required data for each address to the PROM programmer, which then applies the appropriate signals to the PROM to store the data. ROM chips – whether mask-programmable or PROMs – can be built up into large memory systems in an identical way to RAMs. For instance, a 256K by 8 ROM memory system could be built up from four 64K by 8 PROMs in a manner analogous to the RAM chips.

2.15 INTERFACING THE STANDARD PARALLEL PORT

The parallel port of the computer is the most commonly used port for interfacing home made projects. The port allows for the input of up to 9 bits or the output of 12 bits at a time. The port has four control lines, 5 status lines and 8 data lines. The port is found at the back of a general purpose computer as a D-type 25 pin female connector. There may also be a D-type 25 pin male connector.

There are five standard modes of operation of parallel ports under the IEEE 1284 standard of 1994. These five modes of operation are:

- Compatibility mode
- Nibble mode
- Byte mode
- Extended parallel port (EPP) mode
- Enhanced port (ECP) mode

The aim of these various modes is to design new drivers and devices that run at faster speed, but still compatible with each other and also compatible with the standard parallel port (SPP). Compatibility, bytes and nibble modes make use of the standard hardware available on the original parallel port cards while EPP and ECP modes require additional hardware that can run at faster speeds.

Compatibility mode or 'centronics' as it is commonly called can only send data in the forward direction at an average speed of 50 kilo-bytes per second but the speed can be as high as 150 kilo-bytes per second. In order to receive data, compatibility mode must be changed to either nibble or byte mode. Nibble mode can input a nibble (4 bits) in the reverse direction (i.e. from device to the computer). Byte mode uses the parallel's bi-directional feature (found only on some cards) to input a byte (8bits) of data in the reverse direction.

Extended and enhanced parallel ports use additional hardware to generate and manage handshaking. The ECP and EPP can output at around 1-2 megabytes per second.

In Table 2.3, the prefix "n" in front of signal name is used to indicate that such signals are actively low. The "hardware inverted" means that the signal is inverted by the parallel card. The output of the parallel port is normally at TTL logic levels. Most parallel ports implemented in ASIC (Application Specific Integrated Circuit), can sink and source around 12mA. It is always better to have an adequate knowledge of the current capability of each device one is using. This information can be obtained from the manufacturer's data sheet.

2.15.1 Parallel Port Addressing

The parallel port has three commonly used base addresses. These are listed in Table2.4 below. The $3BC_h$ base address was originally used for parallel ports on video cards, but they are now integrated on some motherboards as an option for parallel port. LPT1 is normally assigned base address 378_h, while LPT2 is assigned 278_h. However this may not always be the case as explained later. 378_h and 278_h have always been commonly used for parallel ports. Though this addresses may change from machine to machine. The subscript "h" in front of the addresses denotes that they are in hexadecimal.

Table 2.3: Pin numbering of the 25 pin D-type and the 36-pin centronic connectors

D-type 25 pin Number	Centronics Pin Number	Standard Parallel Port signal	Direction In/Out	Register	Hardware Inverted?
1	1	nStrobe	In/Out	Control	Yes
2	2	Data0	Out	Data	-
3	3	Data1	Out	Data	-
4	4	Data2	Out	Data	-
5	5	Data3	Out	Data	-
6	6	Data4	Out	Data	-
7	7	Data5	Out	Data	-
8	8	Data6	Out	Data	-
9	9	Data7	Out	Data	-
10	10	nAck	In	Status	-
11	11	Busy	In	Status	Yes
12	12	Paper-out/ Paper-End	In	Status	-
13	13	Select	In	Status	-
14	14	nAuto-Linefeed	In/Out	Control	Yes
15	32	nError/nFault	In	Status	-
16	31	nInitialize	In/Out	Control	-
17	36	nSelect printer	In/Out	Control	Yes
18 – 25	19 – 30	Ground	Gnd	-	-

Table 2.4: Port Addresses

Address	Notes:
$3BC_h$ – $3BF_h$	Used for parallel ports which were incorporated onto video cards – Doesn't support ECP addresses
378_h – $37F_h$	Usual addresses for LPT1
278_h – $27F_h$	Usual addresses for LPT2

When the computer is first turned on, BIOS (basic Input/Output System) determines the number of ports on the motherboard of a computer and assigns device labels LPT1, LPT2 and LPT3 to them. BIOS first look at address $3BC_h$. If a parallel port is found here, it is assigned as LPT1. It then searches at location 378h. If a parallel card is found there; it is assigned the next free device label. This would be LPT1 if a card were not found at $3BC_h$ or LPT2 if a card was found at $3BC_h$. The last port of call is 278_h, and this follows the same procedure as the other two ports. Therefore it is possible to have an LPT2 that is at 378_h and not at the expected 278_h address.

However the idea of device label should not be a problem in projects dealing with interfacing devices to the computer. The reason being that in most cases, the base address is used rather than the device label. Moreover, if we are interested in finding the address of LPT1 or any of the line printer devices, we can use a look up table provided by BIOS. When BIOS assigns addresses to a printer device, it stores the address at a specific location in the memory, so we can find them.

2.15.2 SOFTWARE REGISTER FOR THE STANDARD PARALLEL PORT

A port can only perform one function at a time. It either gives out data or takes in data (one way traffic). If the port can input data at one time and output data at another time, such ports are said to be bi-directional. Therefore read and write operations can be performed on their data registry. Table 2.5 through Table 2.7 gives the properties of the various ports as found on the D-25 type connector.

Any data written to the data port (base address) by the computer can be accessed via (pins 2-9). If the port is not bi-directional or does not support read operation, any attempt to read from it will only give back the last bit sent.

The status port (base address + 1) is a read only port. Any data written to this port will be ignored. The status port is made up of 5 input lines (Pins 10, 11, 12, 13, & 15), an IRQ status register and two reserved bits. The control port (base address + 2) with the properties shown in table 2.7 is intended as a write only port. When a printer is attached to the parallel port, four "controls" are used. These are Strobe, Auto linefeed, Initialise and Select printer, all of which are inverted except Initialise.

The printer does not send signal to initialise the computer, nor does it tell the computer to use the auto linefeed. These lines are "open collector" outputs. This means that it has two states, a low state (0V) and a high impedance state (open circuit).

Table 2.5: Data Port

Address	Name	Operation: Read/write	Bit No	Properties
Base + 0	Data Port	Write*	Bit 7	Data 7
			Bit 6	Data 6
			Bit 5	Data 5
			Bit 4	Data 4
			Bit 3	Data 3
			Bit 2	Data 2
			Bit 1	Data 1
			Bit 0	Data 0

Table 2.6: Status Port

Address	Name	Operation: Read/Write	Bit No	Properties
Base + 1	Status port	Read only	Bit 7	Busy
			Bit6	Ack
			Bit 5	Paper-out
			Bit 4	Select In
			Bit 3	Error
			Bit 2	IRQ (Not)
			Bit 1	Reserved
			Bit 0	Reserved

Fig 2.7: Control Port

Address	Name	Operation: Read/Write	Bit No	Properties
Base + 2	Control Port	Read/Write	Bit 7	Unused
			Bit 6	Unused
			Bit 5	Enable Bi-directional Port
			Bit 4	Enable IRQ via Ack Line
			Bit 3	Printer Select
			Bit 2	Initialize Printer (Reset)
			Bit 1	Auto-Linefeed
			Bit 0	Strobe

Normally the printer card has internal pull-up resistors. However, as one would expect, not all of them have. Some have open collector outputs, while others have normal totem pole output. In order to make our device work correctly on as many printer ports as possible, we use an external resistor as well. If there should be an internal resistor on the card, then the two resistors will act in parallel. If it is a totem pole output, the resistor acts as a load. An external 4.7kΩ resistor is usually recommended for pull-up purposes. When in high impedance state, the pin on the parallel port is high. When it is in this state, the external device can pull the pin low and have the control port changed to read a different value. In this way, the four pins of the control port can be used for bi-directional data transfer. However the control port must be set to xxxx0100 for it to be able to read data; that is all pins must be at logic 1 at the port so that it can be pulled down to logic 0.

Bits 4 and 5 are internal controls. Bit 4 enables the IRQ and bit 5 enable the bi-directional port; meaning that we can input 8 bits using (Data 0-7). This mode is only possible if the computer card supports it. Bits 6 and 7 are reserved. Any attempt to write to these two bits will be ignored (Odo, 2002).

CHAPTER 3

DEVELOPMENT AND CONSTRUCTION OF THE WEATHER DATA ACQUISITION SYSTEM

3.1 INTRODUCTION

Precision weather data acquisition and control system design depends on how precisely the electronic parts are carefully designed. It therefore becomes imperative to make use of all the knowledge gained so far on precision circuits and low noise techniques. This chapter gives all the design parameters used in the design and development of the weather data acquisition system.

In other to describe fully the entire circuitry used to accomplish this project, we have divided it into two major parts. The first part describes all the analogue units, while the second part is devoted to the digital aspect. The software design is also considered under the digital design unit.

3.2 TEMPERATURE, PRESSURE & HUMIDITY SENSORS

The choice of any sensor for a particular application is dependent on three major factors. These factors are:

1. The range of measurement that the sensor is expected cover.
2. The environmental condition in which the measurement is taking place.
3. The degree of accuracy, sensitivity and consistency required for repeated measurement operation using the same sensor.

However, if the above stated factors are considered alongside the aim and objectives of this work, it suffices to say that semiconductor monolithic temperature, pressure and humidity sensors are adequate for the work.

Monolithic temperature sensors exist in two major categories. The sensors in the first category always have their output voltage proportional to absolute temperature. They are otherwise known as VPTAT sensors. The output of the second category is a current, which is always proportional to absolute temperature. They are called IPTAT sensors.

The AD590 IPTAT is chosen for this project. The main reasons for this choice are:

- Its temperature range is adequate for our application (-55°C to 150°C).
- Being a current output device, it is better than other thermal sensors in any application that involves remote sensing.

Appendix A gives the detailed specification of the AD590 IC, as released by RS-components.

Semiconductor Monolithic pressure sensors are divided into ***absolute, gauge*** and ***differential***.

For our application, we have chosen the 24PC series of absolute (particularly 24PCCFD6A) pressure sensor. Honeywell Inc manufactures this device. It is rated to have an absolute pressure range between 2 – 15psi, which is equivalent to about 1Bar. The complete manufacturers data sheet is presented in appendix B.

The humidity sensor (HIH 3610 - 001) used in this project is a product of Hycal sensing instruments. It is a capacitive type of humidity sensor with high sensitivity and low hysteresis. The complete manufacturer data sheet is shown in appendix C.

ANALOGUE CIRCUIT DESCRIPTION

3.3.1 SENSOR AMPLIFIERS

In the design of any sensor amplifier, the following points are of importance (Odo, 2002).

1. The type of sensor to be used.
2. The full range of voltage from the sensor
3. The output impedance of the sensor

(a) **Temperature amplifier**

The AD590JH thermal sensor amplifier is built around U1 (Fig. 3.1), a commercially available precision bi-polar op-amp (type OP-07). According to its manufacturer (Analogue Devices, 2001), the op-amp exhibits low drift with time and temperature (0.6mV/°C) and a low offset voltage (7.5mV max). The op-amp U1 is operated in the inverting mode. This condition is required to present low impedance to the sensor, thereby preventing self-heating in the sensor. The sensitivity of the amplifier is equal to R2*1μA/°C. The gain is trimmed to 50mVC by adjusting RV2. By adjusting RV1 an offset of about 273.2μA is generated, which turns the AD590JH IPTAT sensor to a centigrade sensor and at the same time providing external offset nulling. R_3 is included to minimize the effect of the op-amp input bias currents. ($R_3 = R_1//R_2$). The calibration of the amplifier is done in two ways; with the temperature at 0°C (ice bath), R_1 is adjusted for a zero output voltage. With the temperature at 100°C (boiling water), R_2 is adjusted for an output voltage of 5volt.

(b) pressure amplifier

The pressure sensor amplifier (shown in Fig 3.2) is built around U1, U2 and U3 (all OP-07). They are arranged to form an ***instrumentation amplifier***. The amplifier has the following properties (Akpan, 2003):

- Finite, accurate and stable gain.
- Extremely low (ideally zero) output impedance
- Extremely high (ideally infinite) input impedance
- Extremely high (ideally infinite) CMRR so that the device responds only to

differential voltage between the inputs and completely ignoring the common-mode input component.

An ordinary op-amp operated in the differential mode satisfies the first two requirements mentioned above (Adedayo, 2002). All the four qualities are needed in our pressure amplifier, hence our choice of an instrumentation amplifier. The amplifier's gain is given by $(1 + 2R_3/Rg)(R_2/R_1)$. The gain of the amplifier is made variable between 4.3 and 21 via Rg. Rg comprise of both the 10k and 50k port. The 5k port (part of R_2 connected to ground), was included to optimise the CMRR of the amplifier (Franco, 1988). The 50k (*R*comp) was included to minimise any error due to lower temperature effect in the sensor (see appendix B). *R*comp ≈ 10*Rs. Where Rs is the output resistance of the pressure sensor. To further forestall any error due to high temperature, the sensor was driven by a constant current of 2mA (see appendix B). The LM317, U4 (µA741CN) and R_{set} were arranged to form a constant current source. The output current from the source is given as; $1.25/R_{set.}$

For current of 2mA, R_{set} was adjusted for a value of 625Ω. U5 (µA741CN), 3.3V zener and the 1N821 were used to construct a self-regulated voltage reference for the circuit. The output voltage from the reference is given as; $(1 + R4/VR5)*3.3$. The voltage was set to 10V by adjusting *VR*5. 1N821 was included in the circuit to eliminate any drift in the output voltage that may arise as a result of increase in ambient temperature.

(c) Humidity amplifier

No amplifier was constructed for the Humidity sensor because the output from the sensor is large enough to drive the ADC. From the Manufacturer data sheet (Appendix C), the sensor will produce 0.958V at 0% relative humidity at 25C. At 75.3% RH the output is quoted to be 3.268V at 25C.. The only circuitry needed and constructed for the efficient performance of the sensor was a voltage reference similar to the one described above for pressure sensor. The voltage reference was adjusted for an output of 5-Volt.

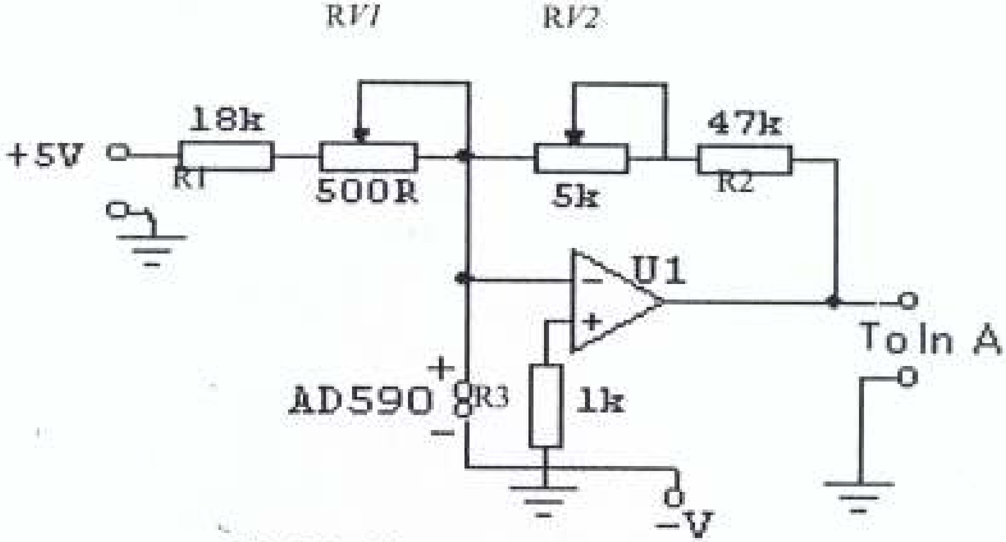


Fig 3.1: The temperature sensor amplifier

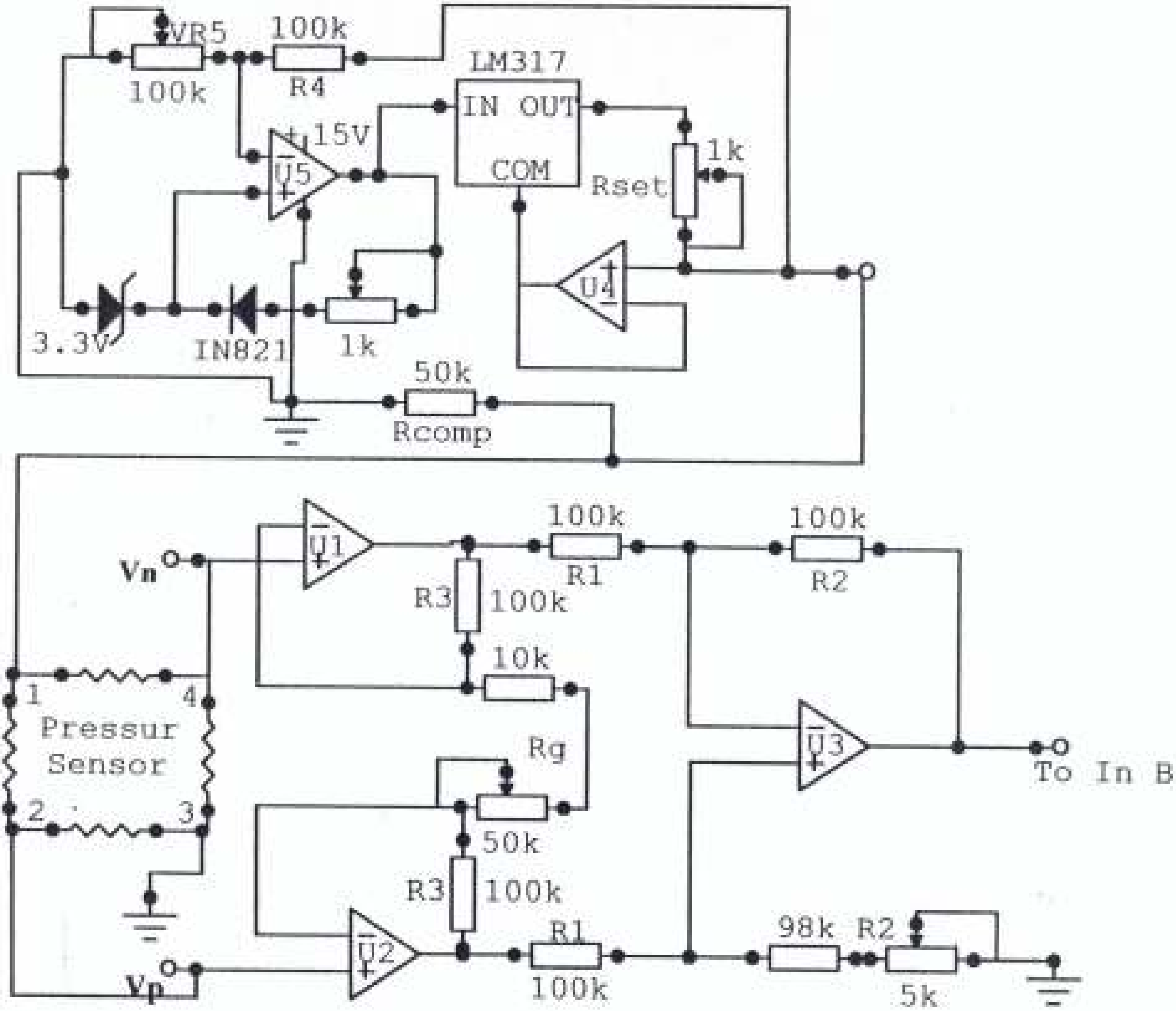


Fig 3.2: The pressure sensor amplifier

3.4 DIGITAL INTERFACING CONSIDERATIONS

For this project, the printer's port of the general-purpose computer was adapted. Fig 3.3 shows how the port looks like at the back of desktop personal computers. The port provides eight TTL outputs D0-D7, five inputs S3-S7 and four bi-directional leads C0-C3. It also provides an easy way to use the PC's interrupt structure.

As we have mentioned earlier in section 3.8, five modes can be used to implement parallel interfacing. Out of these five, this project makes use of the byte mode. This mode enables the input and/or output of 8bits at a time. The TC7109A analogue-to-digital converter used for the interfacing has many features (see appendix D). The ADC is a 12-bit device. Our printer's port can only take 8bits at a time. Driver software was developed to provide a means of taking the information in a succession of eight lower bits followed by the four upper bits. For the general description of the driver software see page 59-60.

3.4.1 DIGITAL CIRCUIT DESCRIPTION

The digital circuit is divided into two parts – the ADC and the software module. Fig 3.4 shows the entire ADC module. The converter was controlled through its pins 26 (RUN/HOLD), 18 (Lower byte enable) and 19 (High byte enable). The Run/Hold pin toggles between 0 and 1, so as to enable the converter to respond and make conversion from more than one source. Pins 18 and 19 were also controlled so that the 12bits from the converter can be conveniently taken through the 8 lines available on the printers' port in succession of the first 8 lower bits, followed by the last 4 upper bits.

The conversion rate of the converter is given as **Fosc** = 0.45/RC. The converter was set to five conversions per second (40.96kHz). In addition, it was ensured that the period of our RC oscillator conformed to this requirement; $2^{n}T_{clock}$ = integral multiples of the line supply voltage (see appendix D). This condition is required to be met when using any integrating type of ADC, so that errors arising from line interference will be eliminated.

A resolution of 1bit/mV was achieved by adjusting the reference input voltage at pin 36. The combination of R_1 and *VR*1 allows for the trimming of the reference voltage. This arrangement allows for a full-scale 12bits output i.e. all outputs bits B1-B12 are high when an input voltage of 4096mV is placed at the input or when the test pin (pin 17) is tied low.

The busses carrying the upper 4bits were connected to the first four busses of the 8 lower bits, such that under software control, the computer successfully read the 12 bits.

All the four control lines available on the printer's port were used in this project. The first two C0 and C1 (pins 1 and 14) were used to control the multiplexer, which selects the

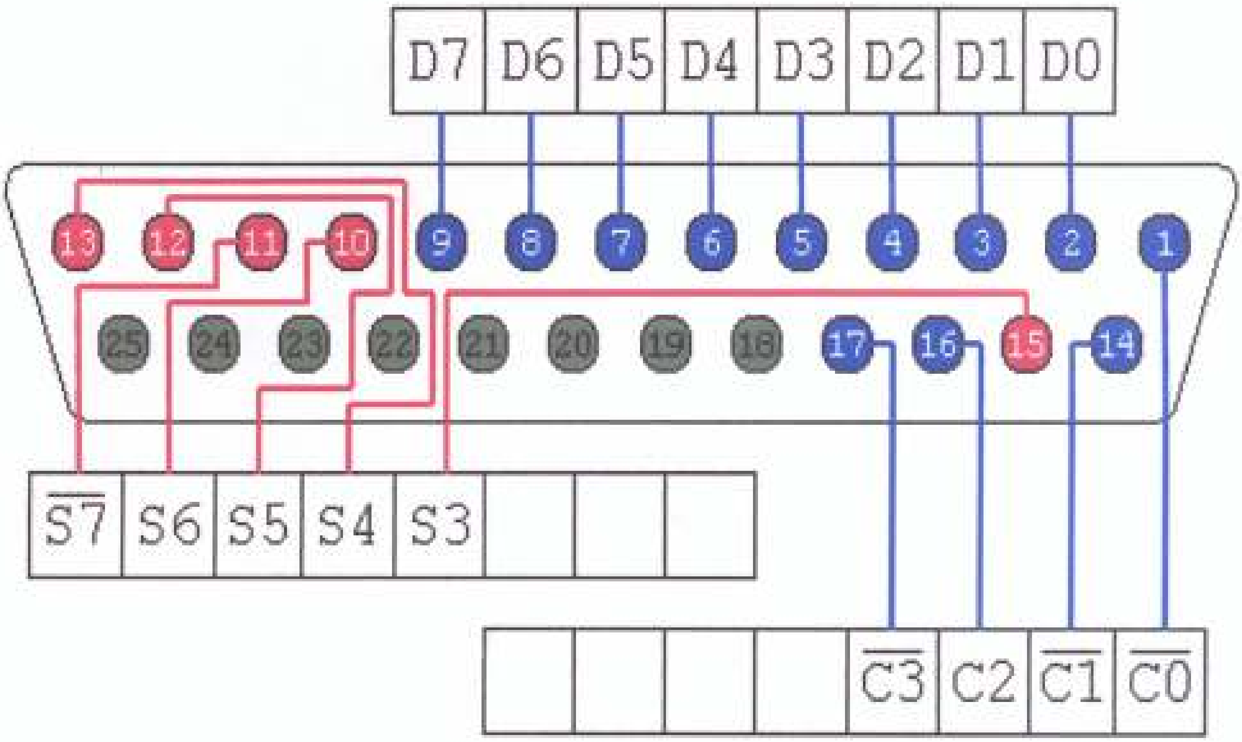


Fig 3.3: 25-way Female D-Type Connector

appropriate input to connect to the ADC. While the other two control lines C2 and C3 (pins 16 & 17) are used to control the ADC.

It is noted however that C0, C1 and C3 (fig 3.3) are all actively low i.e. a high placed on these pins by the computer processor is internally inverted. Therefore, if 1 is placed on pins 1 and 14, it will make the multiplexer to select input 1. If the signals on pins 1 and 14 are 1 and 0 respectively, then the multiplexer will select input 2. Placing 0 and 1 on pins 1 and 14 will cause the multiplexer to select input 3.

When 1 is placed on pin 16 of the control port, the ADC will perform conversion continuously at a rate determined by the ADC's clock (0.45/RC). If pin 16 goes low during any conversion process, the ADC will finish whatever conversion it is handling before it stops. Pin 17 is alternatively toggled between 1 and 0 so as to get the 12bits data B1-B12 through the 8bits data lines D0-D7.

3.4.2 SOFTWARE DESCRIPTION

The entire process of control, data acquisition and processing of the acquired data is carried out by computer software. The software was written in Q-basic and Visual basic.

The entire program is made up of four sub-programs First, a graphical screen was prepared which allows the user to specify the data that he/she wants to log and the logging time interval. When the user clicks "start", a file is opened for both read and write operation and a timer is initiated. The main program then calls the control-sub-program, which is a DLL (Dynamic Link Library). The DLL enables the control program to interact with the external circuitry.

Data are read into a temporary location on the RAM of the computer. The processing of these data is also carried out in the RAM, before they are sent to the Hard Disk Drive for permanent storage.

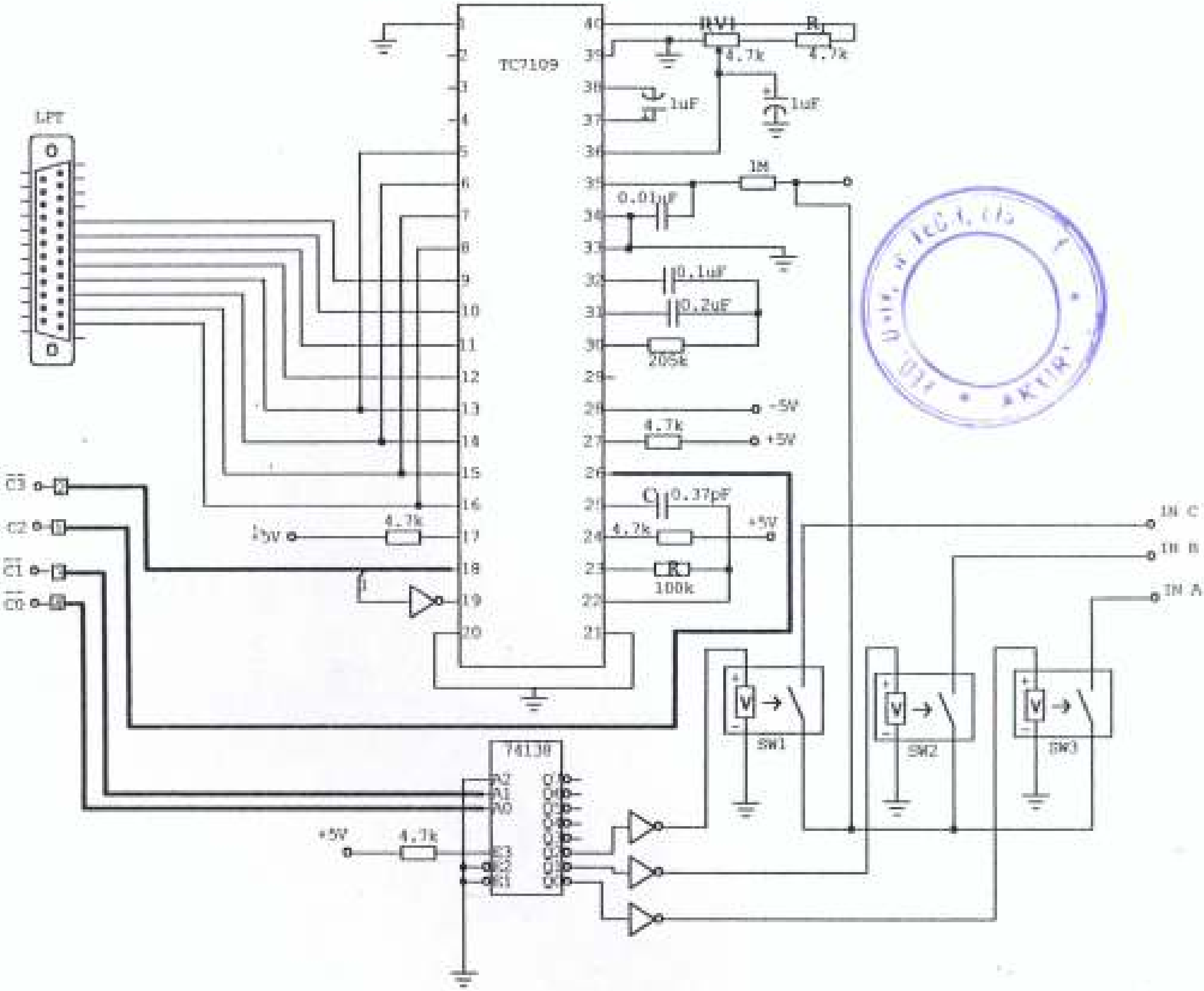


Fig 3.4: Circuit diagram of the Analogue-to-Digital converter interface module

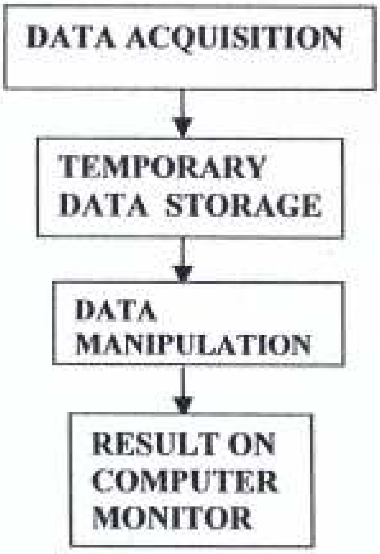


Fig 3.5: Flowchart for the computer interface

```
Option Explicit
1         Dim Lower As Integer
2         Dim Higher As Integer
3         Dim result As Integer
4         Dim p As Integer, d As Integer
    •     Public Function ADC1(channel1 As
          Integer) As Integer
'Run ADC, select channel1 and enable lower byte
5         Out 890, 255
6         For d = 1 To 500
7         Next d
          'Hold ADC
8         Out 890, 251
9         For d = 1 To 500
10        Next d
          'Fetch lower byte data
11        Lower = Inp(888)
12        For d = 1 To 500
13        Next d
          'Enable higher byte
14        Out 890, 243
15        For d = 1 To 500
16        Next d
          'Fetch higher byte data
17        Higher = Inp(888)
18        Select Case Higher
            Case 1: p = 256
            Case 2: p = 512
            Case 3: p = 768
            Case 4: p = 1025
            Case 5: p = 1280
            Case 6: p = 1536
            Case 7: p = 1792
             Case Else: p = 0
            End Select
19          Higher = p
20        ADC1 = Higher + Lower
21        End Function
```

```
Public Function ADC2 (channel 2 As Integer) As
Integer
'Run ADC, select channel2 and enable lower
byte
22        Out 890, 254
23        For d = 1 To 500
24        Next d
          'Hold ADC
25        Out 890, 250
26        For d = 1 To 500
27        Next d
          'Fetch lower byte data
28        Lower = Inp(888)
29        For d = 1 To 500
30        Next d
          'Enable higher byte
31        Out 890, 242
32        For d = 1 To 500
33        Next d
          'Fetch higher byte data
34        Higher = Inp(888)
35        Select Case Higher
            Case 1: p = 256
            Case 2: p = 512
            Case 3: p = 768
            Case 4: p = 1025
            Case 5: p = 1280
            Case 6: p = 1536
            Case 7: p = 1792
            Case Else: p = 0
            End Select
36          Higher = p
37           ADC2 = Higher + Lower
38           End Function
Public Function ADC3(channel3 As Integer) As
Integer
'Run ADC, select channel1 and enable lower
byte
39        Out 890, 253
40        For d = 1 To 500
41        Next d
```

Table 3.1 Cont. Source-code for the ADC control-subprogram.

```
'Hold ADC
Out 890, 249
For d = 1 To 500
Next d
'Fetch lower byte data
Lower = Inp(888)
For d = 1 To 500
Next d
'Enable higher byte
Out 890, 241
For d = 1 To 500
Next d
'Fetch higher byte data
Higher = Inp(888)
Select Case Higher
 Case 1: p = 256
 Case 2: p = 512
 Case 3: p = 768
 Case 4: p = 1025
 Case 5: p = 1280
 Case 6: p = 1536
 Case 7: p = 1792
 Case Else: p = 0
  End Select
  Higher = p
  ADC3 = Higher + Lower
  End Function
```

CHAPTER FOUR

RESULT EVALUATION, DISCUSSION AND RECOMMENDATION

The actual construction of the electronics parts of the project was carried out in phases on different Vero boards. Each section of the entire circuit was tested separately before the final coupling and calibration. This chapter presents the experimental procedure and the results of the various tests carried out on each section of the design and concluding comments about all modifications on the initial design.

4.1 TEMPERATURE SENSOR AMPLIFIER TEST RESULT

The temperature amplifier which forms part of the analogue circuit presented in Fig.3.1 was set to a sensitivity of 50mV/°C using the trimmer resistor *RV*2. For the test, a mercury-in-glass thermometer was used for comparison. The thermometer and the temperature sensor (AD590JH) were both suspended in molten ice until the mercury-in-glass thermometer reads 0°C. At that moment, *RV*1 was adjusted and the amplifier output was set to 0.0 Volt on a digital voltmeter. The sensor and the thermometer were then suspended in boiling water at 100°C, *RV*2 was adjusted and the amplifier output reads 5.00Volt. The temperature of the sensor and thermometer was then varied between 0.0°C and 100°C and the output voltage was recorded. The result are shown in Table 4.1

Fig 4.1 is a plot of the amplifier output versus temperature. The slope of the graph was calculated to be 50mV/°C over the measurement range.

Stability measurement

The temperature amplifier was placed under observation for a very long time and it was discovered that it has a stable output whose drift with time was 10mV over 7days for the same temperature reading. The observed linearity as shown in Fig 4.1, which shows the plot of the result in Table 4.1 is a prove that the temperature sensor amplifier thus constructed meets the requirement of this project. The slope of the graph shows the sensitivity to be constant at 50mV/°C over the measurement range.

Table 4.1: Temperature sensor amplifier characteristics with its sensitivity set at 50mV/°C

Thermometer Reading/°C	OUTPUT VOLTAGE/Volt
0.00	0.0000
5.00	0.2750
10.00	0.5200
15.00	0.7750
20.00	1.0300
25.00	1.2750
30.00	1.5250
35.00	1.7800
40.00	2.0250
45.00	2.2800
50.00	2.5250
55.00	2.7800
60.00	3.0300
65.00	3.2750
70.00	3.5300
75.00	3.7750
80.00	4.0250
85.00	4.2750
90.00	4.5250
95.00	4.7750
100.00	5.0150

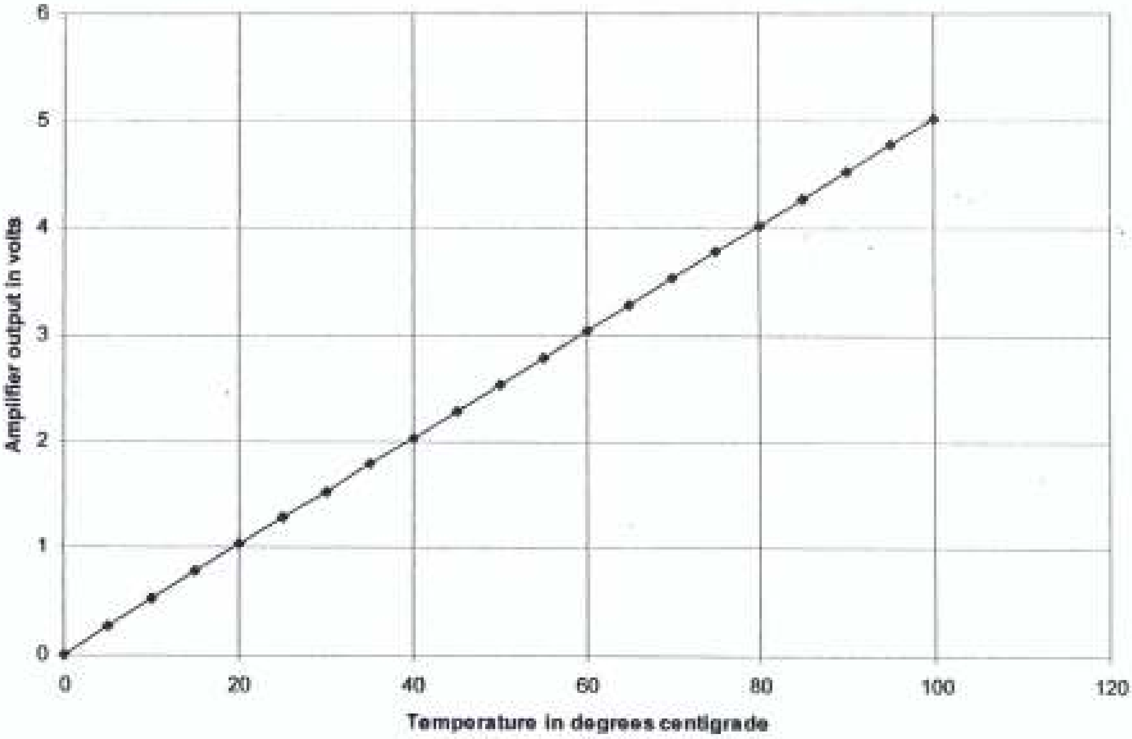


Fig 4.1: The temperature sensor amplifiers output voltage versus environmental temperature

4.2 Pressure amplifier test result

The pressure amplifier is shown in Fig. 3.2. The gain of the amplifier was set to 10 by adjusting the trimmer resistor Rg. For the test, a varying d.c. voltage of between 0 and 217.50mV was applied at the input of the pressure sensor amplifier. This simulates the expected output voltage range from the 24PCCF6DA pressure sensor used. The sensor is rated at 15mV/psi and it has a maximum range of 14.50psia. The corresponding output voltage was measured with a digital voltmeter. The results obtained are tabulated in Table 4.2. The linearity of the amplifier is shown in Fig. 4.2. The amplifier was also placed under observation for many days. It was observed that the amplifier has a stable output.

4.3 COMPUTER INTERFACE ADAPTER AND PROGRAMME TEST

This unit is made up of the 12bits Analogue-to-Digital converter. It was both software and electronically calibrated to ensure its long time integrity. For the electronic calibration, the ADC was set to 1bit/mV by placing a voltage of 4096mV at its input and $RV1$ was adjusted until all the outputs go high. i.e the output was 111111111111

After the initial calibration, different voltages were then applied at the input and the binary output produced by the ADC was monitored with a digital voltmeter. The result is hereby presented in Table 4.3.

The software calibration of the ADC was done by connecting the device to the computer through its printer's port. The control program was then run. First a known input voltage was placed at the input of the ADC; a subprogram called calibrator is activated. The program asks the user to enter a calibrating value into a textbox and then press return. The program then compares the users entered value with the value decoded from a read operation from the printer's port. It then performs the operation (Input voltage by user/Converted voltage from the port), to generate a correcting factor that acts as a multiplier to other converted readings from the port. This further compensates for any error that could arise as a result of changes in environmental condition after the initial electronic calibration.

The output result as displayed on the computer monitor (Fig. 4.3) and the results retrieved from the data logged shows that the software works as expected. The source-code of the calibration program written in Visual Basic is presented in appendix E.

Table 4.2: Pressure sensor amplifier characteristics with its gain set at 10.

Input voltage V_p(mV)	Input voltage V_n(mV)	$V_D = V_p - V_n$ (mV)	Output voltage (mV) at 27°C
0.00	0.00	0.00	0.001
60.00	20.00	40.00	400.001
120.00	40.00	80.00	800.001
180.00	60.00	120.00	1200.001
240.00	80.00	160.00	1600.001
300.00	100.00	200.00	2000.00
360.00	120.00	240.00	2400.00
420.00	140.00	280.00	2800.00

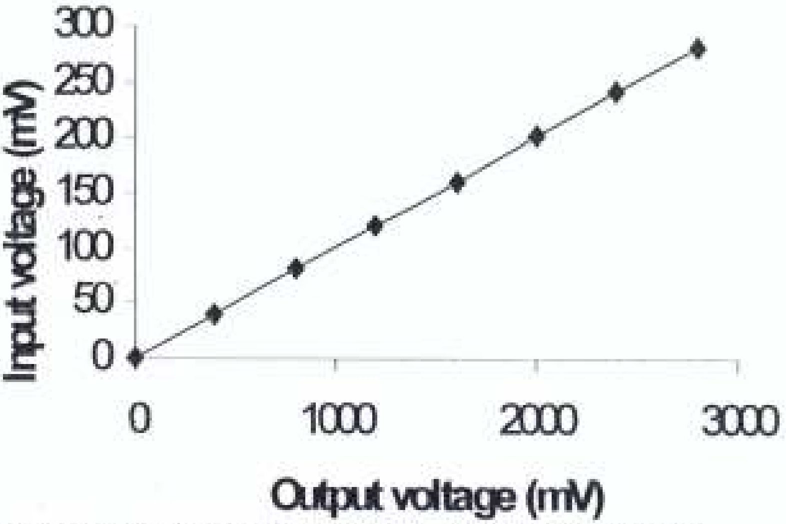
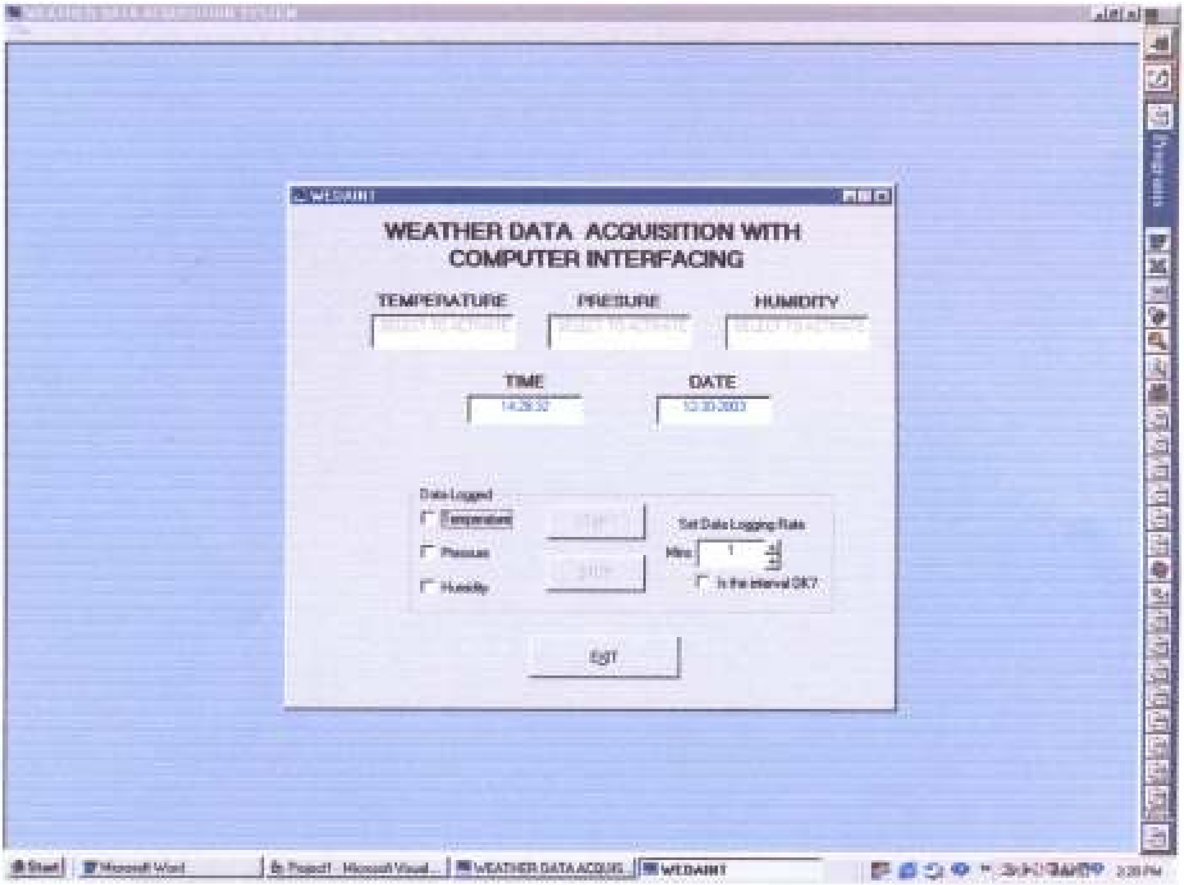


Fig 4.2: The pressure sensor amplifier output versus input voltage

Table 4.3: The Analogue-to-Digital Converter Output Characteristics

Input Voltage (mV)	Voltages recorded by the digital voltmeter; 1 indicates a high output (5V) while 0 indicates low output (0V) B12	B11	B10	B9	B8	B7	B6	B5	B4	B3	B2	B1
10	0	0	0	0	0	0	0	0	1	0	1	0
50	0	0	0	0	0	0	1	1	0	0	1	0
100	0	0	0	0	0	1	1	0	0	1	0	0
200	0	0	0	0	1	1	0	0	1	0	0	0
400	0	0	0	1	1	0	0	1	0	0	0	1
600	0	0	1	0	0	1	0	1	1	0	0	0
1000	0	0	1	1	1	1	1	0	1	0	0	1
1200	0	1	0	0	1	0	1	1	0	0	0	0
1400	0	1	0	1	0	1	1	1	1	0	0	0
1500	0	1	0	1	1	1	0	1	1	1	0	0


WEATHER DATA ACQUISITION WITH COMPUTER INTERFACING

TEMPERATURE: SELECT TO ACTIVATE

PRESURE: SELECT TO ACTIVATE

HUMIDITY: SELECT TO ACTIVATE

TIME: 14:28:32

DATE: 12-30-2003

Data Logged

☐ Temperature

☐ Pressure

☐ Humidity

Set Data Logging Rate

Mins: 1

☐ Is the interval OK?

Fig. 4.3: Weather data acquisition user interface

4.4 PERFORMANCE EVALUATION

In other to ascertain the effectiveness of the weather data acquisition system developed, the system was used to monitor atmospheric temperature, pressure and humidity around Physics department complex Federal University of Technology Akure. The readings obtained were compared with that recorded by an imported weather station installed within the complex. The results are presented in Tables 4.4, 4.5 & 4.6.

4.5 CONCLUSION AND RECOMMENDATION

It has been demonstrated that simultaneous measurement and documentation of continuously changing weather parameters such as Temperature, Pressure and humidity at high speed are quite possible. The advantages of this system over other form of weather monitoring systems are:

- The measurement and storage of data is fully automatic. There, is no need of periodic manual- retrieval of data. In many versions of the weather monitoring systems available, the data is manually retrieved and the system must be reset before it can take new readings.
- Since the data are directly imported to the computer, the large memory facilities available on the computer permits a long time uninterrupted access to data storage. Also the assessment and further processing of the data will be faster.

However, in as much as the objective of this project (see section 1.1) has been attained, I wish to make the following recommendations:

1. Since three quantities (temperature, pressure and humidity) were covered in this project, other parameters such as wind speed and solar intensity could be added in future construction
2. Instead of using cables to transmit signal from the measuring zone to the multiplexer, a radio wave carrier could be considered. The major advantage of this is that in long distance transmission, running multiple cables is expensive. Also this consideration could be advantageous in a centralised weather monitoring system.
3. Instead of the parallel form of transmission used between the ADC and the computer in this project, further work could look at the serial form of transmission.

Table 4.4: Temperature result of the designed equipment versus the result from a laboratory thermometer

Reading from Thermometer/°C	Result from the designed equipment/°C
25.00	24.80
26.00	26.20
27.20	27.00
28.00	28.12
29.80	29.68
30.00	30.18

Table 4.5: Comparison of pressure readings from the designed equipment and imported weather station

Pressure reading from Davis Vantage Pro(mbar)	Reading of pressure from the designed equipment(mbar)
968.352	968.168
969.082	968.952
969.821	969.627
971.081	970.882
971.640	971.501
971.642	971.504
971. 620	971.583
971.624	971.589

Table 4.6: Compared humidity result for the designed equipment.

Readings from Davis vantage Pro(%)	Reading of humidity from the designed equipment(%)
67.00	66.80
74.25	74.08
80.00	80.32
82.50	82.06
85.00	85.06
87.87	89.07
90.20	90.17
92.00	91.67

4. The inherent capability of a general purpose computer to data-log and process large amount of data at high speed and make the result of such computation available almost immediately presents a cheap means of achieving complex control at ease. This great potential should be further harnessed in future research areas that has to do with adaptability of various transducers (active or passive) to control, measurement and instrumentation purposes.
5. Future design should look at the possibility of the device having its own memory facility, instead of relying on the memory facility available on the computer.
6. In addition, future designs should look at the possibility of powering the device from a battery/solar-panel. This will facilitate the usage of the equipment when there is power failure and/or where electricity supply is not accessible.

REFERENCES

Adedayo, K.D., Babalola, M.T. and Ajewole, M.O. (2003): Design and construction of ground electrical conductivity meter. Zuma journal of pure and applied science. 5(1):77-82

Akpan, V. A. (2003)**:** Development of a model Automatic Vending Machine. Unpublished M Tech thesis submitted to Federal University of Technology, Akure

Analogue Devices, (2001): OP-07 Data and application note: pg 6-10

Barney, G.C. (1988):Intelligent Instrumentation – ***Microprocessor application in measurement and control.*** First edition. Prentice-Hall of India: pg233-255

Crecraft, D.I., Sparks, J.J. and Thompson, M., (1993): ***Electronics*** – Analogue and Digital. First edition. Chapman and Hall publishers, London, UK: pg 307-317, 639-648.

Darold, W. (1987): Electronic Design for Instrumentation. Second edition. Cambridge University press. Cambridge, UK: pg 87-120.

Franco, S. (1988): Designing with op-amp and analogue I.C. McGraw-Hill. USA: pg 66-75

Greg, P. (1998): Teach yourself Visual basic in 21 days. SAMS, USA: pg 1-555.

Hall, D.V. (1989): Digital Circuits and Systems. McGraw-Hill International edition. USA: pg 318-342.

Horowitz, P. and Hill, W., (1995): The art of electronics. Second edition. Cambridge University Press, Cambridge, UK: pg 189-211, 392-415.

Isaac, J., Thomas, S. and Philip, J., (1993): General purpose high performance temperature controller. *Int. Journal of electronics*. 74(6): 979-982

Nameroff S. (1989): Q-basic-*The complete reference*. Osborne McGraw-Hill. U.S.A.: pg 55-262

Odo, E.A. and M.T. Babalola, (2003): Computer Interfacing of a 12bit Analogue-to-Digital converter for temperature monitoring. Zuma journal of pure and applied science. 5(1):83-88

Usher, M.J. (1996): Sensors and Transducers. Second edition. Macmillan publishers, London, UK: pg 40-80

Appendix A

AD590JH Temperature Sensor Application Note

RS data

Semiconductor temperature sensor

Stock number 308-809

The RS590 semiconductor temperature sensor is functionally a two terminal I.C. which produces an output current proportional to absolute temperature. For supply voltages between +4V and +30V d.c. the device acts as a high impedance constant current regulator passing μA per degree Kelvin. Linearisation circuitry, precision voltage amplifiers, resistance measuring circuitry or cold junction compensation are not required for basic temperature measurement.

The RS590 is ideal in remote sensing applications. The device is virtually insensitive to voltage drops over long lines due to its high impedance current output provided the connection cable used is a twisted pair and well insulated.

Specification

Typical at +25°C (298.2°K) and V_S = 5V unless otherwise stated.

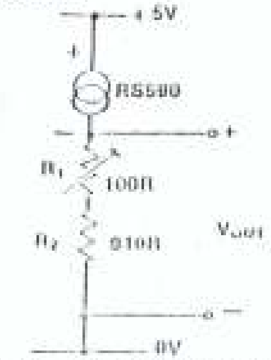
Absolute maximum ratings

Forward voltage (+ to −)	+44V
Reverse voltage (+ to −)	−20V
Breakdown voltage (Case to + or −)	±200V
Rated performance temperature range	−55°C to +150°C
Storage temperature range	−65°C to +175°C
Lead temperature (soldering, 10 sec)	+300°C

Power supply

Operating voltage range	+4V to +30V

Figure 1 **Calibration error trim at single point temperature**



To trim the above circuit, the temperature of the RS590 is measured by a reference temperature sensor and R_1 is trimmed so that V_{out} = 1mV/°K at that temperature.

PIN CONNECTIONS

TO 52 Package

PIN VIEW

Output

Nominal current output	298.2μA
Nominal temperature coefficient	1μA/°C
Calibration error	±2.5°C max
Absolute error[2] (over rated performance temperature range)	
Without external calibration adjustment	±5.5°C max
With +25°C calibration error set to zero	±2.0°C max
Nonlinearity	±0.8°C max
Repeatability	±0.1°C max
Long term drift[1]	±0.1°C max
Current noise	40pA/$\sqrt{Hz}$
Power supply rejection	
$+4V \leq V_S \leq +5V$	0.5μA/V
$+5V \leq V_S \leq +15V$	0.2μA/V
$+15V \leq V_S \leq +30V$	0.1μA/V
Case isolation to either lead	10^{10}Ω
Effective shunt capacitance	100pF
Electrical Turn-on time[2]	20μs
Reverse bias leakage current[3] (Reverse voltage = 10V)	10pA

1 Conditions: constant +5V, constant +125°C
2 Does not include self heating effects
3 Leakage current doubles every 10°C

Operation

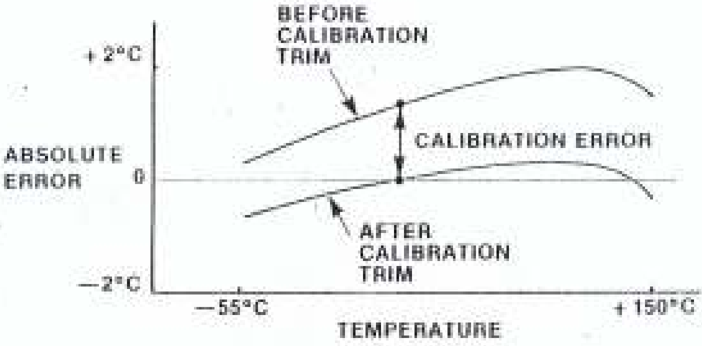
As previously stated the output of the RS590 is basically a proportional to absolute temperature (PTAT) current regulator i.e. the output is equal to a scale factor multiplied by the temperature of the sensor in degrees Kelvin.

Calibration error

The difference between the indicated temperature and actual temperature is called the calibration error. Since this is a scale factor error, it is relatively simple to trim out. Fig. 1 shows the most elementary way of accomplishing this.

Each RS590 is tested for error over the temperature ange with callibration error trimmed out. This error consists of a slope error and some curvature, mostly at the temperature extremes. Fig. 2 shows a typical temperature curve before and after calibration error trimming.

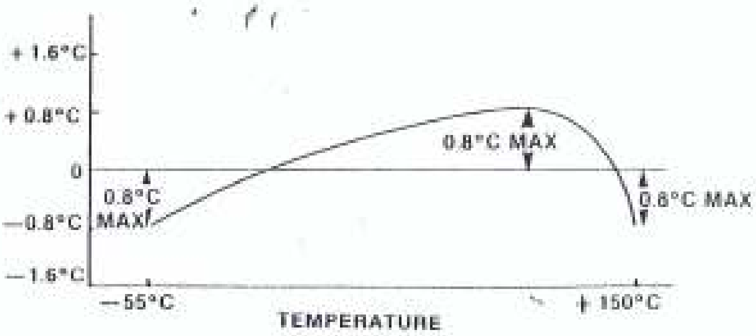
Figure 2 **Effect of calibration error trim on accuracy at single point temperature**



Nonlinearity

Nonlinearity as it applies to the RS590 is the aximum deviation of current over the entire temperature range from a best fit straight line. Fig. 3 shows the nonlinearity of the typical RS590 from Fig. 2.

Figure 3 **Nonlinearity**



ig 4A. shows a circuit in which nonlinearity is the ajor contribution to error over temperature. The rcuit is trimmed by adjusting R_1 for a 0V output ith RS590 at 0°C. R_2 is then adjusted for 10V out ith the sensor at 100°C. Other pairs of temperaures may be used with this procedure as long as ey are measured accurately by a reference sensor. Note that for +15V output (150°C) the V+ supply to the op-amp must be greater than 17V. Also note that V− should be at least −4V: if V− is ground there is no voltage applied across the device.

Note: Resistor values are typical and may need alteration depending upon magnitude of V−.

Appendix B

24PCCF6DA Pressure Sensor Application Note

Pressure Sensors
Absolute Unamplified Noncompensated

NEW

FEATURES

- Absolute pressure measurement
- Miniature package
- 2-15 and 2-30 psi pressure ranges
- 2 mA constant current excitation significantly reduces sensitivity shift over temperature*

24PC PERFORMANCE SPECIFICATIONS

Accuracy Specifications @ 10.0 ± .01 VDC Excitation, 25°C

Parameter	Range psia	bar	Min.	Typ.	Max.	Units
Excitation			—	10	12	VDC
Null Shift	2-15	1		±2.0	±4.0	mV
0 to 25°C, 25 to 50°C	2-30	2		±2.0	±5.5	
Linearity	2-15	1		.10	.20	% Span
B.F.S.L. P2 < P1**	2-30	2		.15	.30	
Sensitivity Shift 0 to 25°C, 25 to 50°C	All			±5.0	±6.5	% Span
Repeatability & Hysteresis	All			±0.5		% Span
Input Resistance			4.0 K	5.0 K	6.0 K	Ohms
Output Resistance			4.0 K	5.0 K	6.0 K	Ohms
Weight			—	2.0	—	grams

ENVIRONMENTAL SPECIFICATIONS

Operating Temperature	−40 to +85°C (−40 to +185°F)
Storage Temperature	−55 to +100°C (−67 to +212°F)
Shock	Qualification tested to 150 G
Vibration	Qualification tested to 0 to 2 kHz, 20 G sine
Media Compatibility	Limited only to those media which will not attack polyetherimide, silicon, fluorosilicone and silicone seals.

*Span: the algebraic difference between output end points

**B.F.S.L.: Best Fit Straight Line

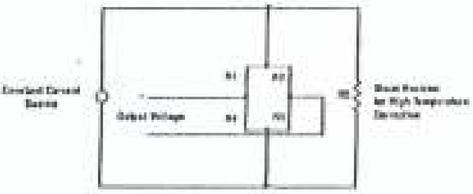
24PC ABSOLUTE ORDER GUIDE

Catalog Listing Type	Pressure Range psia	Span, mV			Null Offset mV			Sensitivity mV/psi Typ.	Over-pressure psia Typ.
		Min.	Typ.	Max.	Min.	Typ.	Max.		
24PCC	2-15	−140	−200	−260	−46	−16	+14	15	45
24PCD	2-30	−160	−300	−440	−61	−16	+29	11	60

*Non-compensated pressure sensors, excited by constant current instead of voltage, exhibit temperature compensation of Span. Application Note #1 briefly discusses current excitation.

Constant current excitation has an additional benefit of temperature measurement. When driven by a constant current source, a silicon pressure sensor's terminal voltage will rise with increased temperature. The rise in voltage not only compensates the Span, but is also an indication of die temperature.

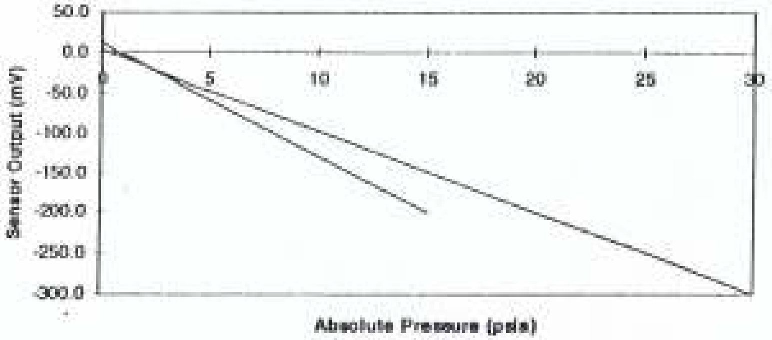
Constant Current Excitation Schematic



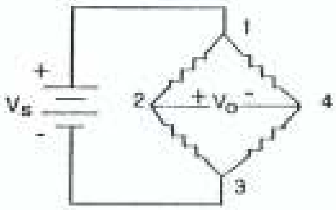
Pressure Sensors

Absolute Unamplified Noncompensated

24PC SERIES ABSOLUTE PRESSURE SENSOR OUTPUT CURVE



EXCITATION SCHEMATIC



TERMINATION STYLE

Style 6 - 1 x 4

Pin 1 = Vs (+)
Pin 2 = Output (+)
Pin 3 = Ground (–)
Pin 4 = Output (–)

Pin 1 is notched
Pin 2 is next to
Pin 1, etc.

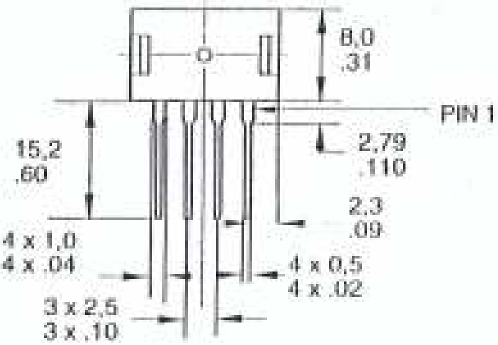
SENSOR SELECTION GUIDE

2 Product Family	4 Circuit Type	PC Pressure Transducer	C Pressure Range	F** Type of Seal	D* Type of Port (P1)	6 Termination Style	A Pressure Measurement
2 20PC Family	4 Standard noncompensated		C 2-15 psia 1 bar D 2-30 psia 2 bar	F Fluoro silicone	A Straight D Modular	6 1 x 4 (.600" long)	A Absolute

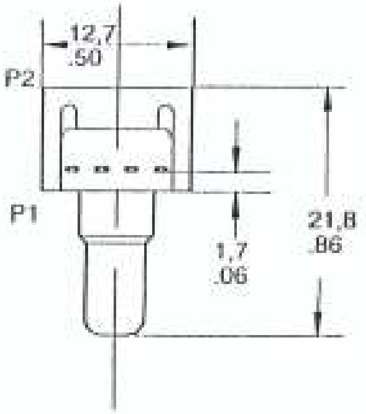
*Port type refers to P1
** Media seal is on P1 side and will not be in contact with media

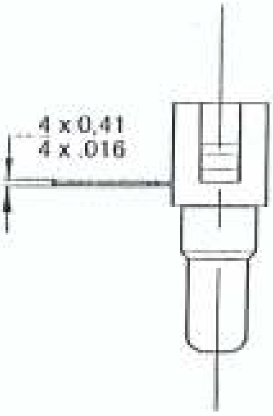
Example: 24PCCFD6A
Non-compensated 15 psi Absolute sensor with fluorosilicone seal, modular port, 1 x 4 terminals, .600" long.
See Accessory Guide, page 27.

MOUNTING DIMENSIONS (for reference only)

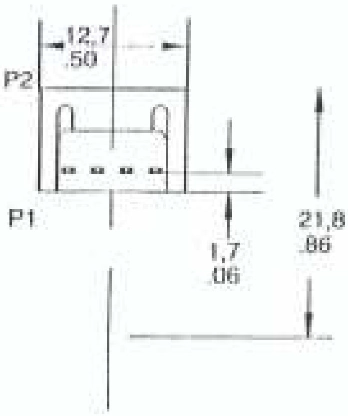


A Straight Port





D Modular Port



Appendix C

HIH-3610-001 Humidity Sensor Application Note

Humidity Sensors
Humidity Sensor

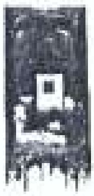
HIH-3610 Series

FEATURES

- Molded thermoset plastic housing with cover
- Linear voltage output vs %RH
- Laser trimmed interchangeability
- Low power design
- High accuracy
- Fast response time
- Stable, low drift performance
- Chemically resistant

TYPICAL APPLICATIONS

- Refrigeration
- Drying
- Metrology
- Battery-powered systems
- OEM assemblies



The HIH-3610 Series humidity sensor is designed specifically for high volume OEM (Original Equipment Manufacturer) users. Direct input to a controller or other device is made possible by this sensor's linear voltage output. With a typical current draw of only 200 µA, the HIH-3610 Series is ideally suited for low drain, battery operated systems. Tight sensor interchangeability reduces or eliminates OEM production calibration costs. Individual sensor calibration data is available.

The HIH-3610 Series delivers instrumentation-quality RH (Relative Humidity) sensing performance in a low cost, solderable SIP (Single In-line Package). Available in two lead spacing configurations, the RH sensor is a laser trimmed thermoset polymer capacitive sensing element with on-chip integrated signal conditioning. The sensing element's multilayer construction provides excellent resistance to application hazards such as wetting, dust, dirt, oils, and common environmental chemicals.

⚠WARNING

PERSONAL INJURY

- DO NOT USE these products as safety or emergency stop devices, or in any other application where failure of the product could result in personal injury.

Failure to comply with these instructions could result in death or serious injury.

⚠WARNING

MISUSE OF DOCUMENTATION

- The information presented in this product sheet is for reference only. Do not use this document as system installation information
- Complete installation, operation, and maintenance information is provided in the instructions supplied with each product.

Failure to comply with these instructions could result in death or serious injury.

TABLE 1: PERFORMANCE SPECIFICATIONS

Parameter	Condition
RH Accuracy[1]	±2% RH, 0-100% RH non-condensing, 25 °C, V_{supply} = 5 Vdc
RH Interchangeability	±5% RH, 0-60% RH; ±8% @ 90% RH typical
RH Linearity	±0.5% RH typical
RH Hysteresis	±1.2% RH span maximum
RH Repeatability	±0.5% RH
RH Response Time, 1/e	15 sec in slowly moving air at 25 °C
RH Stability	±1% RH typical at 50% RH in 5 years
Power Requirements	
Voltage Supply	4 Vdc to 5.8 Vdc, sensor calibrated at 5 Vdc
Current Supply	200 µA at 5 Vdc
Voltage Output	V_{out} = V_{supply} (0.0062(Sensor RH) + 0.16), typical @ 25 °C (Data printout option provides a similar, but sensor specific, equation at 25 °C)
V_{supply} = 5 Vdc	0.8 Vdc to 3.9 Vdc output @ 25 °C typical
Drive Limits	Push/pull symmetric; 50 µA typical, 20 µA minimum, 100 µA maximum Turn-on ≤ 0.1 sec
Temperature Compensation	True RH = (Sensor RH)/(1.093-0.0012T), T in °F True RH = (Sensor RH)/(1.0546-0.00216T), T in °C
Effect @ 0% RH	±0.007 %RH/°C (negligible)
Effect @ 100% RH	-0.22% RH/°C (<1% RH effect typical in occupied space systems above 15 °C (59 °F))
Humidity Range	
Operating	0 to 100% RH, non-condensing[1]
Storage	0 to 90% RH, non-condensing
Temperature Range	
Operating	-40 °C to 85 °C (-40 °F to 185 °F)
Storage	-51 °C to 125 °C (-60 °F to 257 °F)
Package[2]	Three pin, solderable SIP in molded thermoset plastic housing with thermoplastic cover
Handling	Static sensitive diode protected to 15 kV maximum

Notes:

1. Extended exposure to ≥90% RH causes a reversible shift of 3% RH.
2. This sensor is light sensitive. For best results, shield the sensor from bright light.

ESD SENSITIVITY:
CLASS 3

FACTORY CALIBRATION

HIH-3610 sensors may be ordered with a calibration and data printout (Table 2). See order guide on back page.

TABLE 2: EXAMPLE DATA PRINTOUT

Model	**HIH-3610-001**
Channel	92
Wafer	030996M
MRP	337313
Calculated values at 5 V V_{out} @ 0% RH V_{out} @ 75.3% RH	 0.958 V 3.268 V
Linear output for 2% RH accuracy @ 25 °C Zero offset Slope RH	 0.958 V 30.680 mV/%RH (V_{out}-zero offset)/slope (V_{out}-0.958)/0.0307
Ratiometric response for 0 to 100% RH V_{out}	 V_{supply} (0.1915 to 0.8130)

FIGURE 1: RH SENSOR CONSTRUCTION

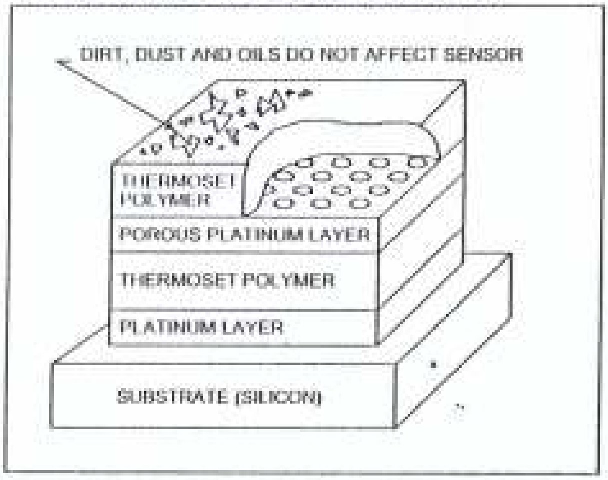


FIGURE 2: OUTPUT VOLTAGE VS RELATIVE HUMIDITY AT 0 °C

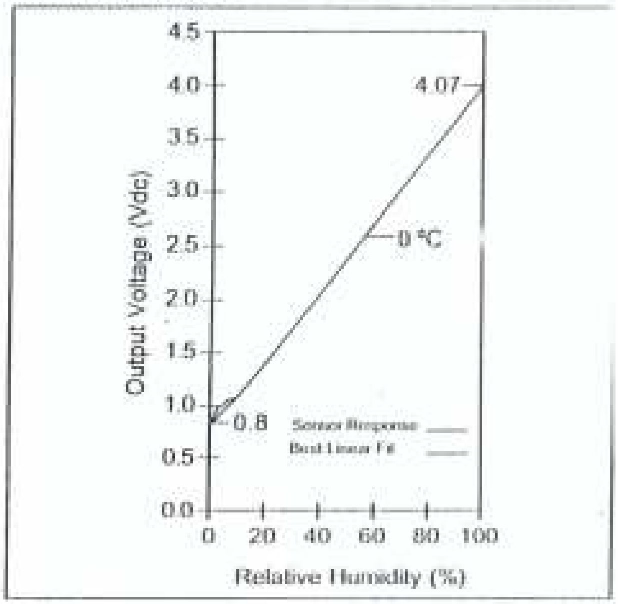
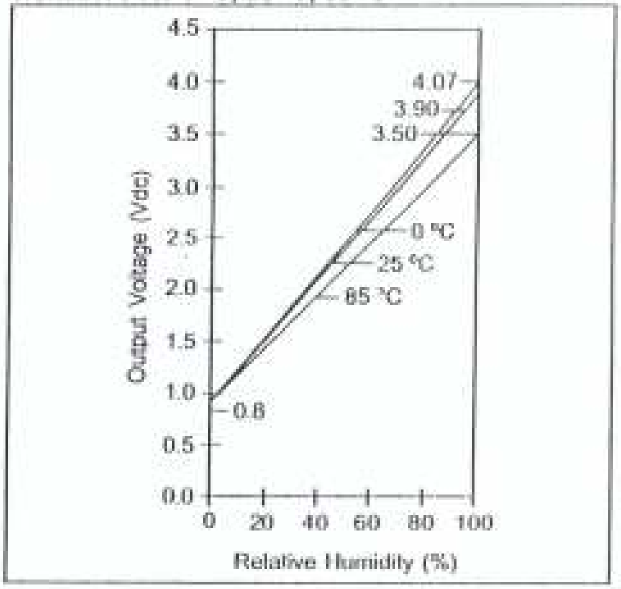


FIGURE 3: OUTPUT VOLTAGE VS RELATIVE HUMIDITY AT 0 °C, 25 °C, 85 °C



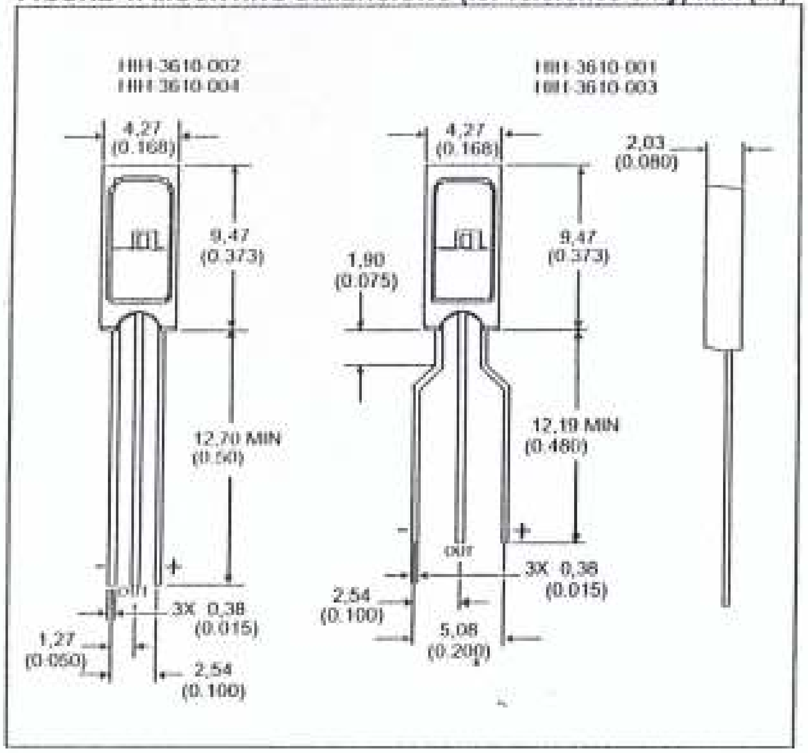
Humidity/Moisture Sensors
Humidity Sensor

HIH-3610 Series

ORDER GUIDE

Catalog Listing	Description
HIH-3610-001	Integrated circuit humidity sensor, 0.100 in lead pitch SIP
HIH-3610-002	Integrated circuit humidity sensor, 0.050 in lead pitch SIP
HIH-3610-003	Integrated circuit humidity sensor, 0.100 in lead pitch SIP with calibration and data printout
HIH-3610-004	Integrated circuit humidity sensor, 0.050 in lead pitch SIP with calibration and data printout

FIGURE 4: MOUNTING DIMENSIONS (for reference only) mm (in)



WARRANTY/REMEDY

Honeywell warrants goods of its manufacture as being free of defective materials and faulty workmanship. Contact your local sales office for warranty information. If warranted goods are returned to Honeywell during the period of coverage, Honeywell will repair or replace without charge those items it finds defective. The foregoing is Buyer's sole remedy and is **in lieu of all other warranties, expressed or implied, including those of merchantability and fitness for a particular purpose.**

Specifications may change without notice. The information we supply is believed to be accurate and reliable as of this printing. However, we assume no responsibility for its use.

While we provide application assistance personally, through our literature and the Honeywell web site, it is up to the customer to determine the suitability of the product in the application.

For application assistance, current specifications, or name of the nearest Authorized Distributor, check the Honeywell web site or call:

1-800-537-6945 USA
1-800-737-3360 Canada
1-815-235-6847 International

FAX

1-815-235-6545 USA

INTERNET

www.honeywell.com/sensing
info.sc@honeywell.com

Honeywell

Sensing and Control
Honeywell
11 West Spring Street
Freeport, Illinois 61032

www.honeywell.com/sensing

Printed with Soy Ink
on 50% Recycled Paper

[illegible] Printed in USA

Appendix D
TC7109A 12-bit ADC Application Note

MICROCHIP

12-BIT µP-Compatible Analog-To-Digital Converters

FEATURES

- Zero-Integrator Cycle for Fast Recovery From Input Overloads
- Eliminates Cross-Talk in Multiplexed Systems
- 12-Bit Plus Sign Integrating A/D Converter With Overrange Indication
- Sign Magnitude Coding Format
- True Differential Signal Input and Differential Reference Input
- Low Noise .... 15µV$_{P-P}$ Typ.
- Input Current .... 1pA Typ.
- No Zero Adjustment Needed
- TTL-Compatible, Byte-Organized Tri-State Outputs
- UART Handshake Mode for Simple Serial Data Transmission

ORDERING INFORMATION

PART CODE TC7109X

A or blank*

Package Code	Package	Temperature Range
CKW	44-PQFP	0°C to +70°C
CLW	44-Pin PLCC	0°C to +70°C
CPL	40-Pin Plastic DIP	0°C to +70°C
IJL	40-Pin CerDIP	− 25°C to +85°C

* The "A" version has a higher I_{OUT} on the digital lines.

GENERAL DESCRIPTION

The TC7109A is a 12-bit plus sign, CMOS low-powe
analog-to-digital converter (ADC). Only eight passive com
ponents and a crystal are required to form a complet
dual-slope integrating ADC.

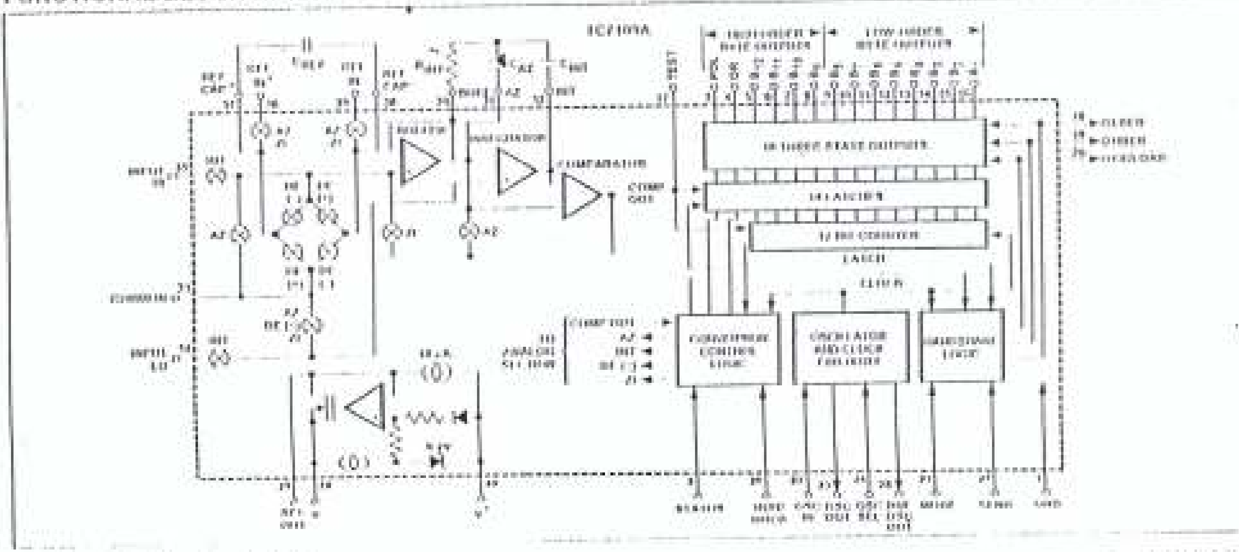
The improved V_{OH} source current TC7109A has fea
tures that make it an attractive per-channel alternative t
analog multiplexing for many data acquisition applica
tions. These features include typical input bias current o
1pA, drift of less than 1µV/°C, input noise typically 15µV
P, and auto-zero. True differential input and referenc
allow measurement of bridge-type transducers such a
load cells, strain gauges, and temperature transducers.

The TC7109A provides a versatile digital interface. I
the direct mode, chip select and HIGH/LOW byte enabl
control parallel bus interface. In the handshake mode, th
TC7109A will operate with industry standard UARTs i
controlling serial data transmission — ideal for remot
data logging. Control and monitoring of conversion timin
is provided by the RUN/$\overline{\text{HOLD}}$ input and STATUS outpu

For applications requiring more resolution, see th
TC500, 15 bit plus sign ADC data sheet.

The TC7109A has improved overrange recovery pe
formance and higher output drive capability than the orig
nal TC7109. All new (or existing) designs should speci
the TC7109A wherever possible.

FUNCTIONAL BLOCK DIAGRAM



ABSOLUTE MAXIMUM RATINGS*

Positive Supply Voltage (GND to V⁺)	+6.2V
Negative Supply voltage (GND to V⁻)	– 9V
Analog Input Voltage (Low to High) (Note 1)	V⁺ to V⁻
Reference Input Voltage (Low to High (Note 1)	V⁺ to V⁻
Digital Input Voltage (Pins 2–27) (Note 2)	GND – 0.3V
Power Dissipation, $T_A \leq 70°C$, (Note 3)	
CerDIP	2.29W
Plastic DIP	1.23W
PLCC	1.23W
PQFP	1.00W
Operating Temperature Range	
Plastic Package (C)	0°C to +70°C
Ceramic Package (I)	– 25°C to +85°C
(M)	– 55°C to +125°C
Storage Temperature Range	– 65°C to +150°C
Lead Temperature (Soldering, 10 sec)	+300°C

*Static-sensitive device. Unused devices must be stored in conductive material. Protect devices from static discharge and static fields. Stresses above those listed under Absolute Maximum Ratings may cause permanent damage to the device. These are stress ratings only and functional operation of the device at these or any other conditions above those indicated in the operational sections of the specifications is not implied. Exposure to Absolute Maximum Rating Conditions for extended periods may affect device reliability.

NOTES:
1. Input voltages may exceed supply voltages if input current is limited to ±100µA.
2. Connecting any digital inputs or outputs to voltages greater than V⁺ or less than GND may cause destructive device latch-up. Therefore, it is recommended that inputs from sources other than the same power supply should not be applied to the TC7109A before its power supply is established. In multiple supply systems, the supply to the device should be activated first.
3. This limit refers to that of the package and will not occur during normal operation.

ELECTRICAL CHARACTERISTICS:

All parameters with V⁺ = +5V, V⁻ = –5V, GND = 0V, T_A = +25°C, unless otherwise indicated.

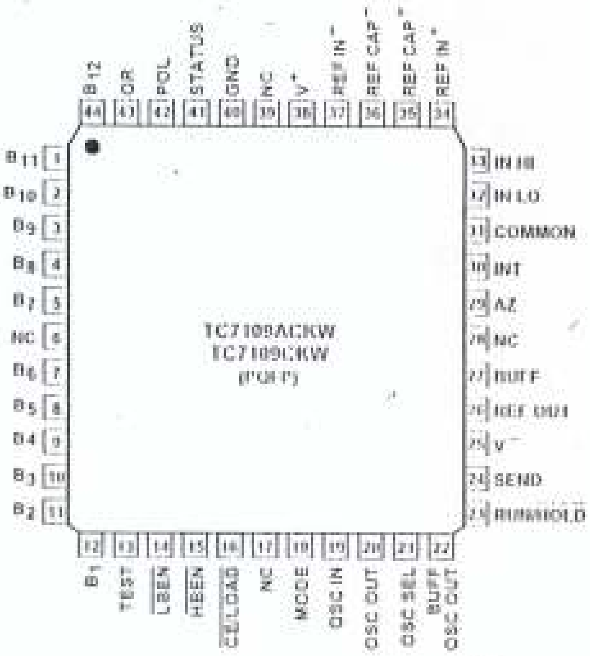
Symbol	Parameter	Test Conditions	Min	Typ	Max	Unit
Analog						
	Overload Recovery Time (TC7109A)		—	0	1	Measurement Cycle
	Zero Input Reading	V_{IN} = 0V Full Scale = 409.6mV	-0000_8	$\pm0000_8$	$+0000_8$	Octal Reading
	Ratio Metric Reading	$V_{IN} = V_{REF}$ V_{REF} = 204.8mV	3777_8	3777_8 4000_8	4000_8	Octal Reading
NL	Nonlinearity (Max Deviation From Best Straight Line Fit)	Full Scale = 409.6mV to 2.048V Over Full Operating Temperature Range	–1	±0.2	+1	Count
	Roll-Over Error (Difference in Reading for Equal Positive and Negative Inputs Near (Full Scale)	Full Scale = 409.6mV to 2.048V Over Full Operating Temperature Range	–1	±0.02	+1	Count
CMRR	Input Common-Mode Rejection Ratio	V_{CM} ±1V, V_{IN} = 0V Full Scale = 409.6mV	—	50	—	µV/V
V_{CMR}	Common Mode Voltage Range	Input High, Input Low, and Common Pins	V⁻+1.5	—	V⁺–1	V
e_N	Noise (P-P Value Not Exceeded 95% of Time)	V_{IN} = 0V Full Scale = 409.6mV	—	15	—	µV
I_{IN}	Leakage Current at Input	V_{IN}, All Packages: +25°C C Device: 0°C ≤ T_A ≤ +70°C I Device: –25°C ≤ T_A ≤ +85°C M Device: –55°C ≤ T_A ≤ +125°C	—	1 20 100 2	10 100 250 5	pA pA pA nA
TC_{ZS}	Zero Reading Drift	V_{IN} = 0V	—	0.2	1	µV/°C
TC_{FS}	Scale-Factor Temperature Coefficient	V_{IN} = 408.9mV = >7770_8 Reading, Ext Ref = 0ppm/°C	—	1	5	µV/°C
I⁺	Supply Current (V⁺ to GND)	V_{IN} = 0V, Crystal Oscillator 3.58MHz Test Circuit	—	700	1500	µA
I_S	Supply Current (V⁺ to V⁻)	Pins 2–21, 25, 26, 27, 29 Open	—	700	1500	µA

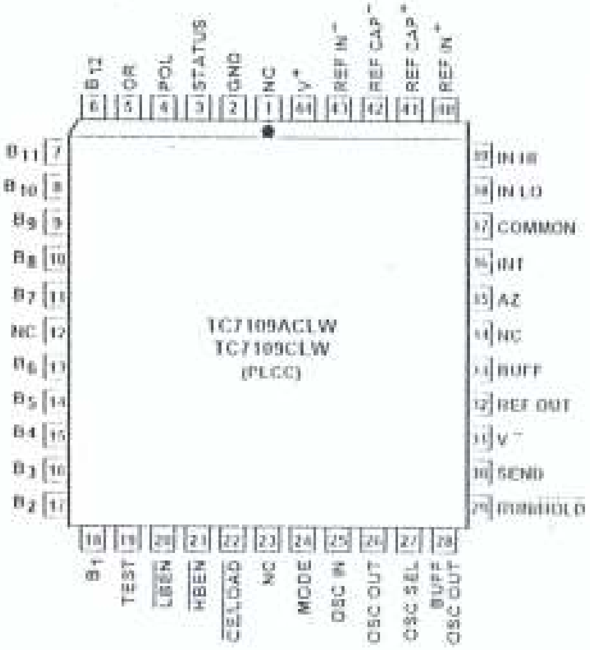
ELECTRICAL CHARACTERISTICS (Cont.)

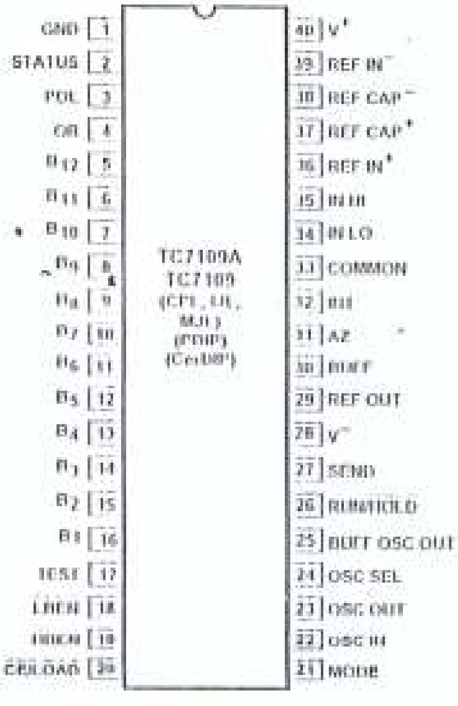
Symbol	Parameter	Test Conditions	Min	Typ	Max	Unit
V_{REF}	Ref Out Voltage	Referenced to V^+, 25kΩ Between V^+ and Ref Out	−2.4	−2.8	−3.2	V
TC_{REF}	Ref Out Temperature Coefficient	25kΩ Between V^+ and Ref Out $0°C \le T_A \le +70°C$	—	80	—	ppm/°C
Digital						
V_{OH}	Output High Voltage TC7109A: I_{OUT} = 700μA	TC7109: I_{OUT} = 100μA Pins 3 – 16, 18, 19, 20	3.5	4.3	—	V
V_{OL}	Output Low Voltage	I_{OUT} = 1.6mA	—	0.2	0.4	V
	Output Leakage Current	Pins 3 – 16 High Impedance	—	±0.01	±1	μA
	Control I/O Pull Up Current	Pins 18, 19, 20 $V_{OUT} = V^+ - 3V$ Mode Input at GND	—	5	—	μA
	Control I/O Loading	$\overline{HBEN}$, Pin 19, $\overline{LBEN}$, Pin 18	—	—	50	pF
V_{IH}	Input High Voltage	Pins 18 – 21, 26, 27 Referenced to GND	2.5	—	—	V
V_{IL}	Input Low Voltage	Pins 18 – 21, 26, 27 Referenced to GND	—	—	1	V
	Input Pull Up Current	Pins 26, 27; $V_{OUT} = V^+ - 3V$ Pins 17, 24, $V_{OUT} = V^+ - 3V$	—	5 25	—	μA μA
	Input Pull Down Current	Pin 21; V_{OUT} = GND + 3V	—	1	—	μA
	Oscillator Output Current, High	V_{OUT} = 2.5V	—	1	—	mA
	Oscillator Output Current, Low	V_{OUT} = 2.5V	—	1.5	—	mA
	Buffered Oscillator Output Current, High	V_{OUT} = 2.5V	—	2	—	mA
	Buffered Oscillator Output Current, Low	V_{OUT} = 2.5V	—	5		mA
t_W	Mode Input Pulse Width		60	—	—	nsec

HANDLING PRECAUTIONS: These devices are CMOS and must be handled correctly to prevent damage. Package and store only in conductive foam, antistatic tubes, or other conducting material. Use proper antistatic handling procedures. Do not connect in circuits under "power-on" conditions, as high transients may cause permanent damage.

PIN CONFIGURATIONS







NC = NO INTERNAL CONNECTION

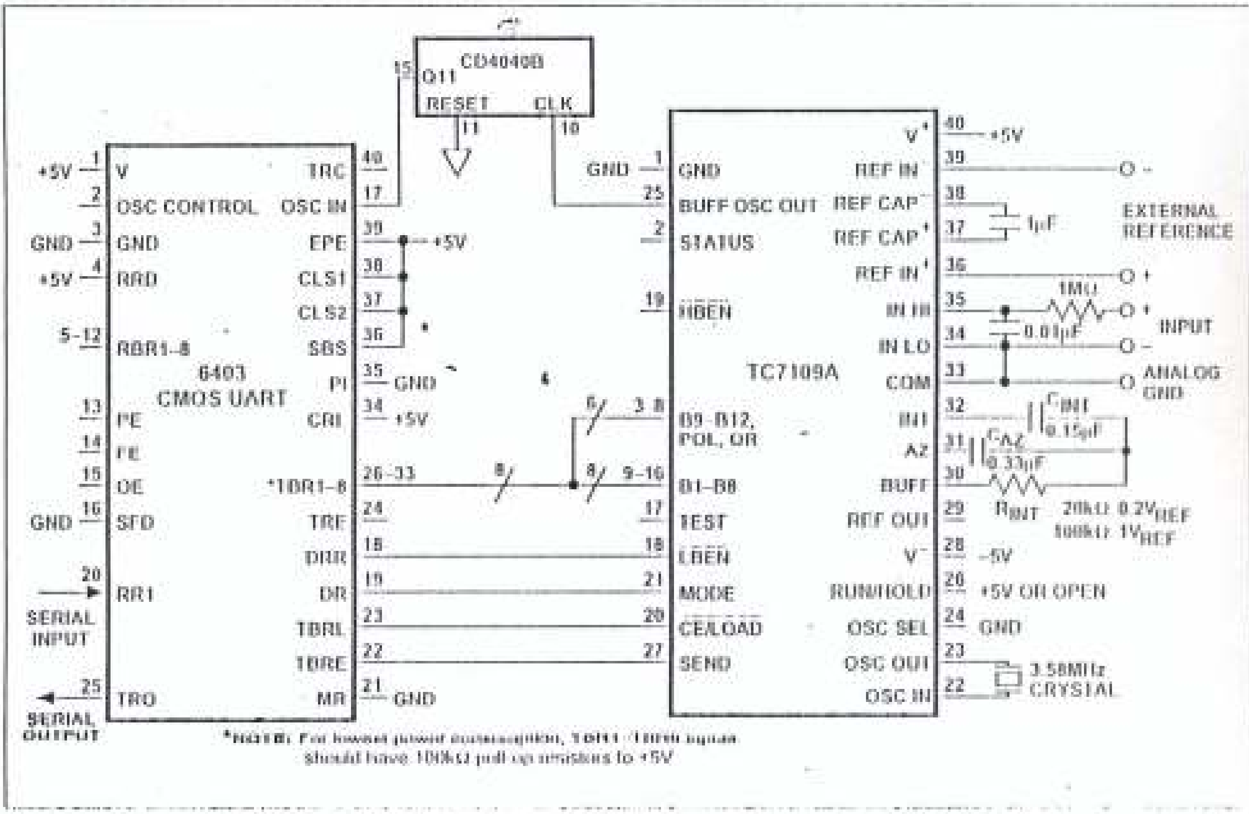
TC7109/A PIN DESCRIPTION

Pin No. (40-Pin PDIP)	Symbol	Description	
1	GND	Digital ground, 0V, ground return for all digital logic.	
2	STATUS	Output HIGH during integrate and deintegrate until data is latched. Output LOW when analog section is in auto-zero or zero-integrator configuration.	
3	POL	Polarity — High for positive input.	
4	OR	Overrange — High if overranged.	All Three-State Data Bits
5	B_{12}	Bit 12 (Most Significant Bit)	
6	B_{11}	Bit 11	
7	B_{10}	Bit 10	
8	B_9	Bit 9	
9	B_8	Bit 8	
10	B_7	Bit 7	
11	B_6	Bit 6	
12	B_5	Bit 5	
13	B_4	Bit 4	
14	B_3	Bit 3	
15	B_2	Bit 2	
16	B_1	Bit 1 (Least Significant Bit)	
17	TEST	Input High — Normal operation. Input LOW — Forces all bit outputs HIGH. **Note:** This input is used for test purposes only.	
18	$\overline{\text{LBEN}}$	Low Byte Enable — With MODE (Pin 21) LOW, and $\overline{\text{CE/LOAD}}$ (Pin 20) LOW, taking this pin LOW activates low order byte outputs, B1-B8. With MODE (Pin 21) HIGH, this pin serves as low byte flag output used in handshake mode. See Figures 7, 8, and 9.	
19	$\overline{\text{HBEN}}$	High Byte Enable — With MODE (Pin 21) LOW, and $\overline{\text{CE/LOAD}}$ (Pin 20) LOW, taking this pin LOW activates high order byte outputs, B9-B12, POL, OR. With MODE (Pin 21) HIGH, this pin serves as high byte flag output used in handshake mode. See Figures 7, 8, and 9.	
20	$\overline{\text{CE/LOAD}}$	Chip Enable/Load — With MODE (Pin 21) LOW, $\overline{\text{CE/LOAD}}$ serves as a master output enable. When HIGH, B1-B12, POL, OR outputs are disabled. When MODE (Pin 21) is HIGH, a load strobe is used in handshake mode. See Figure 7, 8, and 9.	
21	MODE	Input LOW — Direct output mode where $\overline{\text{CE/LOAD}}$ (Pin 20), $\overline{\text{HBEN}}$ (Pin 19), and $\overline{\text{LBEN}}$ (Pin 18) act as inputs directly controlling byte outputs. Input Pulsed HIGH — Causes immediate entry into handshake mode and output of data as in Figure 9. Input HIGH — Enables CE/LOAD (Pin 20), $\overline{\text{HBEN}}$ (Pin 19), and $\overline{\text{LBEN}}$ (Pin 18) as outputs, handshake mode will be entered and data output as in Figures 7 and 8 at conversions completion.	
22	OSC IN	Oscillator Input.	
23	OSC OUT	Oscillator Output.	
24	OSC SEL	Oscillator Select — Input HIGH configures OSC IN, OSC OUT, BUF OSC OUT as RC oscillator — clock will be same phase and duty cycle as BUF OSC OUT. Input LOW configures OSC IN, OSC OUT for crystal oscillator — clock frequency will be 1/58 of frequency at BUF OSC OUT.	
25	BUF OSC OUT	Buffered Oscillator Output	
26	$\overline{\text{RUN/HOLD}}$	Input HIGH — Conversions continuously performed every 8192 clock pulses. Input LOW — Conversion in progress completed; converter will stop in auto-zero seven counts before integrate.	

TC7109/A PIN DESCRIPTION (Cont.)

Pin No. (40-Pin PDIP)	Symbol	Description
27	SEND	Input — Used in handshake mode to indicate ability of an external device to accept data. Connect to V^+ if not used.
28	V^-	Analog Negative Supply — Nominally – 5V with respect to GND (Pin 1).
29	REF OUT	Reference Voltage Output — Nominally 2.8V down from V^+ (Pin 40).
30	BUFFER	Buffer Amplifier Output.
31	AUTO-ZERO	Auto-Zero Node — Inside foil of C_{AZ}.
32	INTEGRATOR	Integrator Output — Outside foil of C_{INT}.
33	COMMON	Analog Common — System is auto-zeroed to COMMON.
34	INPUT LOW	Differential Input Low Side.
35	INPUT HIGH	Differential Input High Side.
36	REF IN$^+$	Differential Reference Input Positive.
37	REF CAP$^+$	Reference Capacitor Positive.
38	REF CAP$^-$	Reference Capacitor Negative.
39	REF IN$^-$	Differential Reference Input Negative.
40	V^+	Positive Supply Voltage — Nominally +5V with respect to GND (Pin 1).

NOTE: All digital levels are positive true.



*NOTE: For lowest power consumption, TBR1 - TBR8 inputs should have 100kΩ pull up resistors to +5V

Figure 1. TC7109A UART Interface (Send Any Word to UART to Transmit Latest Result)

Zero-Integrator Phase

The ZI phase only occurs when an input overrange condition exists. The function of the ZI phase is to eliminate residual charge on the integrator capacitor after an overrange measurement. Unless removed, the residual charge will be transferred to the auto-zero capacitor and cause an error in the succeeding conversion.

The ZI phase virtually eliminates hysteresis or "cross talk" in multiplexed systems. An overrange input on one channel will not cause an error on the next channel measured. This feature is especially useful in thermocouple measurements, where unused (or broken thermocouple) inputs are pulled to the positive supply rail.

During ZI, the reference capacitor is charged to the reference voltage. The signal inputs are disconnected from the buffer and integrator. The comparator output is connected to the buffer input, causing the integrator output to be driven rapidly to 0V (Figure 3). The ZI phase only occurs following an overrange and lasts for a maximum of 1024 clock periods.

Differential Input

The TC7109A has been optimized for operation with analog common near digital ground. With +5V and –5V power supplies, a full ±4V full scale integrator swing maximizes the analog section's performance.

A typical CMRR of 86dB is achieved for input differential voltages anywhere within the typical common mode range of 1V below the positive supply to 1.5V above the negative supply. However, for optimum performance, the IN HI and IN LO inputs should not come within 2V of either supply rail. Since the integrator also swings with the common-mode voltage, care must be exercised to ensure the integrator output does not saturate. A worst case condition is near a full-scale negative differential input voltage with a large positive common-mode voltage. The negative input signal drives the integrator positive when most of its swing has been used up by the positive common-mode voltage. In such cases, the integrator swing can be reduced to less than the recommended ±4V full-scale value, with some loss of accuracy. The integrator output can swing to within 0.3V of either supply without loss of linearity.

Differential Reference

The reference voltage can be generated anywhere within the power supply voltage of the converter. Roll-over voltage is the main source of common mode error, caused by the reference capacitor losing or gaining charge due to stray capacity on its nodes. With a large common-mode voltage, the reference capacitor can gain charge (increase voltage) when called upon to deintegrate a positive signal and lose charge (decrease voltage) when called upon to deintegrate a negative input signal. This difference in reference for (+) or (–) input voltages will cause a roll-over error. This error can be held to less than 0.5 count worst case by using a large reference capacitor in comparison to the stray capacitance. To minimize roll over error from these sources, keep the reference common-mode voltage near or at analog common.

Digital Section

The digital section is shown in the block diagram (Figure 4) and includes the clock oscillator and scaling circuit, a 12-bit binary counter with output latches and TTL compatible three-state output drivers, UART handshake logic, polarity, overrange, and control logic. Logic levels are referred to as LOW or HIGH.

Inputs driven from TTL gates should have 3kΩ to 5kΩ pull-up resistors added for maximum noise immunity. For minimum power consumption, all inputs should swing from GND (LOW) to V⁺ (HIGH).

STATUS Output

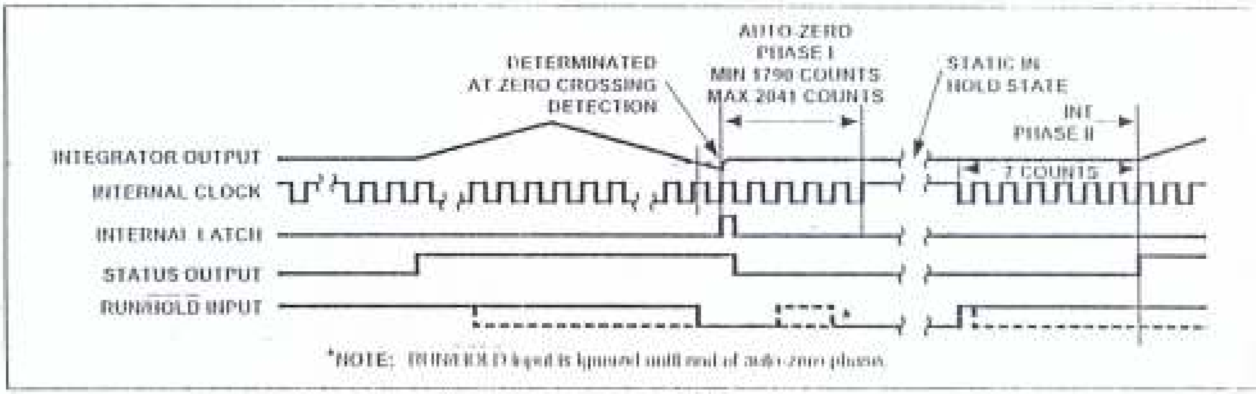
During a conversion cycle, the STATUS output goes HIGH at the beginning of signal integrate and goes LOW one-half clock period after new data from the conversion has been stored in the output latches (see Figure 3). The signal may be used as a "data valid" flag to drive interrupts, or for monitoring the status of the converter. (Data will not change while status is LOW.)

MODE Input

The output mode of the converter is controlled by the MODE input. The converter is in its "direct" output mode, when the MODE input is LOW or left open. The output data is directly accessible under the control of the chip and byte enable inputs (this input is provided with a pull-down resistor to ensure a LOW Level when the pin is left open). When the MODE input is pulsed high, the converter enters the UART handshake mode and outputs the data in 2 bytes, then returns to "direct" mode. When the MODE input is kept HIGH, the converter will output data in the handshake mode at the end of every conversion cycle. With MODE = 0 (direct bus transfer), the send input should be tied to V⁺. (See "Handshake Mode.")

RUN/$\overline{\text{HOLD}}$ Input

With the RUN/$\overline{\text{HOLD}}$ input high, or open, the circuit operates normally as a dual-slope ADC, as shown in Figure 3. Conversion cycles operate continuously with the output latches updated after zero crossing in the deintegrate mode. An internal pull-up resistor is provided to ensure a HIGH level with an open input.



*NOTE: RUN/HOLD input is ignored until end of auto-zero phase.

Figure 5. TC7109A RUN/HOLD Operation

The RUN/HOLD input may be used to shorten conversion time. If RUN/HOLD goes LOW any time after zero crossing in the deintegrate mode, the circuit will jump to auto-zero and eliminate that portion of time normally spent in deintegrate.

If RUN/HOLD stays or goes LOW, the conversion will complete with minimum time in deintegrate. It will stay in auto-zero for the minimum time and wait in auto-zero for a HIGH at the RUN/HOLD input. As shown in Figure 5, the STATUS output will go HIGH 7 clock periods after RUN/HOLD is changed to HIGH, and the converter will begin the integrate phase of the next conversion.

The RUN/HOLD input allows controlled conversion interface. The converter may be held at idle in auto-zero with RUN/HOLD LOW. The conversion is started when RUN/HOLD goes HIGH, and the new data is valid when the STATUS output goes LOW (or is transferred to the UART; see "Handshake Mode"). RUN/HOLD may now go LOW, terminating deintegrate and ensuring a minimum auto-zero time before stopping to wait for the next conversion. Conversion time can be minimized by ensuring RUN/HOLD goes LOW during deintegrate, after zero crossing, and goes HIGH after the hold point is reached. The required activity on the RUN/HOLD input can be provided by connecting it to the buffered oscillator output. In this mode, the input value measured determines the conversion time.

Direct Mode

The data outputs (bits 1 through 8, low-order bytes; bits 9 through 12, polarity and overrange high-order bytes) are accessible under control of the byte and chip enable terminals as inputs with the MODE pin at a LOW level. These three inputs are all active LOW. Internal pull-up resistors are provided for an inactive HIGH level when left open. When chip enable is LOW, a byte-enable input LOW will allow the outputs of the byte to become active. A variety of parallel data accessing techniques may be used, as shown in the "Interfacing" section. (See Figure 6 and Table 1.)

The access of data should be synchronized with the conversion cycle by monitoring the STATUS output. This prevents accessing data while it is being updated and eliminates the acquisition of erroneous data.

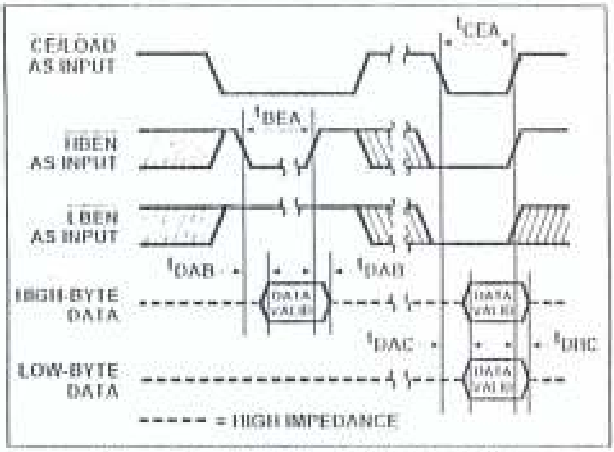


Figure 6. TC7109A Direct Mode Output Timing

Table 1. TC7109A Direct Mode Timing Requirements

Symbol	Description	Min	Typ	Max	Units
t_{BEA}	Byte Enable Width	200	500		nsec
t_{DAB}	Data Access Time From Byte Enable		150	300	nsec
t_{DHB}	Data Hold Time From Byte Enable		150	300	nsec
t_{CEA}	Chip Enable Width	300	500		nsec
t_{DAC}	Data Access Time From Chip Enable		200	400	nsec
t_{DHC}	Data Hold Time From Chip Enable		200	400	nsec

Handshake Mode

An alternative means of interfacing the TC7109A to digital systems is provided when the handshake output mode of the TC7109A becomes active in controlling the flow of data instead of passively responding to chip and byte enable inputs. This mode allows a direct interface between the TC7109A and industry-standard UARTs with no external logic required. The TC7109A provides all the control and flag signals necessary to sequence the two bytes of data into the UART and initiate their transmission in serial form when triggered into the handshake mode. The cost of designing remote data acquisition stations is reduced using serial data transmission to minimize the number of lines to the central controlling processor.

The MODE input controls the handshake mode. When the MODE input is held HIGH, the TC7109A enters the handshake mode after new data has been stored in the output latches at the end of every conversion performed (see Figures 7 and 8). Entry into the handshake mode may be triggered on demand by the MODE input. At any time during the conversion cycle, the LOW to HIGH transition of a short pulse at the MODE input will cause immediate entry into the handshake mode. If this pulse occurs while new data is being stored, the entry into handshake mode is delayed until the data is stable. The MODE input is ignored in the handshake mode, and until the converter completes the output cycle and clears the handshake mode, data updating will be inhibited (see Figure 9).

When the MODE input is HIGH or when the converter enters the handshake mode, the chip and byte enable inputs become TTL-compatible outputs which provide the output cycle control signals (see Figures 7, 8 and 9).

The SEND input is used by the converter as an indication of the ability of the receiving device (such as a UART) to accept data in the handshake mode. The sequence of the output cycle with SEND held HIGH is shown in Figure 7. The handshake mode (internal MODE HIGH) is entered after the data latch pulse (the $\overline{\text{CE/LOAD}}$, $\overline{\text{LBEN}}$ and $\overline{\text{HBEN}}$ terminals are active as outputs since MODE remains HIGH).

The HIGH level at the SEND input is sensed on the same HIGH-to-LOW internal clock edge. On the next LOW-to-HIGH internal clock edge, the high-order byte (bits 9 through 12, POL, and OR) outputs are enabled and the $\overline{\text{CE/LOAD}}$ and the $\overline{\text{HBEN}}$ outputs assume a LOW level. The $\overline{\text{CE/LOAD}}$ output remains LOW for one full internal clock period only; the data outputs remain active for 1-1/2 internal clock periods, and the high-byte enable remains LOW for 2 clock periods. The $\overline{\text{CE/LOAD}}$ output LOW level or LOW-to-HIGH edge may be used as a synchronizing signal to ensure valid data, and the byte enable as an output may be used as a byte identification flag. With SEND remaining HIGH the converter completes the output cycle using $\overline{\text{CE/LOAD}}$ and $\overline{\text{LBEN}}$ while the low-order byte outputs (bits 1 through 8) are activated. When both bytes are sent, the handshake mode is terminated. The typical UART interfacing timing is shown in Figure 8. The SEND input is used to delay portions of the sequence, or handshake, to ensure correct data transfer. This timing diagram shows an industry-standard HD6403 or CDP1854 CMOS UART to interface to serial data channels. The SEND input to the TC7109A is driven by the TBRE (Transmitter Buffer Register Empty) output of the UART, and the $\overline{\text{CE/LOAD}}$ input of the TC7109A drives the TBRL (Transmitter Buffer Register Load) input to the UART. The eight transmitter buffer register inputs accept the parallel data outputs. With the UART transmitter buffer register empty, the SEND input will be HIGH when the handshake mode is entered after new data is stored. The high-order byte outputs become active and the $\overline{\text{CE/LOAD}}$ and $\overline{\text{HBEN}}$ inputs will go LOW after SEND is sensed. When $\overline{\text{CE/LOAD}}$ goes HIGH at the end of one clock period, the high-order byte data is clocked into the UART transmitter buffer register. The UART TBRE output will go LOW, which halts the output cycle with the $\overline{\text{HBEN}}$ output LOW, and the high order byte outputs active. When the UART has transferred the data to the transmitter register and cleared the transmitter buffer register, the TBRE returns HIGH. The high-order byte outputs are disabled on the next TC7109A internal clock HIGH to LOW edge, and one-half internal clock later, the $\overline{\text{HBEN}}$ output returns HIGH. The $\overline{\text{CE/LOAD}}$ and LBEN outputs go LOW at the same time as the low-order byte outputs become active. When the $\overline{\text{CE/LOAD}}$ returns HIGH at the end of one clock period, the low-order data is clocked into the UART transmitter buffer register, and TBRE again goes LOW. The next TC7109A internal clock HIGH to LOW edge will sense when TBRE returns to a HIGH, disabling the data inputs. One half internal clock later, the handshake mode is cleared, and the $\overline{\text{CE/LOAD}}$, $\overline{\text{HBEN}}$ and $\overline{\text{LBEN}}$ terminals return HIGH and stay active, if MODE still remains HIGH.

Handshake output sequences may be performed on demand by triggering the converter into handshake mode with a LOW-to-HIGH edge on the MODE input. A handshake output sequence triggered is shown in Figure 9. The SEND input is LOW when the converter enters handshake mode. The whole output sequence is controlled by the SEND input, and the sequence for the first (high order) byte is similar to the sequence for the second byte.

Figure 9 also shows that the output sequence can take longer than a conversion cycle. New data will not be latched when the handshake mode is still in progress and is therefore lost.

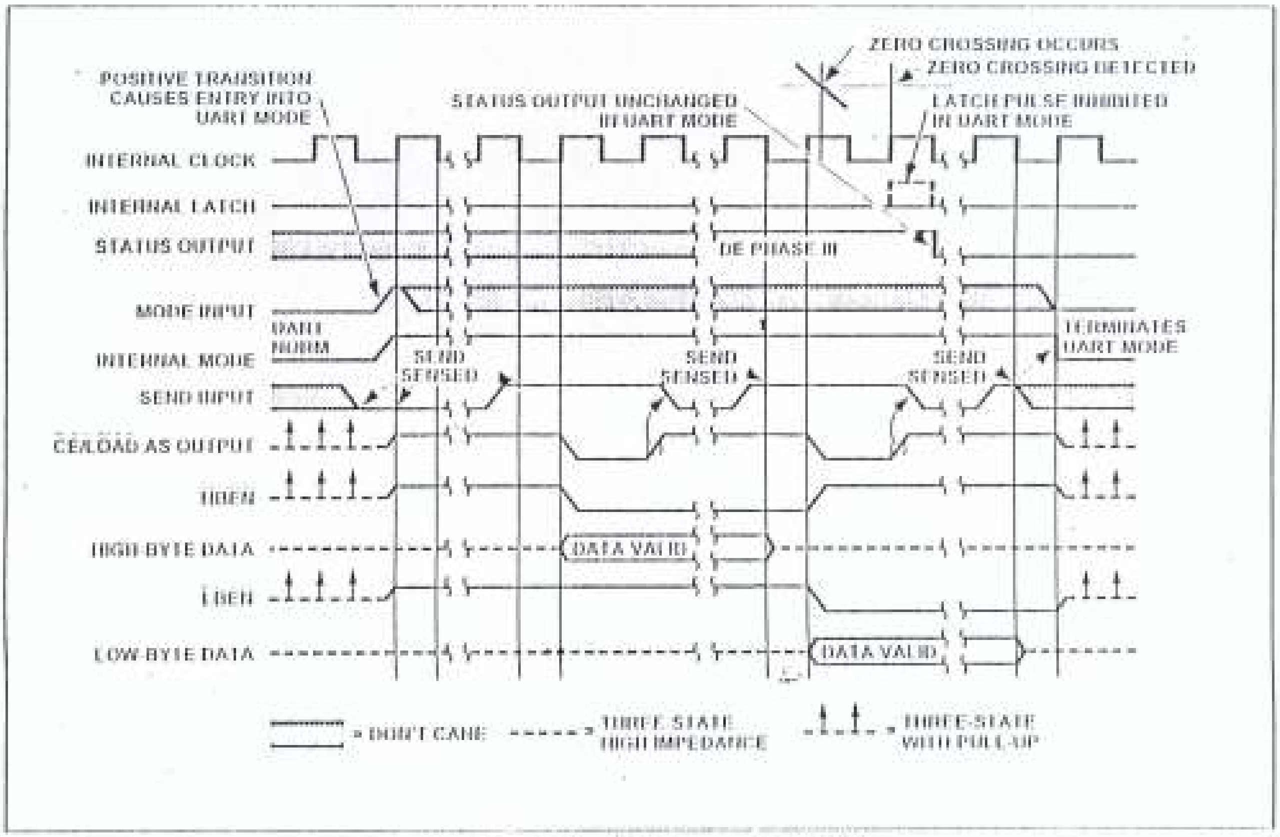


Figure 9. TC7109A Handshake Triggered by MODE Input

Oscillator

The oscillator may be overdriven, or may be operated as an RC or crystal oscillator. The OSCILLATOR SELECT input optimizes the internal configuration of the oscillator for RC or crystal operation. The OSCILLATOR SELECT input is provided with a pull-up resistor. When the OSCILLATOR SELECT input is HIGH or left open, the oscillator is configured for RC operation. The internal clock will be the same frequency and phase as the signal at the BUFFERED OSCILLATOR OUTPUT. Connect the resistor and capacitor as in Figure 10. The circuit will oscillate at a frequency given by f = 0.45/RC. A 100kΩ resistor is recommended for useful ranges of frequency. The capacitor value should be chosen such that 2048 clock periods are close to an integral multiple of the 60Hz period for optimum 60Hz line rejection.

With OSCILLATOR SELECT input LOW, two on-chip capacitors and a feedback device are added to the oscillator. In this configuration, the oscillator will operate with most crystals in the 1MHz to 5MHz range with no external components (Figure 11). The OSCILLATOR SELECT input LOW inserts a fixed ÷58 divider circuit between the BUFFERED OSCILLATOR OUTPUT and the internal clock. A 3.58MHz TV crystal gives a division ratio providing an integration time given by:

$$t = (2048 \text{ clock periods}) \frac{58}{3.58\text{MHz}} = 33.18\text{msec}$$

The error is less than 1% from two 60Hz periods, or 33.33msec, which will give better than 40dB, 60Hz rejection. The converter will operate reliably at conversion rates up to 30 per second, corresponding to a clock frequency of 245.8kHz.

When the oscillator is to be overdriven, the OSCILLATOR OUTPUT should be left open, and the overdriving signal should be applied at the OSCILLATOR INPUT. The internal clock will be of the same duty cycle, frequency and phase as the input signal. When the OSCILLATOR SELECT is at GND, the clock will be 1/58 of the input frequency.

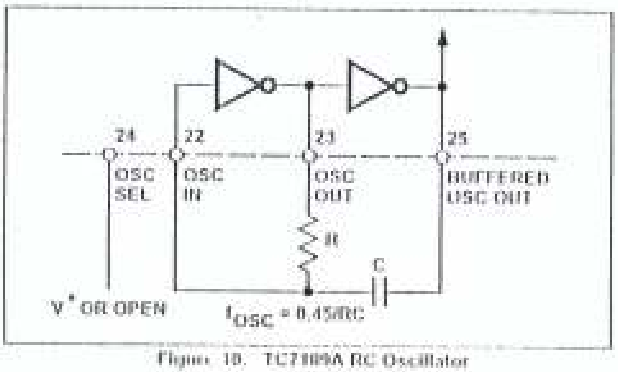


Figure 10. TC7109A RC Oscillator

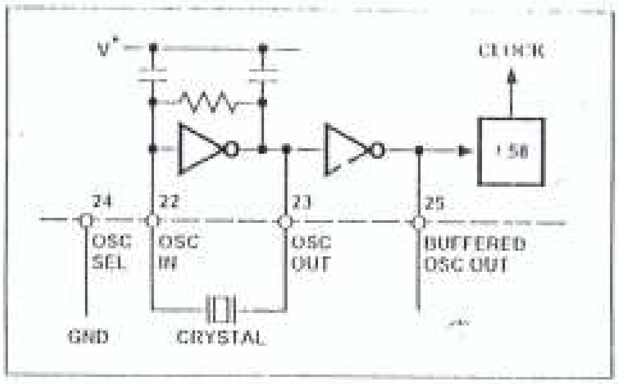


Figure 11. TC7109A Crystal Oscillator

Test Input

The counter and its outputs may be tested easily. When the TEST input is connected to GND, the internal clock is disabled and the counter outputs are all forced into the HIGH state. When the input returns to the 1/2 (V⁺ – GND) voltage or to V⁺ and one clock is input, the counter outputs will all be clocked to the LOW state.

The counter output latches are enabled when the TEST input is taken to a level halfway between V⁺ and GND, allowing the counter contents to be examined anytime.

Component Value Selection

The integrator output swing for full scale should be as large as possible. For example, with ±5V supplies and COMMON connected to GND, the nominal integrator output swing at full-scale is ±4V. Since the integrator output can go to 0.3V from either supply without significantly effecting linearity, a 4V integrator output swing allows 0.7V for variations in output swing due to component value and oscillator tolerances. With ±5V supplies and a common mode voltage range of ±1V required, the component values should be selected to provide ±3V integrator output swing. Noise and roll over errors will be slightly worse than in the ±4V cas For large common mode voltage ranges, the integrat output swing must be reduced further. This will increase bot noise and roll over errors. To improve performance, ±6 supplies may be used.

Integrating Capacitor

The integrating capacitor, C_{INT}, should be selected t give the maximum integrator output voltage swing that w not saturate the integrator to within 0.3V from either supply A ±3.5V to ±4V integrator output swing is nominal for th TC7109A, with ±5V supplies and analog common con nected to GND. For 7-1/2 conversions per second (61.7 kHz internal clock frequency), nominal values C_{INT} and C_A are 0.15μF and 0.33μF, respectively. These values shoul be changed if different clock frequencies are used to main tain the integrator output voltage swing. The value of C_{INT} i given by:

$$C_{INT} = \frac{(2048 \times \text{Clock Period})(20\mu A)}{\text{Integrator Output Voltage Swing}}$$

The integrating capacitor must have low dielectric ab sorption to prevent roll over errors. Polypropylene capaci tors give undetectable errors, at reasonable cost, up t +85°C. Teflon® capacitors are recommended for the militar temperature range. While their dielectric absorption charac teristics vary somewhat between units, devices may b selected to less than 0.5 count of error due to dielectri absorption.

Integrating Resistor

The integrator and buffer amplifiers have a class A output stage with 100μA of quiescent current. They suppl 20μA of drive current with negligible nonlinearity. The inte grating resistor should be large enough to remain in this very linear region over the input voltage range, but small enoug that undue leakage requirements are not placed on the PC board. For 2.048V full-scale a 100kΩ resistor is recom mended and for 409.6mV full-scale a 20kΩ resistor is recom mended. R_{INT} may be selected for other values of full scale by:

$$R_{INT} = \frac{\text{Full-Scale Voltage}}{20\mu A}$$

Auto-Zero Capacitor

As the auto-zero capacitor is made large, the system noise is reduced. Since the TC7109A incorporates a zero integrator cycle, the size of the auto-zero capacitor does no affect overload recovery. The optimal value of the auto-zero capacitor is between 2 and 4 times C_{INT}. A typical value fo C_{AZ} is 0.33μF.

The inner foil of C_{AZ} should be connected to pin 31 and the outer foil to the RC summing junction. The inner foil of C_{INT} should be connected to the RC summing junction and the outer foil to pin 32 for best rejection of stray pickups. For low leakage at temperatures above +85°C, use Teflon capacitors.

Reference Capacitor

A 1µF capacitor is recommended for most circuits. However, where a large common-mode voltage exists, a larger value is required to prevent roll-over error (e.g., the reference low is not analog common), and a 409.6mV scale is used. The roll-over error will be held to 0.5 count with a 10µF capacitor. For temperatures above +80°C use Teflon or equivalent capacitors for their low leakage characteristics.

Reference Voltage

To generate full-scale output of 4096 counts, the analog input required is $V_{IN} = 2\ V_{REF}$. For 409.6mV full scale, use a reference of 204.8mV. In many applications, where the ADC is connected to a transducer, a scale factor will exist between the input voltage and the digital reading. For instance, in a measuring system, the designer might like to have a full-scale reading when the voltage for the transducer is 700mV. Instead of dividing the input down to 409.6mV, the designer should use the input voltage directly and select V_{REF} = 350mV. Suitable values for integrating resistor and capacitor would be 34kΩ and 0.15µF. This makes the system slightly quieter and also avoids a divider network on the input. Another advantage of this system occurs when temperature and weight measurements with an offset or tare are desired for nonzero input. The offset may be introduced by connecting the voltage output of the transducer between common and analog high, and the offset voltage between common and analog low, observing polarities carefully. In processor-based systems using the TC7109A, it may be more desirable to use software and perform this type of scaling or tare subtraction digitally.

Reference Sources

A major factor in the absolute accuracy of the ADC is the stability of the reference voltage. The 12-bit resolution of the TC7109A is one part in 4096, or 244 ppm. Thus, for the on-board reference temperature coefficient of 70ppm/°C, a temperature difference of 3°C will introduce a one-bit absolute error. Where the ambient temperature is not controlled, or where high-accuracy absolute measurements are being made, it is recommended that an external high quality reference be used.

A reference output (pin 29) is provided which may be used with a resistive divider to generate a suitable reference voltage (20mA may be sunk without significant variation in output voltage). A pull-up bias device is provided which sources about 10µA. The output voltage is nominally 2.8V below V^+. When using the on-board reference, REF OUT (pin 29) should be connected to REF^- (pin 39), and REF^+ should be connected to the wiper of a precision potentiometer between REF OUT and V^+. The test circuit shows the circuit for a 204.8mV reference, generated by a 2kΩ precision potentiometer in series with a 24kΩ fixed resistor.

Interfacing

Direct Mode

Combinations of chip enable and byte-enable control signals which may be used when interfacing the TC7109A to parallel data lines are shown in Figure 12. The $\overline{\text{CE}}\text{/LOAD}$ input may be tied low, allowing either byte to be controlled by its own enable (Figure 12A). Figure 12B shows the $\overline{\text{HBEN}}$ and $\overline{\text{LBEN}}$ as flag inputs, and $\overline{\text{CE}}\text{/LOAD}$ as a master enable, which could be the READ strobe available from most microprocessors. Figure 12C shows a configuration where the two byte enables are connected together. The $\overline{\text{CE}}\text{/LOAD}$ is a chip enable, and the $\overline{\text{HBEN}}$ and $\overline{\text{LBEN}}$ may be used as a second chip enable, or connected to ground. The 14 data outputs will be enabled at the same time. In the direct MODE, SEND should be tied to V^+.

Figure 13 shows interfacing several TC7109A's to a bus, ganging the $\overline{\text{HBEN}}$ and $\overline{\text{LBEN}}$ signals to several converters together, and using the $\overline{\text{CE}}\text{/LOAD}$ input to select the desired converter.

Figures 14 – 19 give practical circuits utilizing the parallel three-state output capabilities of the TC7109A. Figure 14 shows parallel interface to the Intel MCS-48, -80 and -85 systems via an 8255 PPI, where the TC7109A data outputs are active at all times. The 8155 I/O ports may be used in an identical manner. This interface can be used in a read-after-update sequence, as shown in Figure 15. The data is accessed by the high-to-low transition of the STATUS driving an interrupt to the microprocessor.

The RUN/$\overline{\text{HOLD}}$ input is also used to initiate conversions under software control. Figure 16 gives an interface to Motorola MC6800 or MOS Technology MCS650X system.

An interrupt is generated through the Control Register B, CB1 line from the high to low transition of the STATUS output. The RUN/$\overline{\text{HOLD}}$ pin is controlled by CB2 through Control Register B, allowing software control of conversions.

Direct interfacing to most microprocessor busses is easily accomplished through the three state output of the TC7109A.

Figures 1, 17 and 18 are typical connection diagrams. To ensure requirements for setup and hold times, minimum pulse widths, and the drive limitations on long busses are met, it is necessary to carefully consider the system timing in this type of interface. This type of interface is used when the memory peripheral address density is low, providing simple address decoding. Interrupt handling can be simplified by using an interface to reduce the component count.

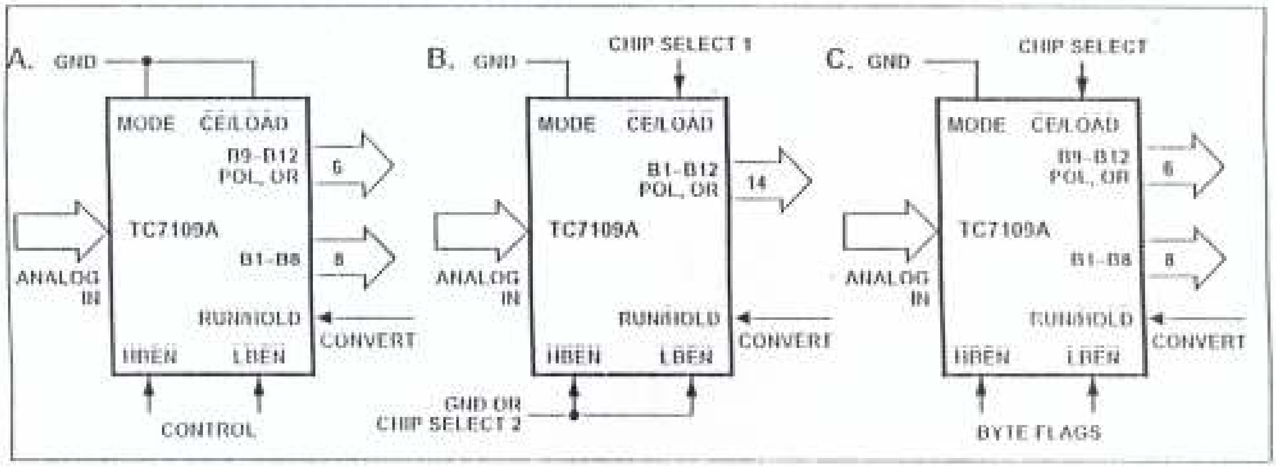


Figure 12. Direct Mode Chip and Byte Enable Combinations

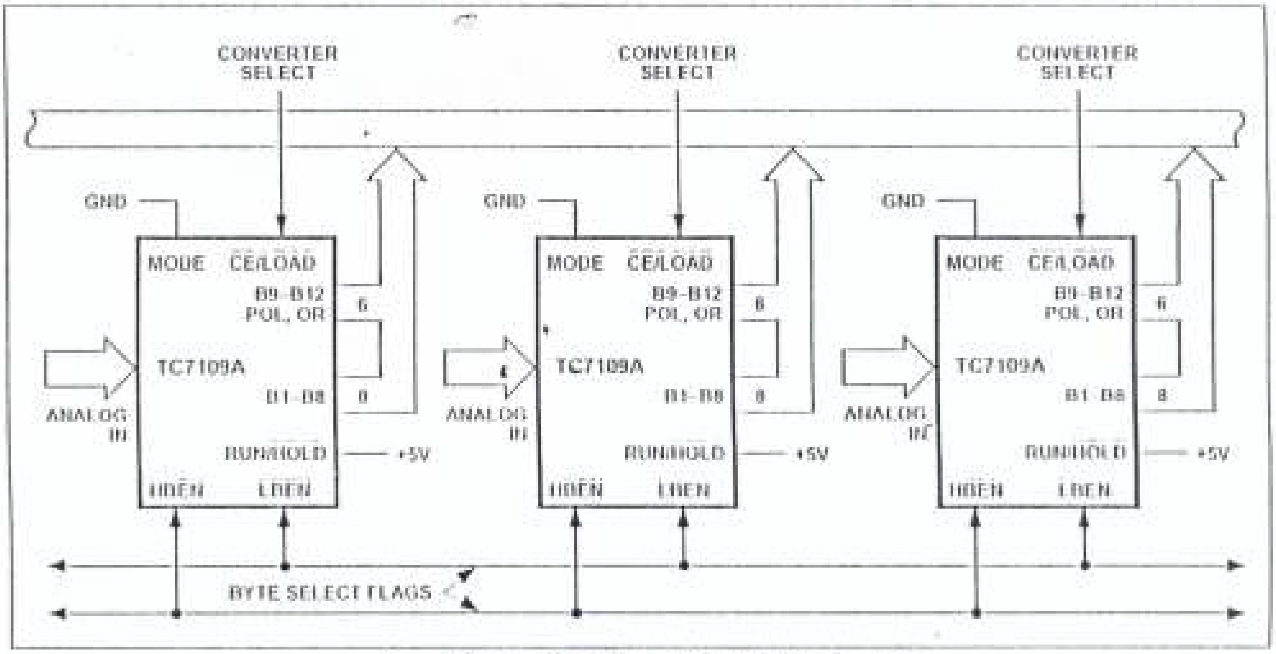


Figure 13. Three-Stating Several TC7109A's to a Small Bus

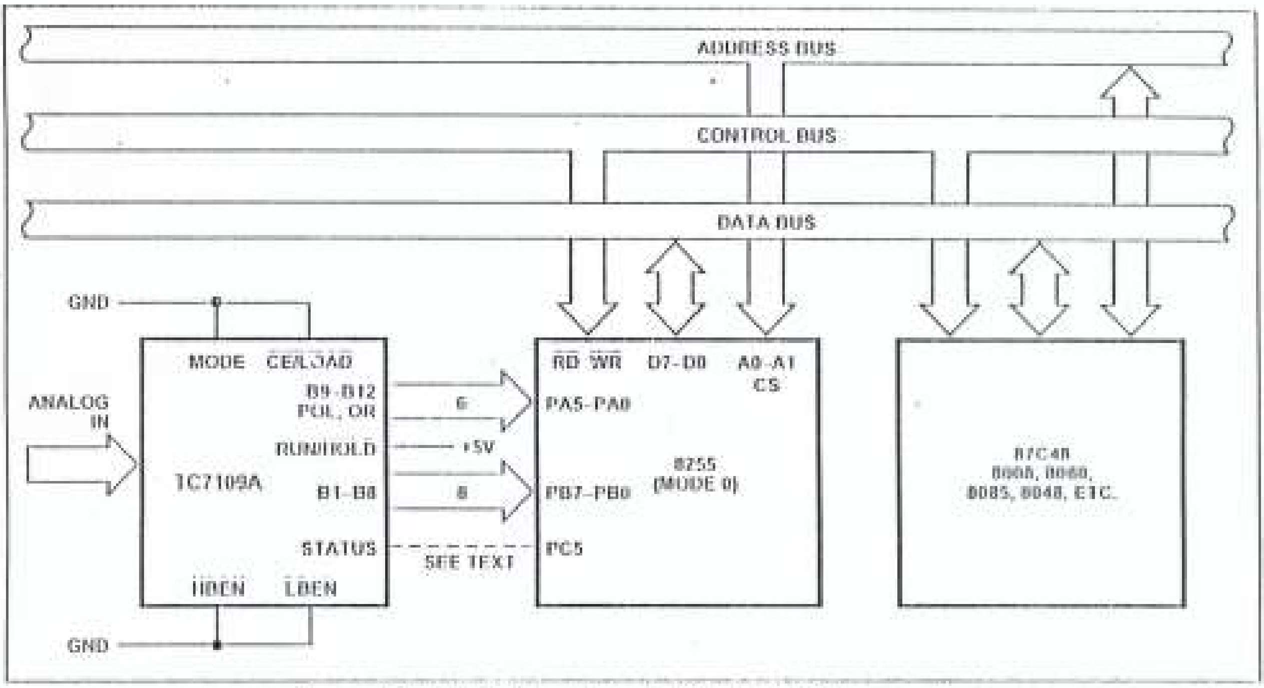


Figure 14. Full-Time Parallel Interface to MCS-48, -80, -85 Microcomputers

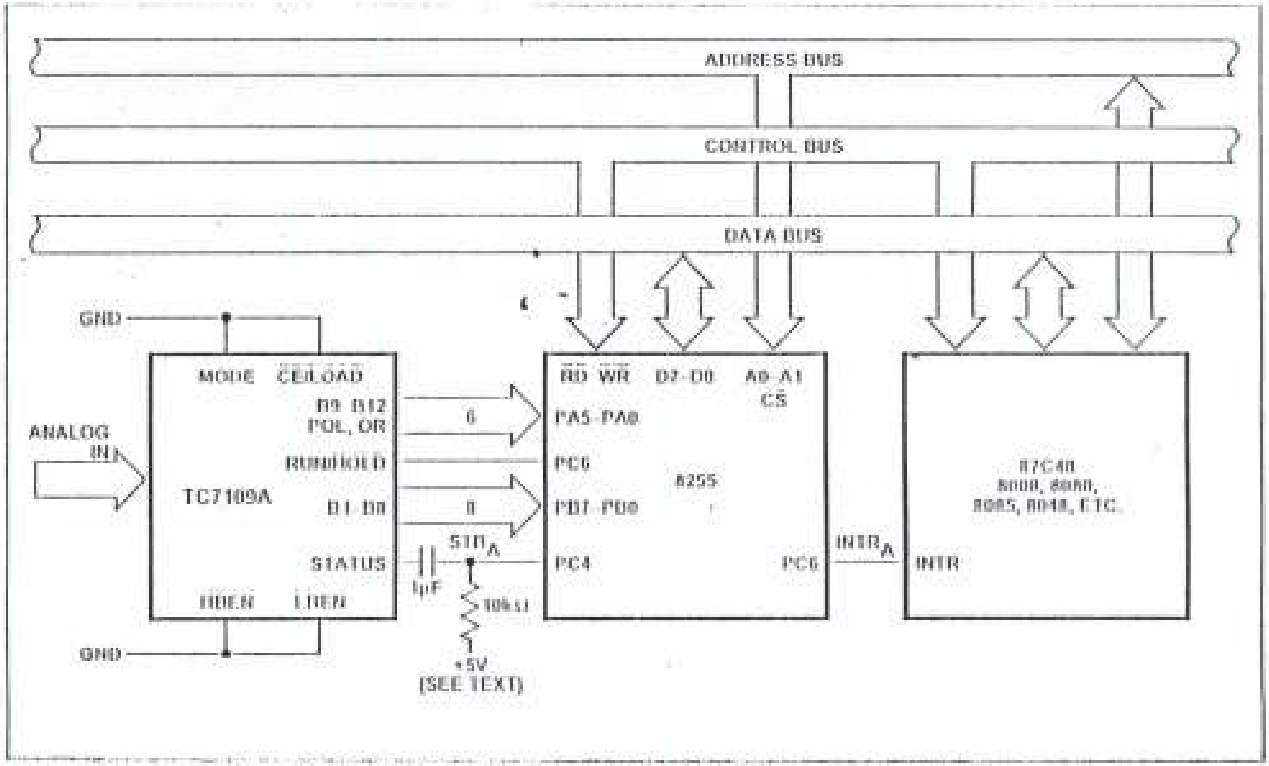


Figure 15. Full-Time Parallel Interface to MCS-48, -80, -85 Microcomputers With Interrupt

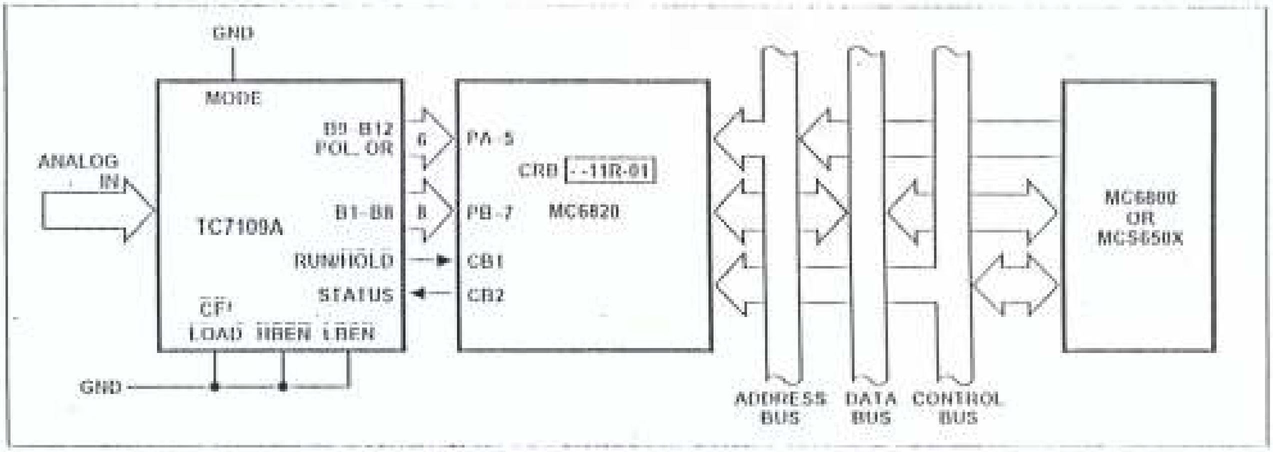
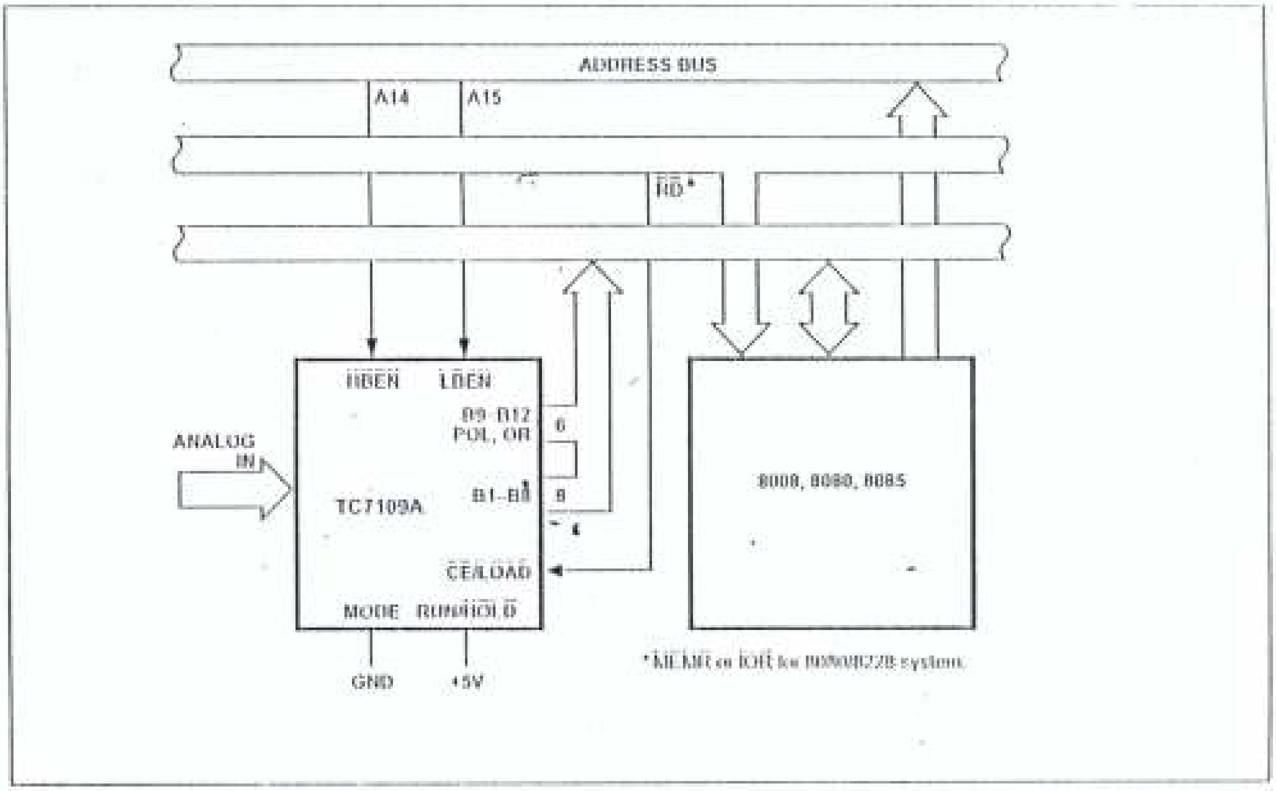


Figure 16. Full-Time Parallel Interface to MC6800 or MCS650X Microprocessor



*$\overline{\text{MEMR}}$ or $\overline{\text{IOR}}$ for 8080/8228 system.

Figure 17. TC7109A Direct Interface to 8080/8085

Appendix E

Source Code for the Software Written in Visual Basic

Option Explicit

```
Dim Rete. d, Temp, Pres, Inter, Logger, Humid As
Integer
```

Option button control

```
Private Sub Check1_Click()
If Check1.Value = 0 Then
txttemp.Text = "SELECT TO ACTIVATE"
txttemp.Enabled = False
Else
txttemp.Text = "Temperature"
txttemp.Enabled = True
End If
End Sub

Private Sub Check2_Click()
If Check2.Value = 0 Then
txtpres.Text = "SELECT TO ACTIVATE"
txtpres.Enabled = False
Else
txtpres.Enabled = True
txtpres.Text = "Pressure"
End If
End Sub

Private Sub Check3_Click()
If Check3.Value = 0 Then
txthumid.Text = "SELECT TO ACTIVATE"
txthumid.Enabled = False
Else
txthumid.Text = "Humidity"
txthumid.Enabled = True
End If
End Sub

Private Sub Check4_Click()
If Check4.Value = 1 Then
cmdstart.Enabled = True
Else
cmdstart.Enabled = False
End If
End Sub
```

Action performed when start button is clicked

```
Private Sub cmdstart_Click()
Open "C:\popo\" & Date$ & ".txt" For Append As #2
Print #2,
Print #2, "WEATHER DATA LOGGED FOR THE DAY " & Date$
Print #2, "Time", "TEMP", "PRES", "HUMID"
Close
cmdstop.Enabled = True
Timer2.Enabled = True
Check1.Value = 1
Check2.Value = 1
Check3.Value = 1
Inter = Val(txtinterval.Text) * 60
'MsgBox inter
'Timer2.Interval = inter
'cmdstart.Enabled = False
'UpDown1.Enabled = False
End Sub
```

Action perfomed when stop button is clicked

```
Private Sub cmdstop_Click()
cmdstart.Enabled = True
cmdstop.Enabled = False
UpDown1.Enabled = True
Timer2.Enabled = False
End Sub

Private Sub Command1_Click()
Unload Me
End Sub
```

Action perfomed at form load

```
Private Sub Form_Load()
Left = (Screen.Width - Width) / 2
Top = (Screen.Height - Height) / 2
Check1.Value = 0
Check2.Value = 0
Check3.Value = 0
txttemp.Text = "SELECT TO ACTIVATE"
txttemp.Enabled = False
txtpres.Text = "SELECT TO ACTIVATE"
txtpres.Enabled = False
txthumid.Text = "SELECT TO ACTIVATE"
txthumid.Enabled = False
cmdstop.Enabled = False
Timer2.Enabled = False
cmdstart.Enabled = False
End Sub
```

Timer initiated actions

```
Private Sub Timer1_Timer()
txttime.Text = Time$
txtdate.Text = Date$
End Sub
Private Sub Timer2_Timer()
Logger = Logger + 1
Rete = Rete + 1
```

```
If Check1.Value = 1 And Rete = 1 Then
For d = 1 To 500
Next d
Temp = ADC1(1)
txttemp.Text = Temp
End If
If Check2.Value = 1 And Rete = 5 Then
For d = 1 To 500
Next d
Pres = ADC2(2)
txtpres.Text = Pres
End If
If Check3.Value = 1 And Rete = 9 Then
For d = 1 To 500
Next d
Humid = ADC3(3)
txthumid.Text = Humid
End If
If Rete = 14 Then
Rete = 0
End If
If Logger = Inter Then
Open "C:\popo\" & Date$ & ".txt" For Append As #3
Print #3, Time$, txttemp.Text, txtpres.Text, txthumid.Text
Close
Logger = 0
Beep
End If
End Sub
```

Updown(down click) control.

```
Private Sub UpDown1_DownClick()
If Val(txtinterval.Text) > 1 Then
txtinterval.Text = Val(txtinterval.Text) - 1
Else
MsgBox "The least logging interval is 1 minute"
End If
End Sub
```

Updown(up click) control

```
Private Sub UpDown1_UpClick()
txtinterval.Text = Val(txtinterval.Text) + 1
End Sub
```

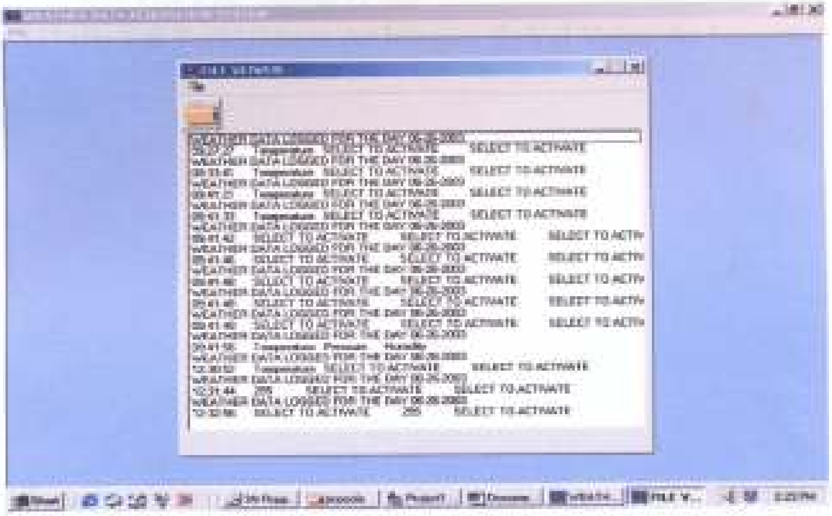
Action to stop operation

```
Private Sub cmdExit_Click()
Unload Me
End Sub
```

Software calibration of the ADC

```
Private Sub cmdGain_Click()
Dim read As Integer
Dim actual As Single
read = ADC1(1)
actual = read * Val(Text3.Text)
Text4.Text = Int(actual)
End Sub

Private Sub cmdoffset_Click()
Dim read As Integer, yourval As Integer, cofactor As Sing
yourval = Val(Text2.Text)
read = ADC1(1)
'MsgBox read
If yourval <> 0 Then
cofactor = read / yourval
Text3.Text = cofactor
Else
MsgBox "Please Enter a Voltage Value for Channel 1
End If
End Sub


Private Sub Form_Load()
Left = (Screen.Width - Width) / 2
Top = (Screen.Height - Height) / 2
End Sub

Private Sub Command1_Click()
ComDiag.ShowColor
List1.BackColor = ComDiag.Color
End Sub

Private Sub Form_Load()
Left = (Screen.Width - Width) / 2
Top = (Screen.Height - Height) / 2
End Sub

Private Sub mnufileexit_Click()
Unload Me
End Sub

Private Sub mnufileopen_Click()
Dim strodo As String
ComDiag.ShowOpen
On Error GoTo handler
Open ComDiag.FileName For Input As #1
List1.Clear
```

```
Line Input #1, strodo
List1.AddItem strodo

Do Until (EOF(1))
Line Input #1, strodo
List1.AddItem strodo
Loop
Close
handler:
End Sub

Private Sub Toolbar1_ButtonClick(ByVal Button As
ComctlLib.Button)
Dim strodo As String
ComDiag.ShowOpen
On Error GoTo handler
Open ComDiag.FileName For Input As #1
List1.Clear
Line Input #1, strodo
List1.AddItem strodo

Do Until (EOF(1))
Line Input #1, strodo
List1.AddItem strodo
Loop
Close
handler:
End Sub
```



The user interface for viewing stored result

User interface for ADC calibration

1. **Dataprop.bas**

```
Type polad
Stime As String
Temp As String
Press As String
Humid As String
End Type
Public DaDa As polad
```

2. Inport 32.bas

```
'Inp and Out declarations for direct port I/O
'in 32-bit Visual Basic 4 programs.

Public Declare Function Inp Lib "inpout32.dll" _
Alias "Inp32" (ByVal PortAddress As Integer) As Integer
Public Declare Sub Out Lib "inpout32.dll" _
Alias "Out32" (ByVal PortAddress As Integer, ByVal Value As Integer)
```