# ANALYSIS AND PERFORMANCE IMPROVEMENT OF INDUCTIVE TRANSDUCERS AND THERMISTORS

**A THESIS**

*submitted in fulfilment of the*
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*By*

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DEDICATED

TO

MY LATE MOTHER-IN-LAW

WHO INSPIRED THE BEGINNING

BUT COULD NOT LIVE TO SEE THE END

## CANDIDATE'S DECLARATION

I hereby certify that the work which is being presented in the thesis entitled ANALYSIS AND PERFORMANCE IMPROVEMENT OF INDUCTIVE TRANSDUCERS AND THERMISTORS in fulfilment of the requirement for the award of the Degree of Doctor of Philosophy submitted in the Department of ELECTRICAL ENGINEERING of the University is an authentic record of my own work carried out during a period from Dec. 1986 to Apr. 1990 under the supervision of Dr. S.C. Saxena, Professor.

The matter embodied in this thesis has not been submitted by me for the award of any other Degree.

30-4-90.

Candidate's Signature

This is to certify that the above statement made by the candidate is correct to the best of my knowledge.

Date: 30.4.1990

Signature(s) of Supervisor(s)

The candidate has passed the Viva-Voce examination held on 8.6.1991 at Roorkee The thesis is recommended for award of the Ph.D. Degree.

Signature of Guide(s)

8/6/91

Signature of External Examiner(s)

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(S.B.LAL SEKSENA)

## ABSTRACT

Transducers are widely used for measurement of physical quantities in Biomedical, Aerospace, Meteorological, Industrial and Process Instrumentation systems. These are essential link between the physical systems and electronic signal processing, conditioning, and handling units. Commonly encountered problem with transducers is variation in their performance characteristics due to changes in excitation parameters and in internal and surrounding environmental conditions. The responses of most of the transducers are non-linear throughout their operating range. There is also another problem of stability in their long term use. Thus, there is continual effort by the researchers right from the beginning to improve the performance characteristics of transducers either by developing a new transducer or by improving the design of existing transducer. In view of the importance of this area, for this thesis work, two types of transducers have been selected for the improvement of their performance characteristics. Differential inductive transducers have been selected due to their extensive use in the measurement of large number of physical parameters, namely displacement, thickness, force, pressure, flow, level, velocity, acceleration, vibration, torque etc. NTC thermistors have been selected as the other transducer as these are most commonly used for the measurement of temperature due to their convenient shape, size and high order of sensitivity.

The emphasis has been placed on the improvement of performance of differential inductive transducers by providing self-compensation as these are used in almost all types of environments including hostile conditions. The response linearization of thermistors have been taken as the parameter of interest as this is the main problem encountered in their use.

In the first part of the work, three type of inductive transducers have been developed with improved performance characteristics, one based on linear variable differential transformer (LVDT) principle and other two based on ratio and differential combinations of two inductive coils. The LVDT has been developed using dual set of secondary windings for self-compensation. This transducer is immune to variations in excitation conditions and changes in environmental temperatures. The output voltage signal from one set of secondary windings is taken as difference of induced voltages and is termed as the differential output, whereas the output signal from other set of the secondary windings is taken as the sum of the induced voltages and is termed as a reference output. The final output is the ratio of these two output signals and does not contain terms influenced by variations in excitation parameters and ambient temperature. The transducer is extensively tested for changes in excitation conditions and environmental temperature . The performance has been evaluated and compared with that of the conventional uncompensated LVDT transducer. It is found from the results that

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In the second part of the work, the problem of nonlinearity of the responses of transducers has been tackled. Two analog and one digital techniques have been developed for response linearization of thermistors. In one analog technique, thermistor is connected in feed-back path of an active-bridge circuit. To linearize the input-output characteristic of the bridge circuit, Taylor's theorem has been used. In second analog method, a log-amplifier in conjunction with analog divider unit has been used for response linearization. The divider unit is used as an inverter to invert the converted variable. The linearized output responses of these circuits are proportional to input temperature at the transducer end.

The linear output of single-thermistor-active-bridge amplifier has been used to develop a microprocessor-based multichannel temperature scanning system for the measurement of temperature, using PWM technique. The scheme has been tested for the measurement of temperature from eight different locations.

A digital technique has also been developed using scale factor polynomial approach for the linearization of response characteristics of thermistors. It has been implemented around a digital computer. The scope of the technique is not only confined to linearization of the thermistor response, but can be extended to response linearization of all type of non-linear response characteristics of other transducers. The validity of the scheme has been tested on the response curves of thermistor and LVDT transducer (the former falling into the category of exponentially decaying characteristic, whereas the latter one into the category of exponentially rising characteristic). It has been found that the linearity of response improves with piece-wise segment linearization. The results are satisfactory and are in conformity with the results obtained by other existing digital techniques.

To sum up at the end, it can be stated that these inductive transducers are suitable for all types of indoor and outdoor applications including hostile environments owing to their high performance stability. These are the improved and modified versions in the existing family of inductive transducers. The analog linearization techniques are applicable only for NTC thermistor whereas the digital technique is useful for all types of transducer response characteristics.

# CHAPTER - I

## INTRODUCTION

### 1.1 GENERAL

Transducer is a device that converts variation of one physical quantity into an usable optical, mechanical, or more commonly an electrical signal. In particular for control and instrumentation field, the transducer is defined as the device which converts the measurand (measured quantity, property, or condition) into a usable electrical output [74]. This is called electrical transducer. The energy conversion process is referred to as transduction or transformation. Electrical transducers are broadly classified into various categories on the basis of their principle of transduction, form of measurand and the requirement of external excitation. Transducers are extensively used for the measurement of physical parameters in almost all fields of engineering and science. Bio-medical, Aerospace, Meteorological, Industrial, and Process instrumentation systems make extensive use of transducers. These are essential links between physical systems and front end electronic signal processing and conditioning systems.

Number of problems are encountered with transducers when they are put to use in different fields. Common problems are variation in their performance characteristics like non-linearity in response and instability throughout their operating range due to changes in excitation parameters and internal and surrounding environmental conditions. The accuracy of measurement is affected due to these influencing parameters. These problems have drawn considerable attention of investigators right from the beginning and there is continual efforts to improve the performance characteristics of transducers either by developing new transducers or by making certain improvements in the design of existing transducers. By realizing the importance of this area, following two types of transducers have been selected for the improvement of their performance characteristics in this work;

i) Inductive Transducers

ii) Thermistors.

Inductive transducers are extensively used for the measurement of large number of physical (mechanical) quantities, namely displacement, thickness, force, pressure, flow, level, velocity, acceleration, vibration, torque, etc. The work has been carried out for performance improvement of inductive transducers by providing self-compensation as these are used in almost all types of environments including hostile conditions. Negative temperature coefficient (NTC) thermistors are most frequently used for the measurement of temperature in bio-medical, marine-science, atmospheric-science, and other measuring systems due to their convenient shape and size, high temperature coefficients of resistance, high resistivity, and good reliability. Thermistors are basically temperature sensitive resistors which exhibit relatively large nonlinear resistance change over narrow temperature range. From the initial stage of their development, nonlinearity of response is the major problem encountered in their use. The output response linearization of thermistor has been taken up as the other parameter of interest in this work. Following sections briefly review the developments which have taken place for the self-compensation of inductive transducers and response linearization of thermistors.

## 1.2 INDUCTIVE TRANSDUCERS

Physical input parameters to be measured, hereafter termed as measurand, can vary the inductance of a coil. This variation in inductance can be used to measure the value of physical parameter. Electrical and electronic circuits are used to convert and measure the change of inductance. In such cases, the inductance coil forms an integral part of the circuit. Transducers operating on this principle are called variable inductance transducers.

The inductance of an uniformly wound single layer coil is expressed as

$$L = \frac{4\pi^2 n^2 r^2}{\ell . 10^7} \text{ (H)} \quad \text{(for air core)} \qquad \ldots (1.1)$$

where,

n is number of turns of coil,

$\ell$ is length of coil (m),

r is radius of coil (m),

and a is cross-sectional area of coil ($m^2$) is equal to $\pi r^2$.

This expression of equation (1.1) can be rewritten as

$$L = n^2 G \mu \quad \ldots (1.2)$$

where,

$\mu$ is the effective permeability of medium in and around the coil and is equal to $4\pi \times 10^{-7}$ H $m^{-1}$

G is the geometric form factor.

Since a change in the inductance of the coil can be obtained by varying n, $\ell$, a, or $\mu$, the transducers are classified in the following categories:-

i) Variable-reluctance type

ii) Variable-permeability type

iii) Variable coupling (between two or more coils) type

iv) Eddy current type

v) Magnetoresistive type

In general, variation of G is confined to the transducers employing air-cored coils only. In variable reluctance type transducers, inductance of a coil is changed by variation of length of magnetic circuit by the measurand.An iron armature and a coil wound on a permanent magnet make this type of simple transducer. Reluctance type transducer is also made by using two or more ac excited coils (changing the reluctance by movement of the armature). Inductive transducers, which operate by the displacement of a ferromagnetic core (armature) by changing the magnetic flux linkages of a coil or coils, are broadly divided into two categories. One category of such transducers consists of three coaxial windings. The primary winding is energized by sinusoidal excitation. Two identical secondary windings are connected in series - opposition and produce output voltages in response to flux linkages from primary winding. The transducer produces an ac output (differential) in response to input displacement of core. Transducers belonging to this category are called linear variable differential transformers (LVDTs). The other category of transducers is configured by two matched (identical) and coaxial coils in push-pull arrangement. A core moves inside the coil assembly and changes the flux linkages between them. Transducers belonging to this class are called variable inductance displacement transducers (VIDTs). These transducers are invariably used for the measurement of dynamic physical parameters, particularly displacement in almost all fields of science and engineering. A variety of commonly used variable inductance transducers are shown in Figs.1.1 and 1.2.

Measurement of nonelectrical quantities (particularly displacement) using inductive transducers by exploiting the property of variation of inductance in response to corresponding change in the position of core inside the coil is extensively used in many fields. Lion [62], Neubert [73], Norton [74], and Doebelin [30] explained the working principle of these transducers in the respective works. The details are also given regarding the operating ranges, deviation from linearity, and their different uses. Thompson [110] and Lion [63] have

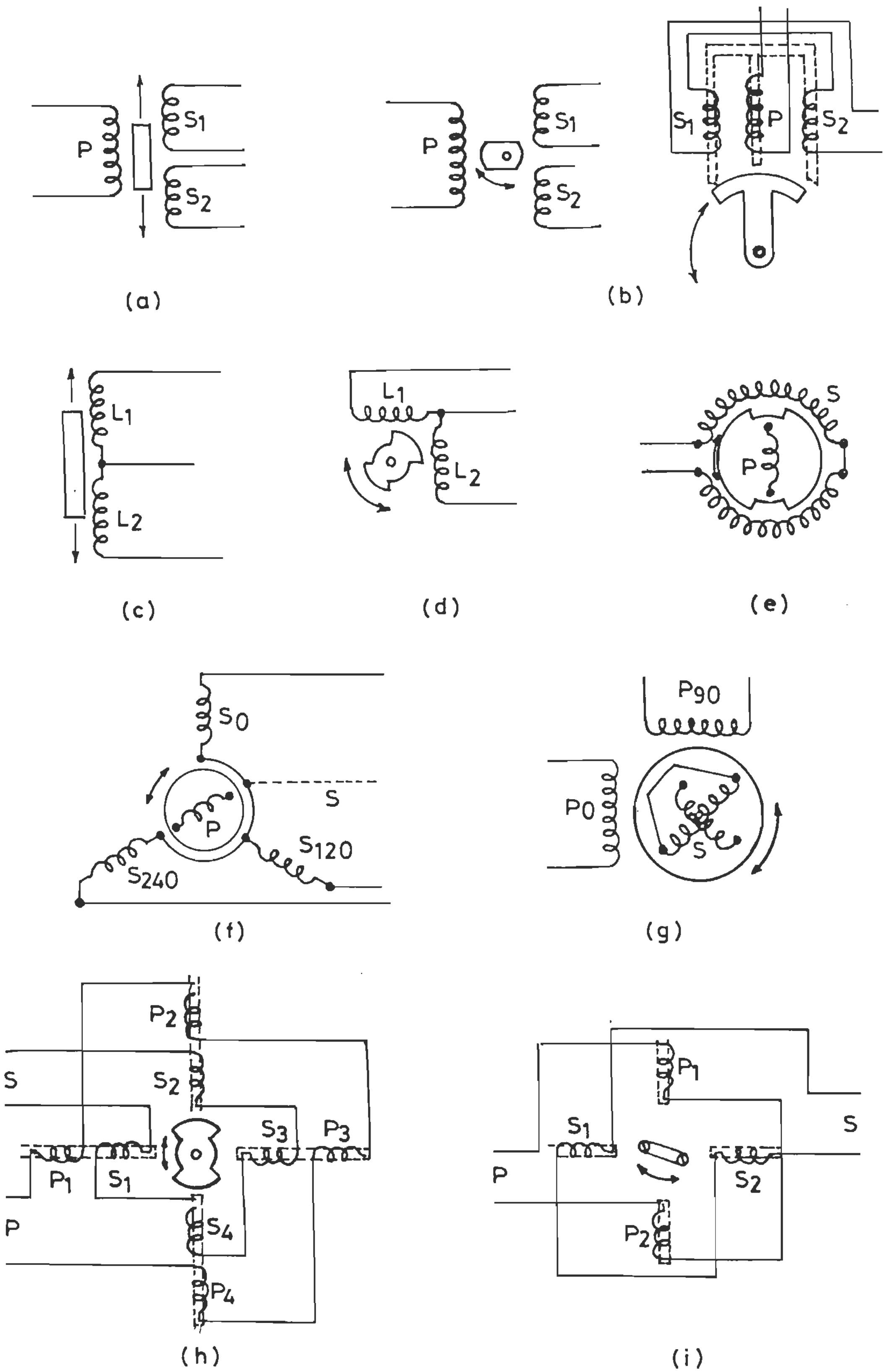



FIG. 1.1 RELUCTIVE DISPLACEMENT TRANSDUCERS
(a) DIFFERENTIAL TRANSFORMER (LINEAR); (b) DIFFERENTIAL TRANSFORMERS (ANGULAR); (c) INDUCTANCE BRIDGE (LINEAR); (d) INDUCTANCE BRIDGE (ANGULAR); (e) INDUCTANCE POTENTIOMETER; (f) SYNCHRO; (g) RESOLVER; (h) MICROSYN; (i) SHORTED-TURN SIGNAL GENERATION.

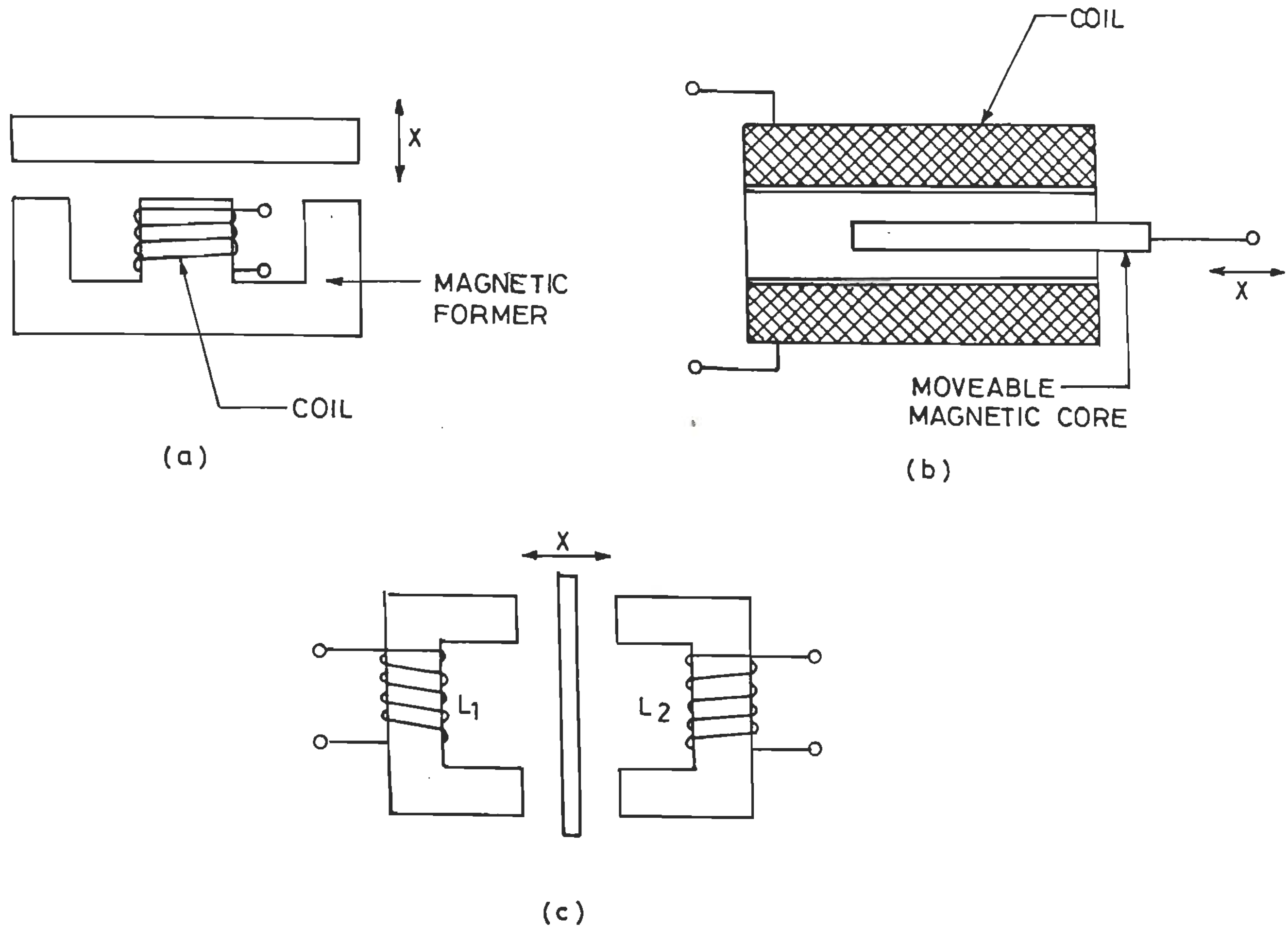



FIG. 1.2 EXAMPLES OF VARIABLE-RELUCTANCE INDUCTIVE TRANSDUCERS

provided a general review on transducers including the basic background types, and associated problems and prospects. In the year 1983, Payne [80] has presented a good review article on sensors covering their impact on modern industries, applications in various fields, and new emerging technologies. The resistivity of metals (e.g., bismuth) at low temperature changes, if the metal is exposed to a magnetic field. By making use of this phenomenon, Williamson [119] developed a magneto-resistance displacement transducer to detect (theoretical) minimum displacement of about $1.3\times10^{-11}$ cm and obtained sensitivities of upto $1.1\times10^{2}$ V/m for the transducer. Another magnetoresistor displacement transducer was developed by making use of shaded pole pieces in place of rectangular pole pieces. The linearity of 5 per cent was achieved over a (displacement) range of 14 mm by Ferris and Ivision [31]. Ara [3] developed a linear variable differential transformer transducer for compensation against the variations in excitation parameters and changes in the environmental temperature. In general, the immunity to such variations has been obtained by using another reference LVDT, NTC thermistor or additional associated front end electronic circuitry. Sydenham [109] thoroughly reviewed resistive, inductive, capacitive, and other alternate methods for the measurement of distances upto micrometer range. Weckenmann [117] described an inductive transducer with short-circuiting ring for the measurement of displacement upto 10 cm. It gave an error in linearity of 0.3 per cent for its full scale range. Hugill [42] carried out synthesis of inductive displacement transducers on a computer using a prior knowledge of magnetic flux distribution within the transducer. This method is known as field modelling technique. The synthesised system was compared with experimental results obtained from the same system under similar design specifications. It was observed that sensitivities of both the systems were within 2 per cent. Abdullah [1] has suggested computer - aided design method for developing LVDT transducer to bring down the error limit even upto fractions of a per cent. According to Hillyard [40], Portasilo Research incorporated LVDT with a weighing system which is suitable for use in hazardous areas. This LVDT has a core travel of +0.025 inch with an output sensitivity of 200 mV per 0.001 inch. Lane [60] discussed possibility of using LVDT displacement sensor for pressure measurement with good stability and linearity for small displacement of 0.02 inch. Penny [83] developed a circuit to provide compensation for nonlinearity of the LVDT used for pressure measurement by Lane [60]. Reviewing different methods for pressure measurement, Shephard [97] suggested the use of LVDT to convert pressure proportional to displacement (which in turn is equivalent to its electrical output). Garrantt [33] carried out a good review work on displacement transducers for 50 mm core travel and provided number of the related references. A variable-inductance type transducer was developed by Jaura [46] for the measurement of displacement in the range of ±4mm with a sensitivity of 270 mV/mm. This transducer assembly consists of two coils arranged in push-pull fashion and forms two arms of a four arm-ac-bridge circuit. The other two arms are resistive in nature. The bridge is excited by an ac supply of 4V, 400 Hz. Uno [115] described a practical method with ferrite core for the measurement of inductance of a coil from 250 to 350 μH. Clabburn [23] developed a resistive displacement transducer for the measurement of acceleration upto to $10^5$ $ms^{-2}$. He achieved an accuracy of 0.5 per cent of full scale under static calibration and within 2 per cent of full scale under dynamic conditions. Later on in 1984, Groenland [37] developed a magneto-resistive transducer for the measurement of absolute linear or angular position detection. He also performed the experiments using the sensors with a resolution of 250 μm and 1 mm. By applying the principle of linear variable differential transformer, Kano, et al. [48] developed two transducers for detecting the position of linear dc motors in one- or two-dimension in the operating ranges of ±20 mm and ±8 mm. The linear range for the displacement is ±15 mm and ±4 mm in each case. Magnetic position sensor based on effective permeability of a slander ferromagnetic core element with new construction ideas was realized by Garshelis and Fiegel.[34]. A novel variable reluctance sensor with two possible and easy-to-realize design configurations has been reported by Lemarquand [61]. Literature further reveals that inductive transducers have been developed for the measurement of displacement in the range from 0.5 mm to 2500 mm with linearity within ±0.5 per cent of full scale value by Haslam, et al. [39]. Typical response characteristic of the transducer is shown in Fig. 1.3.

The performance characteristics of inductive transducers reviewed in the preceding paragraph are severely affected by variations in excitation parameters and changes in ambient temperature in and around the transducer assembly. As such, these can not be put in long term use without adequate compensation for changes in undesired inputs and ambient temperatures. There is necessity to develop self-compensated and smart trnsducers for use in different type of environments including hostile conditions. Three types of inductive transducers one based on linear variable differential transformer (LVDT) principle and other two based on inductive ratio (IR) and inductive different differential (ID) principles have been developed with overall improved performance characteristics. Here after, they are referred to as LVDT, IRT, and IDT transducers in this thesis.

## 1.3 THERMISTORS

Temperature is one of the most commonly measured fundamental parameters in all large electrical apparatus and equipment, steel, fertilizer, and ceramic industries, and in various other process, control, and instrumentation schemes. The phenomena which are exploited to get electrical output as a function of temperature include thermal expansion (bimetallic elements), Seeback Voltage (thermocouples), resistance effects

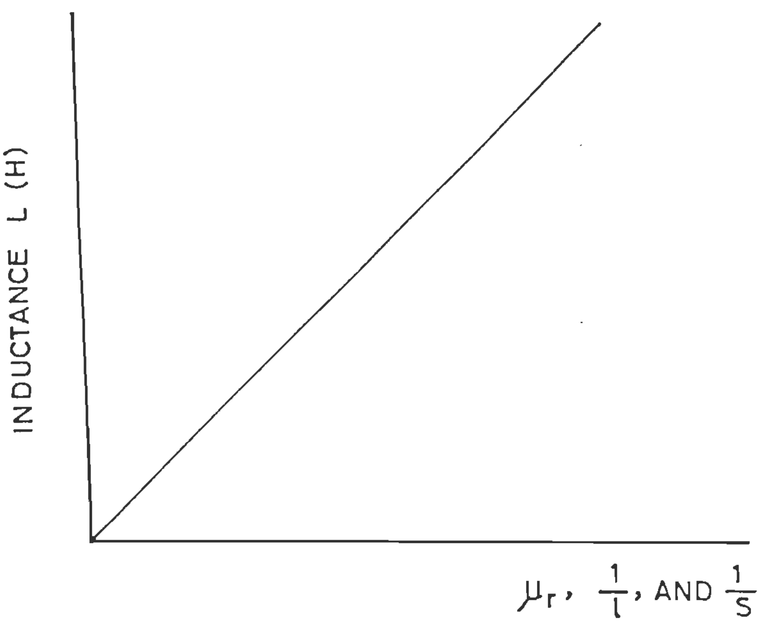



FIG. 1. 3 CHARACTERISTIC OF INDUCTIVE TRANSDUCER

(RTDs and thermistors), and semiconductor junction effects (diodes) [Sheingold [95], Damljanvoic [26], Bently [7]]. The bimetallic thermal switch is the most elementry temperature sensor and it utilizes metals with different thermal coefficients of expansion to physically make or break electrical contact at a preset temperature in thermostates which are used in heating and airconditioning systems. These are used in the temperature ranges from sub-zero to 300°C. Thermocouples, platinum resistance thermometers (PRTs), and thermistors are three most widely used temperature sensors. Thermocouples are popular industrial temperature sensors as by proper selection of thermocouples (wires) materials, it is possible to measure temperature over a wide range in many processes. A copper constantan (type T) thermocouple can be used to measure temperature down to -100°C, and platinum/rhodium-platinum (type R) thermocoupels are used in the temperature range from 0 to 1400°C. Unless each thermocouple is individually calibrated, the over all measurement error can not be reduced below ±1 per cent of full span. Resistance temperature detectors (RTD's), usually in the form of solid conductor made of platinum, copper, tongston, and nickel exhibit small increase in resistance as temperature increases and are characterized by positive temperature coefficients of resistance. Platinum resistance wire is regarded as the standard for accuracy and repeatability among temperature sensors. Platinum resistance thermometers are used to measure temperature in the range of -200 to +500°C with an accuracy of ±0.2 per cent of full span. Thermistors are categorised as negative temperature coefficient (NTC) and positive temperature coefficient (PTC) types. Both are manufactured by different processes from materials processing negative and positive temperature coefficients and are used to measure temperatures in wide ranges. NTC thermistor has been selected for present work. The resistance of NTC thermistor decreases exponentially with increase of its body temperature. They are made from oxides of manganese, cobalt, chromium, or nickel. These thermistors have high sensitivity particularly between -50 and +100°C. This makes them especially suitable for detecting small changes in or near ambient temperature encountered in environmental, meteorological, medical and other such applications. Thermistors are limited to temperature ranges between -80 and +300°C. PTC thermistors are made from silicon (having positive temperature coefficient) and the resistance of a typical element increases from 500 ohms at 55°C to 1900 ohms at 125°C. Sufficient literature Becker, et al. [6], Smith [98,99], Candy [18], Beck [5], Scarr and Setterington [92], Sapoff and Openheim [89], Brauer and Fenner [13], Cirkler and Lobl [22], Prudenziati, et al. [84], and Hyde [43] is available giving their theory, manufacturing process, properties, and applications. Thermistors have advantages of small physical size, low cost, ruggedness, and a large temperature coefficient. One of the main problem encountered with thermistors is their highly nonlinear response characteristic throughout their operating range. A sharp exponential decay in resistance for this circuit element is observed with increase in temperature of medium. From its infancy, copious work has been done by many researchers to improve the gain-nonlinearity of NTC thermistors. In the recent past, useful developments for gain-linearization in desired operating range, have taken place by making use of various analog and digital techniques. The detail survey of analog schemes, reported so far in the literature, is given in chronological order in this section. These include, typical combinations of single thermistor and a fixed resistor, multiple use of thermistors and resistors, thermistors in logarithmic-divider unit with an active compensating network, thermistor in a delay network, and thermistor-resistor combination in conjunction with logarithmic converter and FET.

In early analog schemes, the linearizing network invariably incorporated a fixed value resistance in series with a thermistor. The resultant voltage output from the network was nearly linear over a narrow temperature range. Beakley [4] developed a network of a thermistor and a resistor of fixed value and observed deviation of measured values with the expected ones within 0.024°C in limited temperature range from 10 to 30°C. The value of fixed resistance was selected such that second and higher derivatives of Taylor's expansion of resulting output voltage of the network vanished at a desired temperature. Cole [25] developed a linearization technique using a dc Wheatstone bridge configuration with a slide wire (for ratio arms) employing a thermistor in one of the arms of the bridge. He obtained a linear relationship for response between balance point of this bridge and temperature in the range of 0 to 40°C with a constant sensitivity of 0.1°C. Further, Godin [36] described a Wheatstone bridge method using single thermistor and obtained linear response in a wider temperature range from 280 to 330 K. Pitts and Priestley [82] described a simple constant sensitivity Wheatstone bridge method for thermistor thermometers. In this attempt, compensation for change in sensitivity when the bridge is balanced, has been obtained by a variable resistor. Part of the resistor is in series with the thermistor and the remaining (part) is in series with the bridge voltage supply. Sensitivity of 2.45 $\mu A°C^{-1}$ over a narrow working range of temperature of 15°C (i.e., 56 to 82°C) was achieved by them. Boel and Erickson [9] presented a Wheatstone bridge configuration using a thermistor for response linearization in the range from 0 to 100°C. Inflected curve between bridge open-circuit output voltage and the absolute temperature was obtained and linearity in input - output parameters was observed around point of inflection in a narrow range. The differences in temperature upto 0.01°C were obtained without additional calibration of the bridge. Diamond [27] improved upon the existing technique for response linearization of thermistor around point of inflection. The improved technique is generalized in nature and can be used with constraints for other transducers in narrow operating range of temperature. A general approach for response linearization of thermistor thermometer was developed by Bowman [11] under different operating modes and was well supported by theoretical and experimental results for a temperature range from 0 to 100°C.. The maximum departure from linearity was observed within ±0.2°C. Later on Bowman and Sagar [12] proposed a different approach for response linearization of a self-heated thermistor. A numerical method was suggested for finding a suitable criteria for linearity. Stankovič [101] modified the nonlinearity analysis of uncorrected linearized thermistor Wheatstone bridge and the absolute nonlinearity calculated differed by about 1 to $2 \times 10^{-3}$ °C in corrected and uncorrected expressions.

The approaches using combination of a fixed resistance and thermistor, discussed so far, emphasize linearization of the function in a narrow temperature range as slope of the curve (inflected) changes very slowly (i.e., remains nearly constant) in the vicinity of the point of inflection. Moreover, the fixed value linearizing resistance is also selected on certain assumptions.

Trolander, et al. [114] investigated multiple use of thermistors and resistance in a network. The linearity of the order of 0.1 per cent of full span was obtained which was better than the previous attempts. But it is a function of number of thermistors connected in the proposed network and further, thermistors must be identical and must sample the similar temperature. Broughton [15], later on, analyzed one-thermistor temperature measuring circuits by transforming their transfer functions into linear fractions. The transfer function is expanded by using Taylor's infinite series about a reference temperature and second and higher derivatives of the series are equated to zero to obtain necessary condition for gain-linearization and selection of circuit elements. The response is linear upto 80 per cent range of full span. The condition of linearity achieved, suits a particular thermistor because it is dependent upon specifications of the thermistor in the circuit. Such circuits are either much complicated in design or make use of extensive mathematical computations to obtain optimum input - output linearity.

In the past twenty years the problem of response linearization using delay network has drawn considerable attention of investigators. The concept of temperature measurement using (temperature) sensors in astable multivibrator bridge was first introduced in the year 1967 by Maher [67]. The sensitivity of PTC sensistor (86 mV $°C^{-1}$), of Si-transistor (73 mV$°C^{-1}$) and of Ge-transistor (650 mV $°C^{-1}$) was investigated by him in the range of 0 to 40°C.

Lövborg [64] converted temperature-to-frequency, linearly, by using a thermistor as an element of a RC-oscillator circuit. The (temperature-frequency) response of oscillator was recorded in the range from 5 to 45°C with an accuracy of about ±0.1°C. Weaving [116] developed a portable temperature integrator using a low frequency silicon unijunction transistor (UJT) in the narrow range from 10 to 32°C. In this integrator, series combination of thermistor and resistance is used to obtain good response linearity and sensitivity. Value of the resistance is obtained by the method developed by Lövborg [64]. The frequency of oscillation (of the developed circuit) is dependent upon the resistance of thermistor. Patranabis and Sen [79] proposed a temperature-to-frequency convertor circuit which provides a constant amplitude square wave output. Analysis was carried out for linearity and sensitivity and practical results were presented. Stanković and Simic [105] described a monostable multivibrator bridge with a capacitor or a resistance transducer, whose output was obtained to be a linear function of the transducer impedance. Sensitivity claimed by them is four times higher than that obtained by Wheatstone bridge circuit employing similar type of transducer. The proposed monostable bridge is complex than the astable multivibrator. Later on, Stanković [100] described a simple and practical temperature-to-frequency convertor around an astable multivibrator bridge circuit using thermistor as a bridge element in the narrow range of temperature form 20 to 50°C. The linearity of the bridge was within ±0.25°C and its dependence on ambient temperature was less than 0.027 per cent °C $s^{-1}$. Buckley and McLeod [16], explained a digital temperature integrator for the range from 40 to 110°F. It uses a linearizing Wheatstone bridge, differential amplifier voltage-to-frequency convertor, divider circuit, and a readout unit. A positive temperature coefficient thermistor was used as temperature sensor. The accuracy of the readout unit was obtained within ±2 per cent of the readout value. The details of the voltage-to-frequency converter are nicely presented by O'Haver [75]. Further, in the year 1974, Stanković [103] developed thermistor-astable multivibrator and monostable multivibrator bridge circuits for response linearization in the range of 30 to 60°C with mean sensitivity of 135 mV°$C^{-1}$ and 20 to 60°C for sensitivity value of 160 mV °$C^{-1}$ and the drift was 0.21 and 0.38 mV °$C^{-1}$, respectively. Here the lower stability of monostable multivibrator bridge (MMB) compared to astable multivibrator bridge (AMB) is due to the obvious reason of increased complexity of the former bridge configration. Stanković [102] described theoretically a general method for obtaining linear temperature-to-frequency and temperature-to-time conversions by means of thermistor circuit in which possibilities of connecting a passive resistance either in series or parallel with thermistor were investigated. Ikeuchi, et al. [45] described (analytically) a linear temperature-to-frequency converter by making use of an astable multivibrator bridge employing single thermistor. The centre frequency of the converter is independent of the linearizing condition. With well supported analysis, Natrajan [71] suggested two square wave generator circuits around operational amplifier for linear temperature-to-frequency conversion using a thermistor or combination of thermistor and fixed resistor. Deviation characteristics for both the circuits were also given. Djordevich [29] proposed a scheme in which temperature dependent thermistor voltage is converted into frequency employing a current comparator, transistorized timing-network, switches, and stabilized voltage supply for an accurate adjustment of timing network. The obtained input-output characteristic of the scheme is linear over a temperature range from 273 to 400 K. Natrajan and Bhattacharya [72], suggested astable and monostable multivibrator circuits around operational amplifier using thermistor as a circuit element to convert temperature-to-time linearly. The conditions for linearization around a reference temperature were obtained on some of the circuit parameters. In addition, they have obtained theoretical deviatioin curves. Stanković and Elazar [104] theoretically analysed symmetrical multivibrator bridge with five possible configurations using thermistor as temperature sensors. Alongwith temperature-to-frequency conversion, the possibility of measuring temperature linearly, was also explored in the operating temperature intervals from -65 to 47°C and -1 to 125°C. The sensitivity obtained by them was 1 per cent $°C^{-1}$. Khan and Sengupta [49] presented a thermistor astable multivibrator bridge circuit for linear conversion of temperature-to-frequency in the range from 0 to 86°C and obtained a sensitivity of 21 Hz $°C^{-1}$. This approach is also applicable in the lower temperature range from -50 to 0°C. Sundquist [108] described a linear temperature-to-frequency converter using a monostable multivibrator with a standard thermistor. The linearity of better than 0.5K over 75K range was obtained. This converter circuit can be used to linearzie response to any transducer exhibiting exponential behaviour between input and output signals. Khan and Sengupta [50] presented a simpler version of temperature-to-frequency converter in the operating range of 248 to 523 K. The performance of the tested circuit was found to be linear in the range of 273 to 430 K. Nonlinearity, was observed as 1.2 per cent at 248K and 4.8 per cent at 523K. Later on, Khan and Sengupta [52] proposed an improved linear temperature-to-frequency converter using delay network employing a thermistor in the range of 253K to 523K with maximum deviation of 4.8 per cent at 523K and 1.2 per cent at 253K. It incorporates lesser components and does not requrie a stabilized supply and the calibration is done only at room temperature. Recently in 1988, Khan, et al. [56] suggested a linearizing scheme using a delay network in conjunction with analog multiplier. It was another variation of delay network scheme foer gain-linearization of the thermistor. The scheme has been tested in the range of 283 to 493K and departure from linearity was observed to be decreasing form 6 per cent to 0.6 per cent in the temperature range from 283 to 320K. Sengupta [94] described a temperature measuring circuit using thermistor in a pulse generator for obtaining highly linear relationship between absolute temperature and frequency. The sensitivity of 9.6 Hz $°C^{-1}$ was obtained. The experimental results were obtained for a range from 5 to 85°C, but the circuit has been designed for entire range of temperature measurement (i.e., -100 to 225°C) using thermistor.

Thermistor has been used as a circuit element in logarithmic network for response linearization. Khan and Sengupta [51] developed a technique using thermistor-registor parallel combination with logarithmic amplifier. The technique provides a linear analog temperature-voltage relationship over a wide temperature range from -25 to 110°C. The output of this scheme is expanded in terms of a well established criteria of Taylor's infinite series expansion. Later on in 1985, Khan [53] improved this technique by avoiding the use of logarithmic amplifier chip, which is not easily available in many countries. The performance of the improved circuit was tested in the range from 273 to 433 K and the obtained experimental results are in conformity of the analysis. An error of less than 0.9 per cent was observed in the range of 293 to 410 K. In segments of the temperature from 273 to 293 K and from 410 to 493 K the maximum error observed was less than 2.6 per cent. Khan [54] investigated the theoretical aspects of response linearization schemes with thermistor in logarithmic amplifier network as well as in other conventional schemes in the range from 263 to 453 K with a sensitivity figure of 14 mV $°C^{-1}$. Experimental results obtained thus, were in total conformity with the theoretical analysis. Further response of the proposed thermistor-log amplifier circuit is independent of the fluctuations in the ambient temperature. The power dissipation in thermistor is around 0.33 mW and 48 $\mu$W at 273 and 453 K, respectively. In 1987, Khan, et al.[55]

demonstrated an improved version of logarithmic technique for response linearization of thermistor. The input-output relationship is linear in the range from 253 to 493K with an error less than 1 per cent of full span. Patranabis, et al. [78] developed a general response linearizing scheme applicable for all types of transducers whose input-output characteristics do fall into two categories, i.e., exponentially rising and exponentially decaying. Thermistor coupled logarithmic amplifier is employed with FET as a circuit element to provide compensation for nonlinearities inherited from response curve of transducer. Further, it has been mentioned that it is difficult to invert the converted variable obtained after transduction.

Apart from the analog schemes reported so far in the literature for response linearization, there are some more general schemes which could not be accommodated in the earlier classifications. Trofimenkoff and Smallwood [112] suggested a simple circuit using junction field effect transistor (JFET) for response linearization of transducer and described a specific case of response linearization of thermocouple in range of ±50°C. Thermistor response linearization using single operational amplifier has been discussed and tested in narrow range from 10 to 50°C, by Stockert and Nave [106]. Trofimenkoff and Smallwood, [113] suggested a general technique for response linearization of transducer using analog multiplier in feedback circuit. It is well supported by mathematical analysis. The same circuit can be used to linearize higher order response functions (of transducers) by employing additional multipliers in feed back path. The analog multiplier linearization techique can be used to linearize thermocouple outputs in the range of ±50 °C for the quadratic system and ±100°C for the cubic system with errors less than ±0.1°C. Hoge [41] examined and compared eight linearizing circuits containing single thermistor, excluding logarithmic amplifiers circuits. All such circuits are found to be equal in linearizing capabilities. Theoretical findings were based on simplicity, stability, accuracy, and convenient adjustment of circuit elements. White [118] proposed an unified response linearization technique applicable to all resistance temperature detectors (RTDs). In this technique, by choosing appropriate feed back in circuit the linearization is achieved, e.g., negative feedback is given to the circuits employing NTC thermistor and nickel thermometer, where as positive feedback is applied for platinum resistance thermometer (PRTs). He presented practical results for platinum resistance thermometer in the range from -40 to 200°C with an error of less than 0.02°C. The proposed technique is once again based upon the method of Taylor's series expansion. Trofimenkoff and Nordquist [111] described a feedback technique employing single amplifier resistance bridge for response linearization of resistance temperature detectors. Test results for platinum resistance temperature have been given in range of -100 to 300°C. Sengupta [93] suggested thermistor response linearization using relaxation oscillator circuit in which fewer inexpensive and easily available circuit elements are used. With this approach the measuring accuracy was obtained to be better than 0.1 K. Diamond [28] described a theoretical method to improve upon the nonlinearity offered by platinum resistance thermometer. In this method, a second metal which will compensate the nonlinearity offered by platinum, is placed in series with platinum. Another generalized technique for linearizing transducer signal using ratiometric property of analog-to-digital converter is presented by Iglesias and Iglesias [44]. In this technique analog-to-digital conversion and linearization of the signal is achieved at the same time. This technique is applicable for linearization of transducer response, falling in monotonic concave upward or downward categories.

Having gone through sufficient literature for gain-linearization of thermistor employing analog (linearization) schemes, there is an open question as to which scheme promises a better linearization or which technique has got an edge over the other. There is still necessity of developing generalized techniques which may be used for all types of thermistors in their different ranges of measurements as per the requirement of their use. Some techniques have been developed for restricted range while the others are for wider ranges. There is not a single technique which can be used for all types of thermistors in their entire range of measurements. In the present work, two analog linearizing schemes, fully supported by theoretical analysis and experimental results, have been developed and tested for wider range of temperature measurement with less expansive and off-the-shelf electronic components.

Besides analog techniques, a number of digital techniques have also been developed for the linearization of input-output characteristics of transducers. Digital techniques are of generalized type and can be used for all types of response characteristics with little modification in their procedure of implementation. These provide higher accuracy and better noise rejection and make digital conversion fairly easy for multi-input data acquisition systems and data loggers, around microprocessors and microcomputers.

Kollataj and Harkonem [57] developed a digital method employing an integrated circuit for linearizing the response characteristics of both weak (from thermocouples and resistance temperature detectors) and strong nonlinear signals. In this method, linearization of error curve, i.e., the difference between nonlinear and desired linear characteristics, is approximated by straight line segments and the necessary correction according to slope at the segments (in the form of control bits) is applied via an IC logic card. Accuracies of $\pm 0.2°F$ upto $250°F$ and $\pm 1°F$ for high temperature upto $1500°F$ were achieved in the operating temperature range from 0 to $1500°F$. The method produces higher accuracy but is difficult to realize and at the same time is not exclusively software oriented. In an other digital technique described by Mayer, et al. [68], the characteristic parameters for each transducer are stored. This technique involves approximation of the error curve with a straight line polygon. An accuracy of 0.1 per cent of full span was achieved using the technique. In many cases, transfer function of transducers is not known explicitly and each transducer requires its own software routine for calculation. Krishnan [59] presented a theoretical method for correcting the gain-nonlinearlity in working region of thermistor using analog computer setup which is easily realizable. But the method is expensive and complex. Later on, Burke [17] developed another technique around microcomputer in which linearizing resistor was added either in parallel or in series with a thermistor (NTC) and linearity is attained between temperature-to-resistance and temperature-to-voltage. The maximum errors of ±0.01°C upto 10°C, ±0.5°C upto 30°C, ±0.6°C upto 50°C, and ±3°C upto 100°C were observed. The value of linearizing resistor is obtained with the help of corresponding resistances at three equidistant temperature points on R-T characteristic of the sensor. Brignell and Dorey [14] developed software technique using look-up table approach for compensation of nonlinearity offered by most of the transducers. The approach is entirely based on microprocessor but is sequential and slow. Thereafter, Bolk[10] presented a general digital technique for gain-linearization of transducer, around microprocessor - based system and accounted for manufacturing tolerances, zero offset and scaling errors. The test results are given for eight temperature sensors, including thermocouples, platinum and nickel thermometers, in range from 0 to 300°C and for thermistors (NTC and PTC) in range from 0 to 60°C. The superiority of the technique has been proved by comparing the results with those obtained earlier. Mahana and Trofimenkoff [66] discussed a method of corrections for input-output nonlinearities and for the effects of disturbing variables in transducers. This method made direct use of interpolating polynomials for generating multidimensional look-up table for performing correction with the help of 8-bit microprocessor. In another digital technique, calibration and response linearization of analog Hall-effect position sensor was done by Kondraske and Ramaswamy [58] over 360° angular range using look-up table technique around a microprocessor-based system with maximum error of 3.2 per cent of full scale value. Recently, Patranabis and Ghosh [77] and Ghosh and Patranabis [35] developed a new algorithm for linearizing the response characteristics of thermistor and alike sensors. The method has been experimentally verified in temperature range of 273 to 373K. This technique is confined to correct the nonlinearity of thermistors and other similar sensors which have exponentially decaying transfer response.

Realizing the importance of digital techniques for response linearization of transducers with microprocessor and microcomputer based data acquisition systems, a general digital technique has been developed for response linearization of transducers in this work.

## 1.4 ORGANIZATION OF THE THESIS

The work embodied in this thesis deals with the analysis and performance improvement of inductive transducers and thermistors. The first part, which is the major portion of the thesis, is devoted for the development of (improved) smart inductive transducers for the measurement of linear displacement. In the second part of the work, the problem of nonlinearity of responses of transducers is tackled. The thesis contains six chapters. A brief description of the contents of each chapter is given in this section as follows:

The first chapter deals with the introduction of inductive transducers and thermistors and also with the common problems encountered in their use in different instrumentation schemes. The review in chronological order of existing literature on inductive displacement transducers and NTC thermistors is presented in this chapter. New techniques to over come the problems with the aim to achieve improved performance are also suggested.

In the next chapter, analysis and development of three self-compensated inductive transducers, namely linear variable differential transformer, inductive ratio transducer and inductive differential transducer has been carried out. The performance has been evaluated and compared with that of the conventional uncompensated LVDT transducer. To achieve quotient of two analog output signals from the transducer, analog divider unit is discussed and its constructional details are given. In next two sections, two other types of inductive transducers, one based on ratio principle and other based on differential principle have been developed and their performances are evaluated under the influence of varying excitation conditions and change in environmental temperature. Experimental test results are presented for both transducers. Comments and discussions are also incorporated in the last section of this chapter.

Linearization of thermistor response using two new analog schemes is presented in third chapter. In one technique, thermistor is connected in feed-back path of an active bridge amplifier. The condition of linearity between input-output relation is obtained by carrying out detailed mathematical analysis. In second method, a logarithmic convertor in conjunction with the analog divider unit is used for gain-linearization. Experimental test results for linear input-output relationships of both the schemes are given. The critical discussions on constraints of the existing analog techniques developed by differnt workers are also given in this chapter.

Chapter IV deals with a microprocessor based multichannel temperature scanning and telemetry scheme. Review along with merits and demerits of existing methods for temperature measurement in electrical apparatus are presented in first section of the chapter. The output voltage of analog linearizing scheme, which has been dealt with in the third chapter, has been used for quick and continuous detection and monitoring and temperature from eight different locations in a sequential order. The snags of the earlier techniques have been minimized. Experimental results and discussions are included in the last section of this chapter.

In fifth chapter, a general digital technique has been developed using scale factor polynomial approach for linearization of response characteristics of thermistor (a case of exponentially decaying characteristic) and LVDT transducer (a case of exponentially rising characteristics). The new algorithm for one segment/single stroke as well as for piece-wise linearization is given. The method has been tested on main frame computer. In the last section of this chapter, the results in tabular form along with conclusions are given.

The last chapter is the concluding chapter of the thesis. The important and significant conclusions arrived at during the course of the work are summarized. The suggestions and possibilities for further work are also presented in this chapter.

# CHAPTER - II

# DEVELOPMENT OF SELF-COMPENSATED INDUCTIVE DISPLACEMENT TRANSDUCERS

## 2.1 GENERAL

Inductive transducers are commonly used for the measurement of physical quantities in response to applied physical inputs in different types of environments by making use of changes in reluctance, inductance, permeability and eddy current. Linear variable differential transformer and variable inductance displacement transducers are extensively used in industry since 1930 [42]. The performance of these transducers is adversely affected by variations in excitation conditions and changes in ambient and winding temperature. This makes them unsuitable for many applications. Thus, there has always been the necessity of development of smart and intelligent transducers which can provide a self-compensated output for undesirable input parameters. In the past, some effort has been put by the researchers to develop self-compensated inductive transducer [3]. This chapter deals with three self-compensated inductive displacement transducers, namely linear variable differential transformer transducer, inductive ratio transducer and inductive differential transducer which have been developed during the course of this work.

## 2.2 SELF-COMPENSATED LINEAR VARIABLE DIFFERENTIAL TRANSFORMER (L.V.D.T.) TRANSDUCER

LVDTs have good sensitivity, linearity, consistency of response and high reliability, good stability, and sufficient durability in their use over a wide range. The performance of the transducers is influenced considerably by their geometry, arrangement of primary and secondary coaxial windings, quality and configuration of core material, variation of electrical input parameters and changes in ambient and winding temperatures [30,62,74]. The primary excitation quantities influence the impedence and temperature distribution of coil assembly. The core permeability is also affected by the temperature changes and by variations in excitation current and frequency. LVDT, as such, in its primitive configuration is uncompensated transducer. Lion [62] suggested the use of negative temperature coefficient thermistor, in the input primary circuit of differential transformer transducer for minimizing the surrounding temperature effects. Efforts have been made by Ara [3] in developing a differential transformer which is insensitive to the variations in excitation parameters and to the changes in ambient and winding temperature. The undesired variations in the measurement of output of LVDT are minimized by employing temperature-sensitive elements such as thermistors, reference LVDT and associated electronic circuit. As usual, thermistors are used to provide compensation for temperature variations only. The use of second LVDT for providing reference output signal is not convenient, because in addition to space requirement for housing this individual assembly as close as possible with measuring transducer, it also increases the over all system weight and its cost. In addition to this, it necessarily requires the maintenance of environmental conditions, similar to those of measuring LVDT. The investigations to obtain self-compensated output from the transducer were carried out by him for a core travel of ±4 mm.

In the present work, a new concept of dual secondary (identical) windings for self-compensation is used in linear variable differential transformer against variations in excitation conditions and changes in ambient temperature in and around the differential transformer transducer assembly. After theoretically analyzing the basic equations, a self-compensated LVDT has been fabricated and tested. The testing has been carried out under both favourable and hostile environmental conditions and also for variations in excitation parameters. The details are given in the following section.

### 2.2.1 Principle of Compensation

Differential transformer transducer (usually plunger or aramture type) is frequently used to give the analogue of physical variables to outside world in the form of useful electrical signal. This type of transducer is known as a linear variable differential transformer. Fig.2.1 shows the schematic diagram of conventional uncompensated transducer

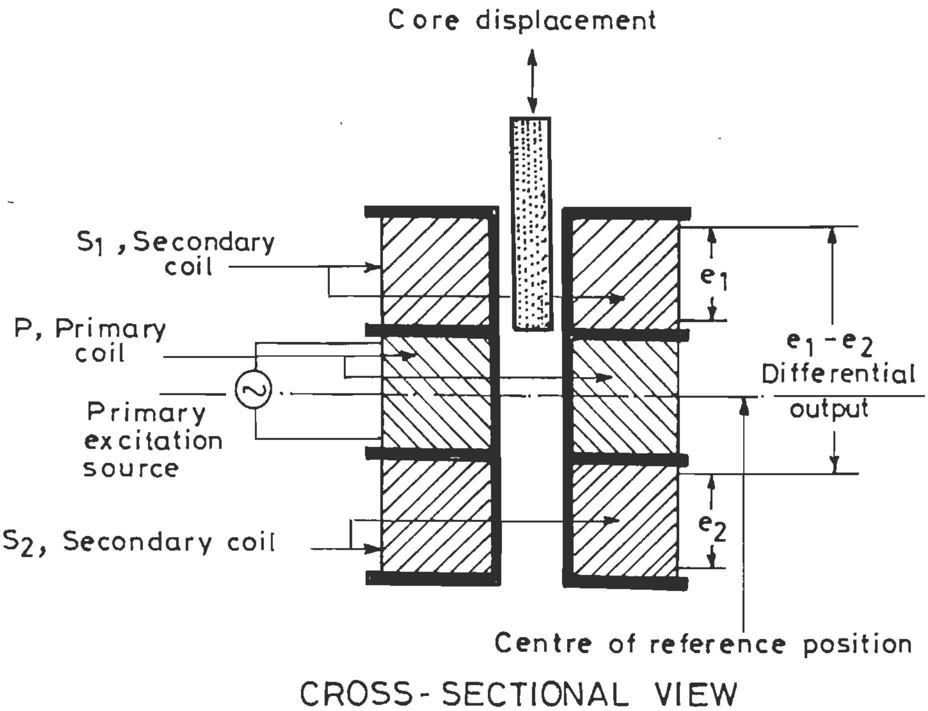



FIG. 2.1 UNCOMPENSATED LVDT TRANSDUCER

in its most elementry form. It consists of three coaxial and uniform coil windings, $S_1$,$P_1$, and $S_2$, wound over a hollow, non-magnetic and insulating former. Out of these, the primary winding is in the centre and the secondary windings are to either side of primary winding. The winding assembly is housed in a cylindrical case made of ferromagnetic material which also acts as a magnetic sheild. End rings (guides) are provided on both ends to provide friction-free and perfect axial movement of the core. An a.c. excitation, normally in the range of 3 to 15 volts r.m.s. at 60 to 20000 Hz is applied to the primary winding. A core of ferromagnetic material moves freely within the coil assembly in axial direction. In some design, the core is threaded for attachment to one of the sensing shafts shown in Fig.2.2. The alternating magnetic field induces voltages $e_1$ and $e_2$ in the nearly matched coils $S_1$ and $S_2$, respectively. To obtain net electrical output signal, i.e., $e_o = (e_1 - e_2)$, both secondaries are connected (externally) in series-opposition. With complete symmetry, the output signal is zero in the reference positon of the core, i.e., when the core is held stationary in the centre of the core travel. Core displacement causes a magnetic asymmetry. It is due to unsymmetrical magnetic coupling between primary and each of two secondaries as the core from null position causes a larger mutual coupling for one transducer coil and a smaller for the other. As a result, the output voltage experience a change in its magnitude and undergoes 180° phase shift at the output terminals while going through central position.

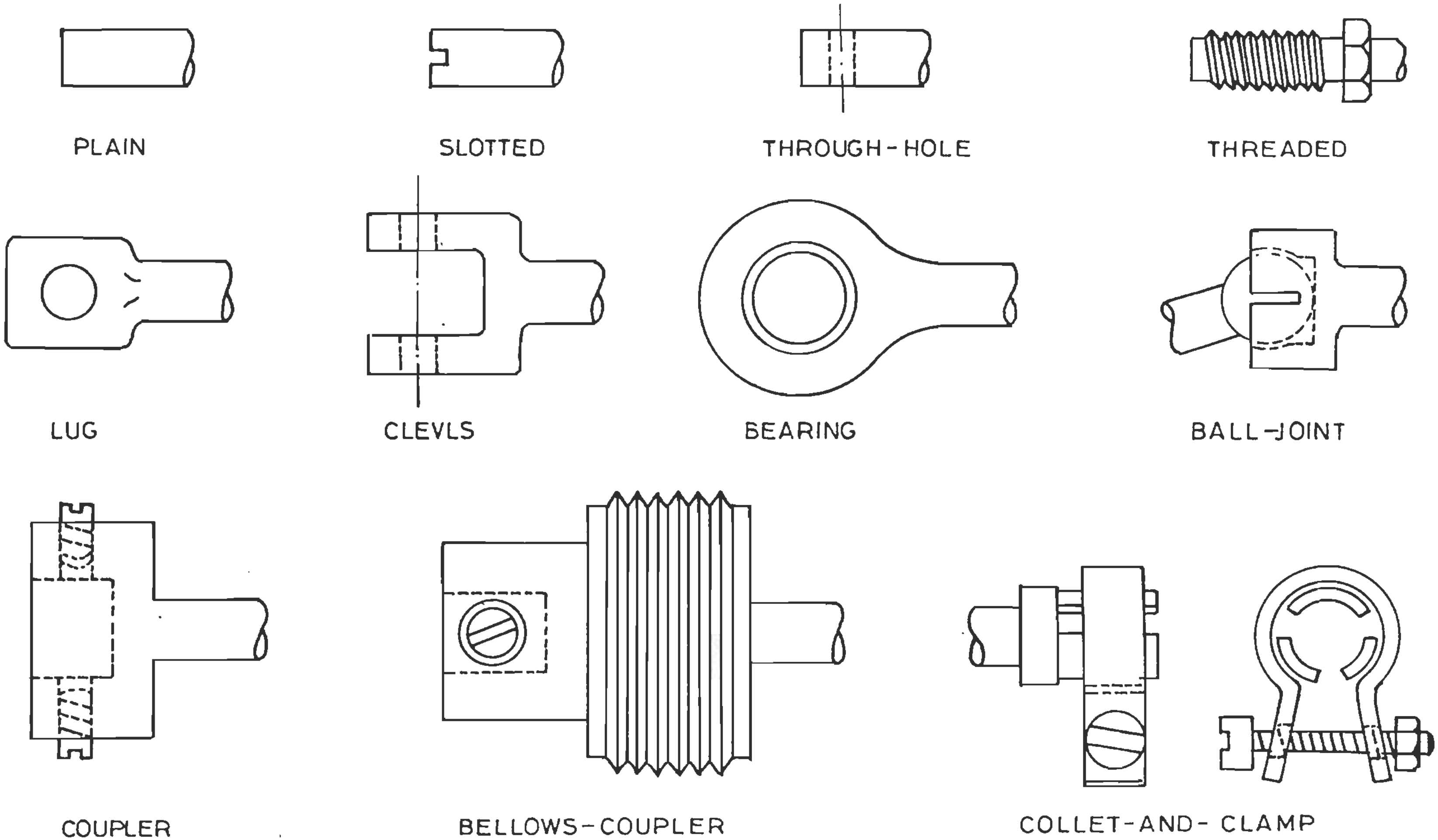



FIG. 2.2 TYPICAL DISPLACEMENT-TRANSDUCERS SHAFTS AND COUPLINGS

The response characteristic (output voltage $e_0$ versus core displacement d) is linear over a wide range of displacement on either side of the centre position and flattens out on both ends as shown in Fig.2.3 . As discussed earlier, the output voltage at the null position should remain zero, but it does not happen in practice. It is due to harmonics in excitation voltage and stray capacitance coupling between primary and secondary which results in a small but non-zero null voltage. It is also called as residual voltage. This is less than 1 per cent of maximum full scale output voltage in linear range and is acceptable. Methods of reducing the null voltage, when it is unaccepbtale are available in almost all the texts dealing with this topic [30,62,76].

The arrangement of the dual set secondary windings for providing self-compensation in conventional LVDT is shown in Figs.2.4(a) and (b). It has two sets of identical secondary windings $S_1$, $SC_1$ and $S_2$, $SC_2$. Under this arrangement of windings, the differential electrical output signal, $(e_1-e_2)$ is obtained from one set while, additive reference signal, $(e_1+e_2)$ is obtained from the other set.

For LVDT transducer, the induced output voltages, $e_1$ and $e_2$, are given as [3]

$$e_1 = k_1 F_1(\mu, I_p, f, x) \qquad \ldots (2.1)$$

$$e_2 = k_2 F_2(\mu, I_p, f, -x) \qquad \ldots (2.2)$$

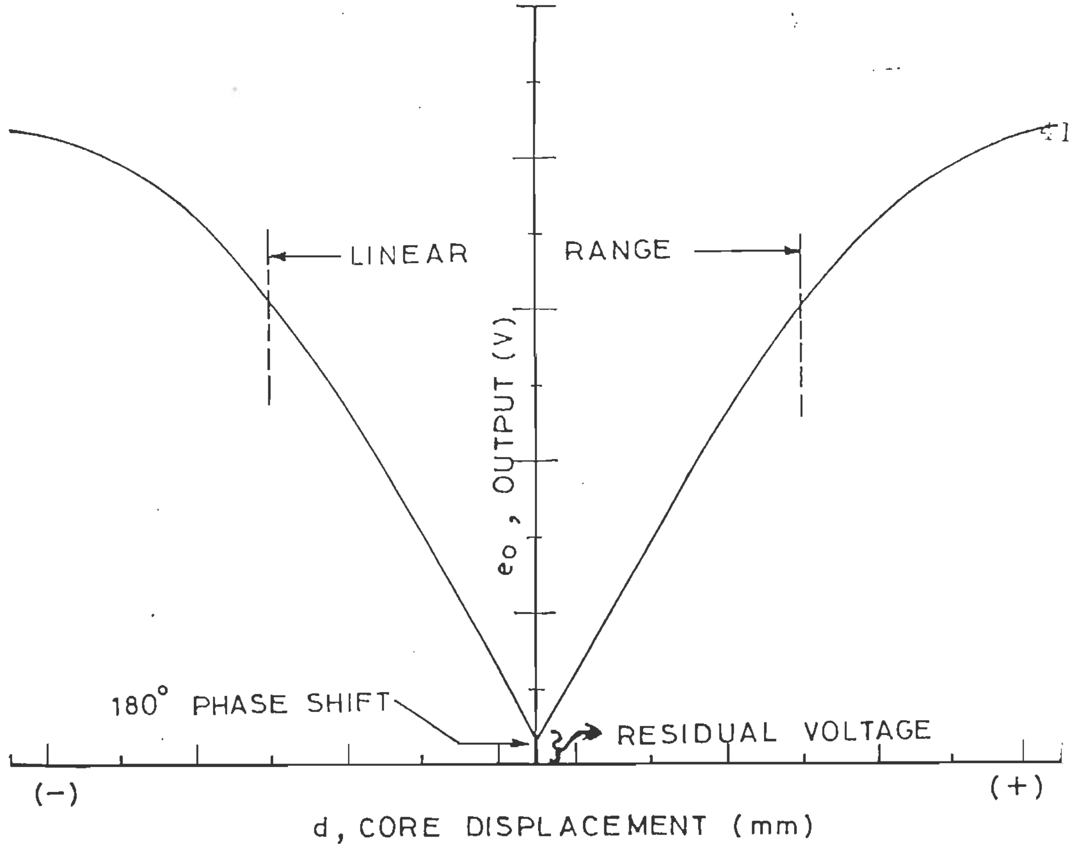



FIG. 2.3 RESPONSE CHARACTERISTICS OF UNCOMPENSATED LVDT TRANSDUCER

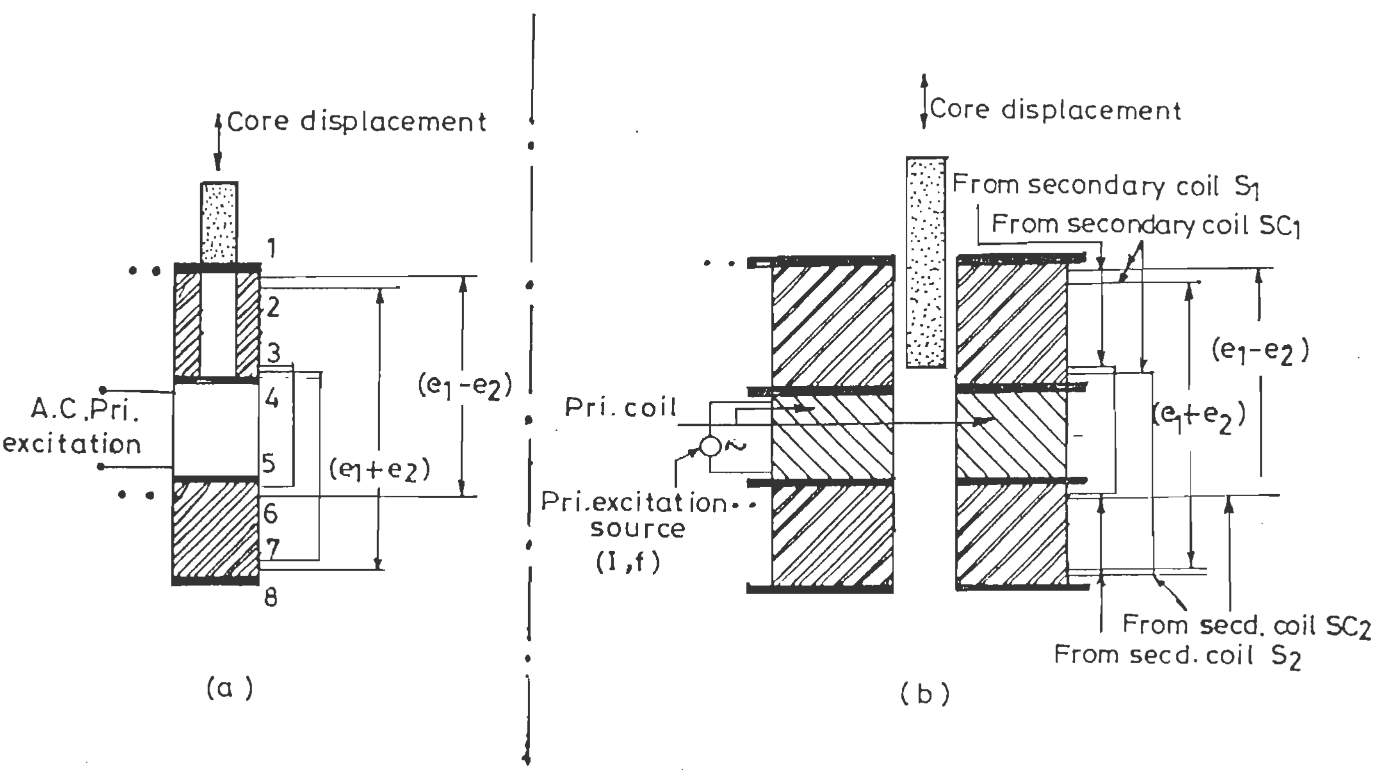



FIG. 2.4 SELF-COMPENSATED LVDT TRANSDUCER

and $\mu = \mu_0(I_p, T, f)$ ... (2.3)

where,

$e_1$ is the secondary voltage from $S_1$ and $SC_1$ coils,

$e_2$ is the secondary voltage from $S_2$ and $SC_2$ coils,

$\mu$ is the permeability of core material,

$I_p$ is the primary excitation current,

$f$ is the primary excitation frequency,

$N_p$ is the number of turns in primary winding,

$N_s$ is the number of turns in secondary winding,

$x$ is the displacement of core w.r.t. central (null) position of the core travel,

$T$ is the temperature, and

$k_1, k_2$ are the constants considering dimensions of coil, core, and other assembly and number of turns in primary and secondary windings.

For uncompensated LVDT, the differential output, is given as

$$(e_1 - e_2) = k_1 F_1(\mu, I_p, f, x) - k_2 F_2(\mu, I_p, f, -x) \quad \ldots (2.5)$$

If the (linear) displacement is in the normal operating range, then the variations of x can be separated in a function F(x). For a perfect symmetrical transducer assembly, $k_1 = k_2 = k$.

The terms can be separated as

$$F_1(\mu, I_p, T, x) = H(\mu, I_p, f).F(x) \quad \ldots (2.6)$$

$$F_2(\mu, I_p, T, -x) = H(\mu, I_p, f).F(-x) \quad \ldots (2.7)$$

and $\mu = \mu_0(I_p, T, f)$,

$$H(\mu, I_p, f).F(x) = J(T, I_p, f).F(x) \qquad \ldots (2.8)$$

$$H(\mu, I_p, f).F(-x) = J(T, I_p, f).F(-x) \qquad \ldots (2.9)$$

By making substitutions from equations (2.6 to 2.9) in equation (2.5), the differential output is obtained as

$$(e_1 - e_2) = kJ(T, I_p, f).[F(x) - F(-x)] \qquad \ldots (2.10)$$

It is apparent from equation (2.10) that the differential output voltage signal is dependent on ambient temperature (T), excitation current ($I_p$), and excitation frequency (f).

Like differential output signal, the additive output signal, is obtained as

$$(e_1 + e_2) = kJ(T, I_p, f).[F(x) + F(-x)] \qquad \ldots (2.11)$$

For self-compensated LVDT, the output is taken as quotient of differential and additive signals and is expressed as,

$$e\text{-output} = \left|\frac{(e_1 - e_2)}{(e_1 + e_2)}\right| = \frac{kJ(T, I_p, f).[F(x) - F(-x)]}{kJ(T, I_p, f).[F(x) + F(-x)]}$$

$$= \left|\frac{F(x) - F(x)}{F(x) + F(x)}\right| \qquad \ldots (2.12)$$

The output voltage of the self-compensated LVDT transducer expressed by equation (2.12) does not contain the terms influenced by T, $I_p$, and f. Thus, the output of tansducer becomes independent of influencing parameters unlike to that of equation (2.5). The term [F(x)+F(-x)] also remains constant as increase due to F(x) is equal to decrease due to F(-x) in the linear displacement range. The output signal, e-output, becomes dependent upon the differential (measuring) output signal which in turn is a function of core position [F(x)-F(-x)]. This self-compensation approach has been implemented by using a dual set of secondary windings. Here the secondaries are arranged exactly in parallel turn-by-turn on each side of the primary to obtain similar effects from them due to changes in environment and in excitatioin conditions. All windings are arranged coaxially and uniformly close to each other with possible perfection by keeping a thin insulation between them. The desired quotient of both the differential and reference signals is obtained using an IC analog divider unit without causing much complexity to the measuring system. The details of transducer assembly and electronic circuitry are given in next section.

### 2.2.2 Transducer Assembly and Electronic Circuitry

The approach has been tested by assembling both uncompensated and self-compensated LVDTs. The performance of these transducers are evaluated and are compared under adverse conditions of variations in excitation parameters and environmental temperature. The cross-sectional view of the compensated transducer with relevant dimensions is shown in Fig.2.5. The increased diameter of the secondaries due to dual set of arrangement makes all the difference in the design of this transducer compared to unconventional type. Due to

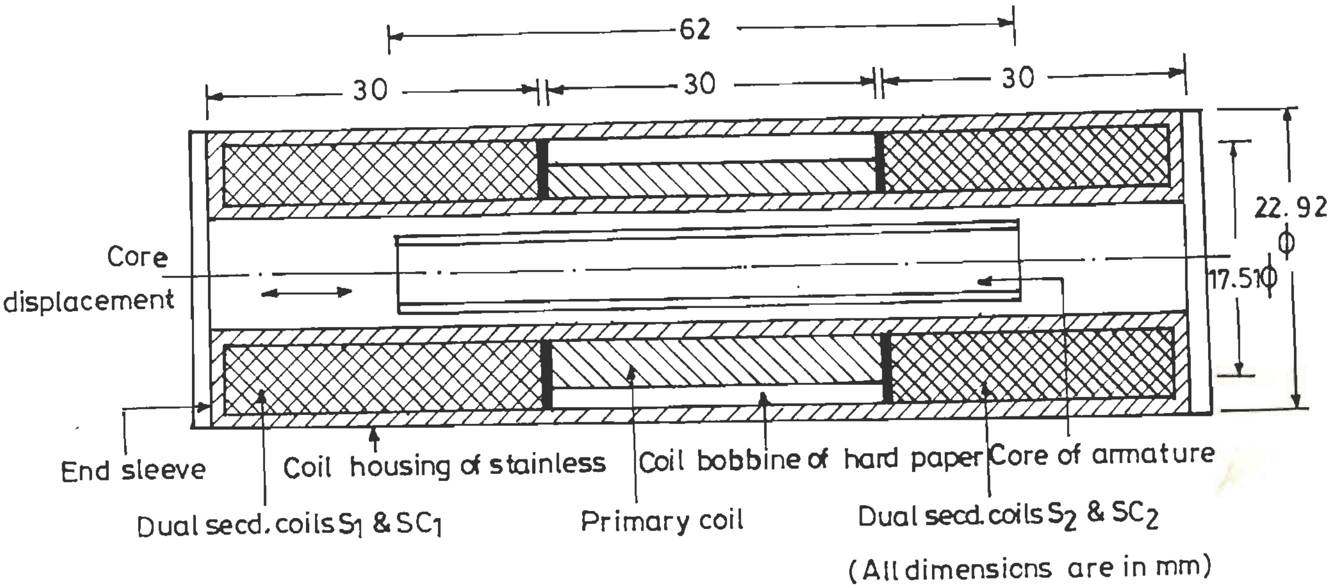
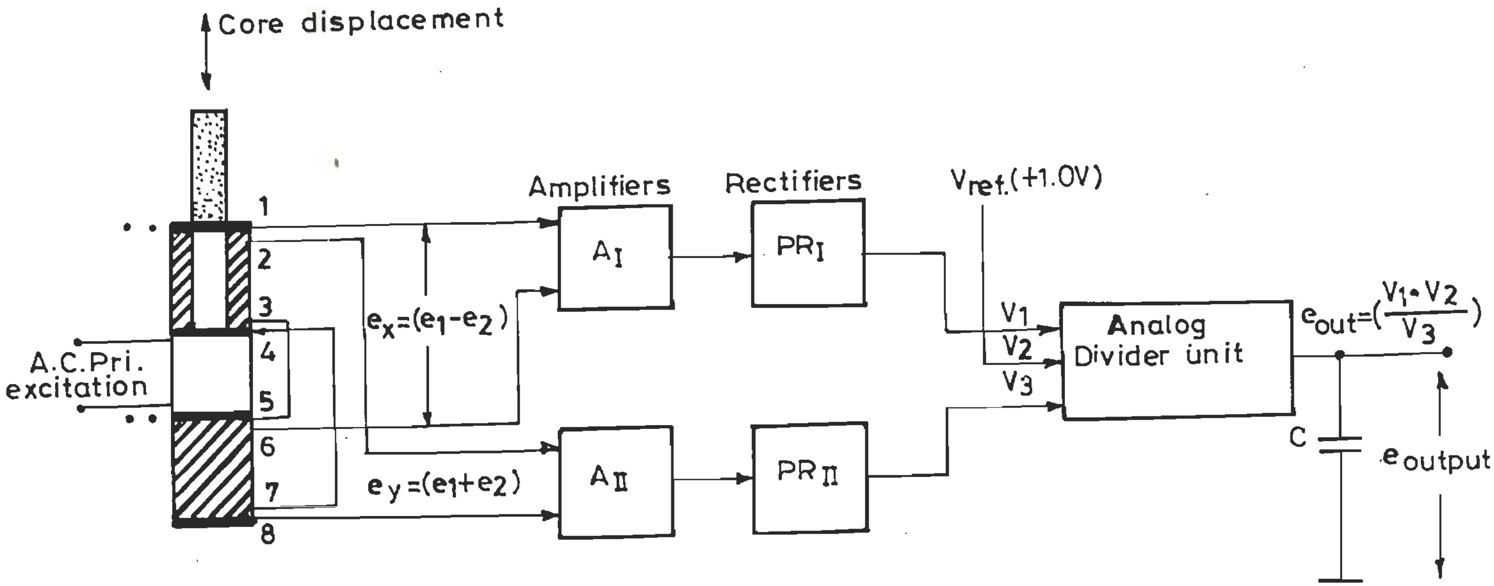



FIG. 2. 5 CROSS SECTION OF THE COMPENSATED LVDT USED FOR THE EXPERIMENTS

dual set of secondaries, the overall weight of the assembly is increased to a smaller extent. The 36 S.W.G., S.E.C. conductor is used for coil windings. The mandrel (bobbin) of the windings is made of hard paper conduit and the coil housing is made of ferromagnetic material so as to avoid stray magnetic field. The end-guides are provided for smooth and axial movement of core within the coil assembly. Connecting leads are brought out of this housing. As discussed in the earlier section, the coaxial coil windings experience similar effect of influencing parameters under varying test conditions to facilitate performance evaluation.

The complete block diagram representation of the instrumentation scheme is shown in Fig.2.6 . It consists of compensated LVDT, signal processing, signal conditioning, and analog divider unit. The output signals $(e_1-e_2)$ and $(e_1+e_2)$, from the compensated LVDT, are separately taken from main and compensating windings and are processed through amplifiers $A_I, A_{II}$ and precision rectifiers $P_{RI}$, $P_{RII}$ units as shown in Fig.2.6. After processing and conditioning, these signals are fed to the analog divider unit. Since the introduction of monolithic multipliers in 1969, these devices have proved their versatility and potentiality in mathematical computations [2,19,21,96]. Log-antilog method [107] has been used to develop analog IC divider unit around quad amplifier LM-324 in the present work. The log-antilog method depends on the mathematical relationship, as the sum of the logarithm




[$A_I$, $A_{II}$ ; Amplifiers , $PR_I$, $PR_{II}$ ; Precision Rectifiers]

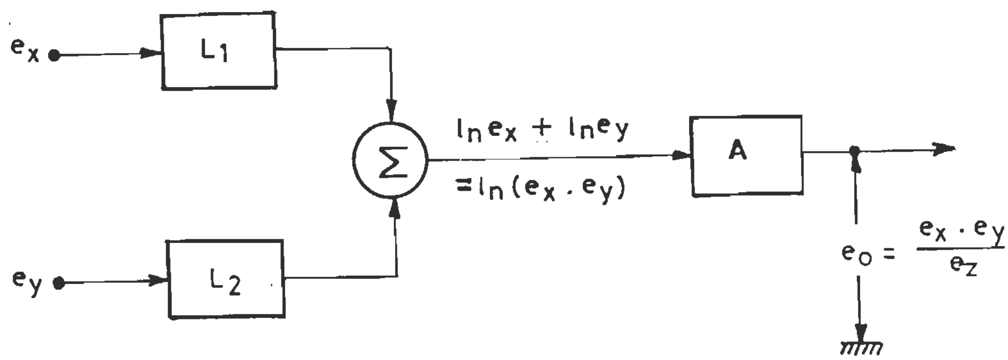
FIG. 2.6 CIRCUIT CONFIGURATION TO PERFORM SELF-COMPENSATION of two numbers equals the logarithm of the product of those numbers. Because logarithms are exponents, they are added to multiply and substracted to divide. The antilog then provides the desired result. Fig.2.7 shows the block-diagram representation of a log-antilog multiplier unit. Voltages $e_x$ and $e_y$ are inputs to $L_1$ and $L_2$ log amplifiers, respectively and resulting output of these amplifiers is addition of logarithms of the input voltages. which is expressed as

$$\ell_n e_x + \ell_n e_y = \ell_n(e_x \cdot e_y) \qquad \ldots (2.13)$$

The antilog amplifier A, serves dual purpose of undoing the logarithms and scaling the voltages. The net output voltage of this unit is $e_o$. By properly choosing the voltages, $e_x$, $e_y$, and $e_z$ the unit is used as divider, multiplier, or square-root generator. The output of the divider unit is expressed as

$$e_o = \frac{e_x \cdot e_y}{e_z} \quad \text{volt} \qquad \ldots (2.14)$$

If $e_y$ is held constant at +1.0 volt as reference signal, the unit works as divider unit and the output is proportional to $[e_x/e_z]$ volt. Further, if $e_z$ is held constant at a reference of +1.0 volt, this unit behaves as multiplier whose output voltage $e_o$ equals to $(e_x \cdot e_y)$. Another variation is achieved as square-root generator when $e_x = e_y$ and $e_z$ is again at a constant reference of +1.0 volt. Log-antilog dividers requires the input and reference voltages of positive polarity, and operation is restricted to only first quadrant. Four quadrant operation is possible by properly shifting the input and

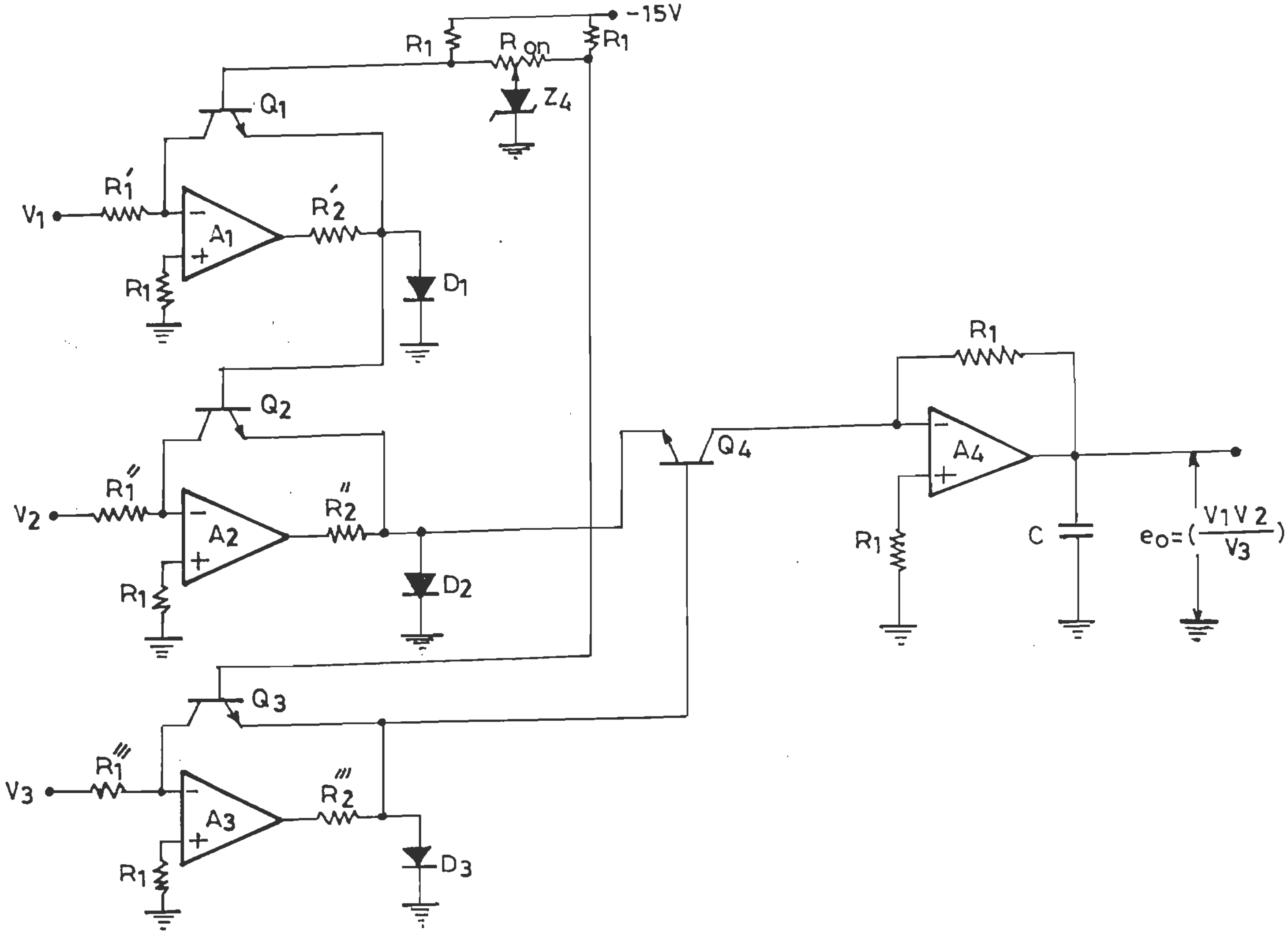



[$L_1$, $L_2$ ; Log amplifier, $A_1$ ; Antilog amplifier and scaling]

FIG. 2.7 LOG-ANTILOG MULTIPLIER / DIVIDER BLOCK DIAGRAM output levels. Log-antilog dividers / multipliers often have one-tenth of total error and lower drifts associated with a transconductance multiplier. But log-angilog divider is more expensive and slower than transconductance multiplier.

The analog divider unit built, as shown here in Fig.2.8. for obtaining the quotient of $(e_1 - e_2)$ and $(e_1 + e_2)$ has three single ended log amplifiers $A_1$, $A_2$, and $A_3$ and one antilog amplifier $A_4$,. $Q_1$, $Q_2$ and $Q_3$ are acting as logging transistors in the feedback loops of amplifiers. The collector current of transistor is equal to input voltage divided by the input resistor at the amplifier stage. The emitter-base junctions of $Q_1$ and $Q_2$ are connected effectively in series. This results in addition of logarithms, i.e., multiplication of $V_1$ and and $V_2$, where $V_1$ equals to $(e_1 - e_2)$ volts and $V_2$ equals to +1.0 volt (reference). Inut voltage $V_3$ to the third amplifier is equal to $(e_1 + e_2)$ volt, and its output is treated differently. It is connected in series with antilog transistor $Q_4$ and the combination is connected in parallel with the output of other two log amplifiers. The output of $A_3$ is substracted from the sum of log outputs of $A_1$ and $A_2$. Substraction of logarithmic quantities results in division. The collector current of transistor $Q_4$ is proportional to product of $V_1$ and $V_2$ divided by $V_3$. The output of antilog amplifier $A_4$ is the final result from analog divider unit and is expressed as

$$e_o = \left[ \frac{V_1 \cdot V_2}{V_3} \right] \text{ volt} \qquad \ldots (2.15)$$




[ $R_1 = R'_1 = R''_1 = R^*_1 = 10K$ ; $R'_2 = R''_2 = R'''_2 = 1K$ ; $R_{on} = 220\ \Omega$ , ($D_1 - D_3$); IN4148,
$Z_4$; LM 103, 2.4V, ($Q_1 - Q_4$); 2N 3728 (Matched pair), ($A_1 - A_4$); LM-324 ]

FIG. 2.8 PRECISION ANALOG DIVIDER

where,

$V_1$ is the $(e_1-e_2)$ volt,

$V_2$ is the +1.0 volt reference, and

$V_3$ is the $(e_1+e_2)$ volt.

The divider circuit has an accuracy of 1 per cent for an input voltage variation from 500 mV to 50 V provided input offsets of $A_1$,$A_2$ and $A_3$ are properly handled. The Zenner diode $Z_4$ increase the collector-base voltage across logging transistor to improve high current operation. The performance of divider unit is independent on change in temperature as variation in log amplifiers is compensated by equal change in antilog amplifier. The transistor must be identical (matched-pair) and must be kept in same housing to maintain identical environmental conditions. The offset null voltage for transistors is obtained by the resistor Ron. The gain is dependent on the operating levels of logging transistors. To limit the maximum loop gain, resistors $R_2'$, $R_2''$, and $R_2'''$ with compensation capacitor C are used. Resistors $R_1'$, $R_1''$, $R_1'''$ and $R_1$ are identical. The maximum limit of the reverse breakdown voltage of transistors $Q_1$,$Q_2$, and $Q_3$ for negative inputs is ensured by diodes $D_1$,$D_2$, and $D_3$. Instead of going for building analog divider with few opamps and discrete components, packaged multiplier/divider, AD 533 and AD 534 are also available. But these devices were not available here at the time of experimentation.

### 2.2.3 Experimental Results and Discussions

Figs.2.9 and 2.10(a-c) show the response characteristics of the uncompensated and self-compensated LVDT transducers at excitation levels of 3,4, and 5 volts at 1 kHz for different values of linear displacement in both positive and negative directions w.r.t. null position of the core travel for ambient temperature at 290, 323 and 343 K. The response characteristics of self-compensated transducer have been plotted separately for each level of excitation. It is not possible to show them clearly on the same graph due to overlapping. The response of self-compensated type is more symmetrical in nature and is immune to variations in excitation levels and ambient temperature. The response of uncompensated LVDT has larger variations due to influence of undesired parameters compared to that of self-compensated type.

Table-I shows variations in sensitivity for both the uncompensated and self-compensated types of transducers. Negligible variation in sensitivity for self-compensated LVDT is observed only at third place after decimal for the similar values of core displacement in either directions w.r.t. to null position. It may be due to human observational error in noting down the displacement or the output voltage. As such high degree of stability and reproducibility of sensitivity is reflected by the values given in the Table-I for compensated LVDT.

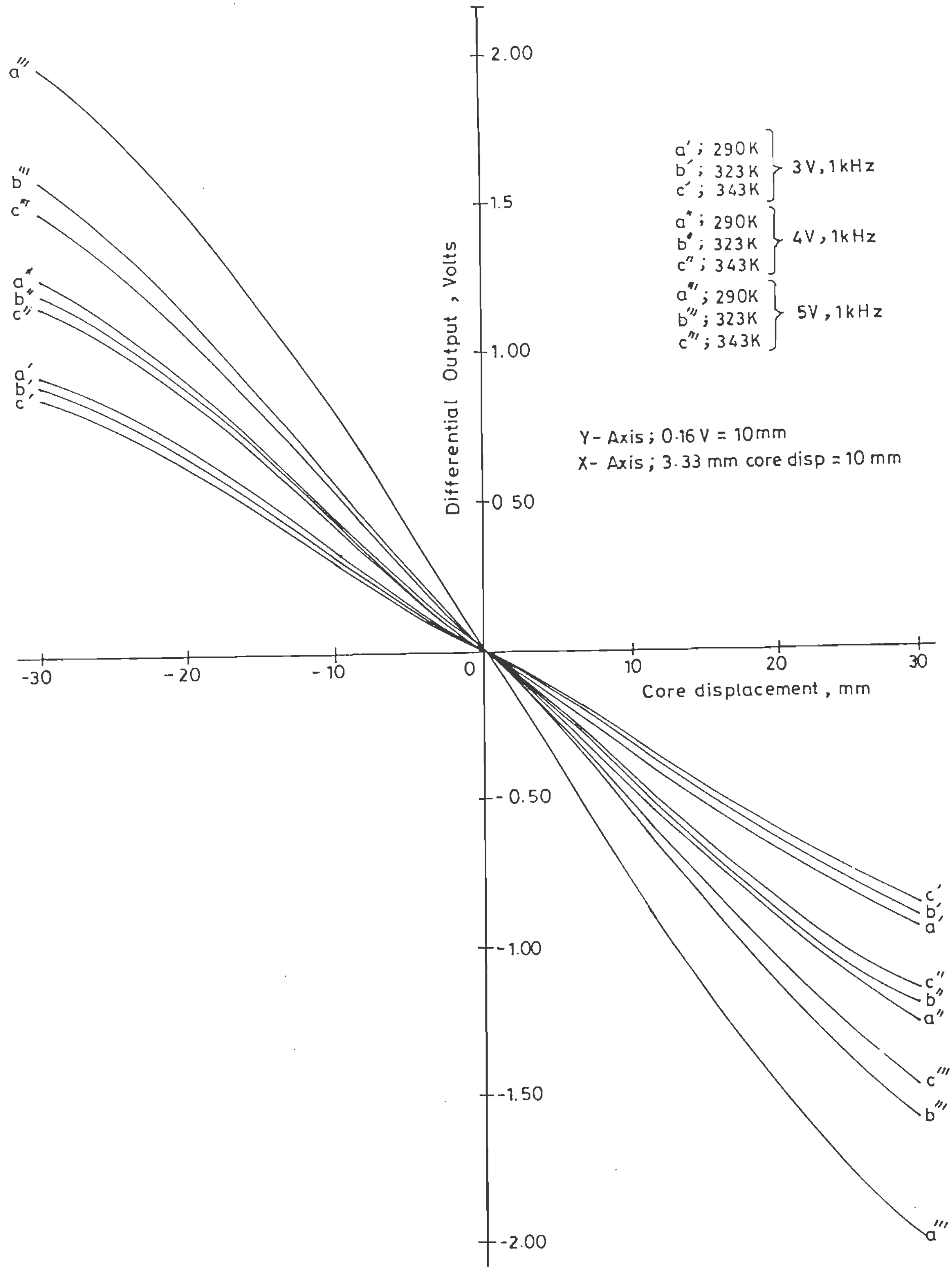
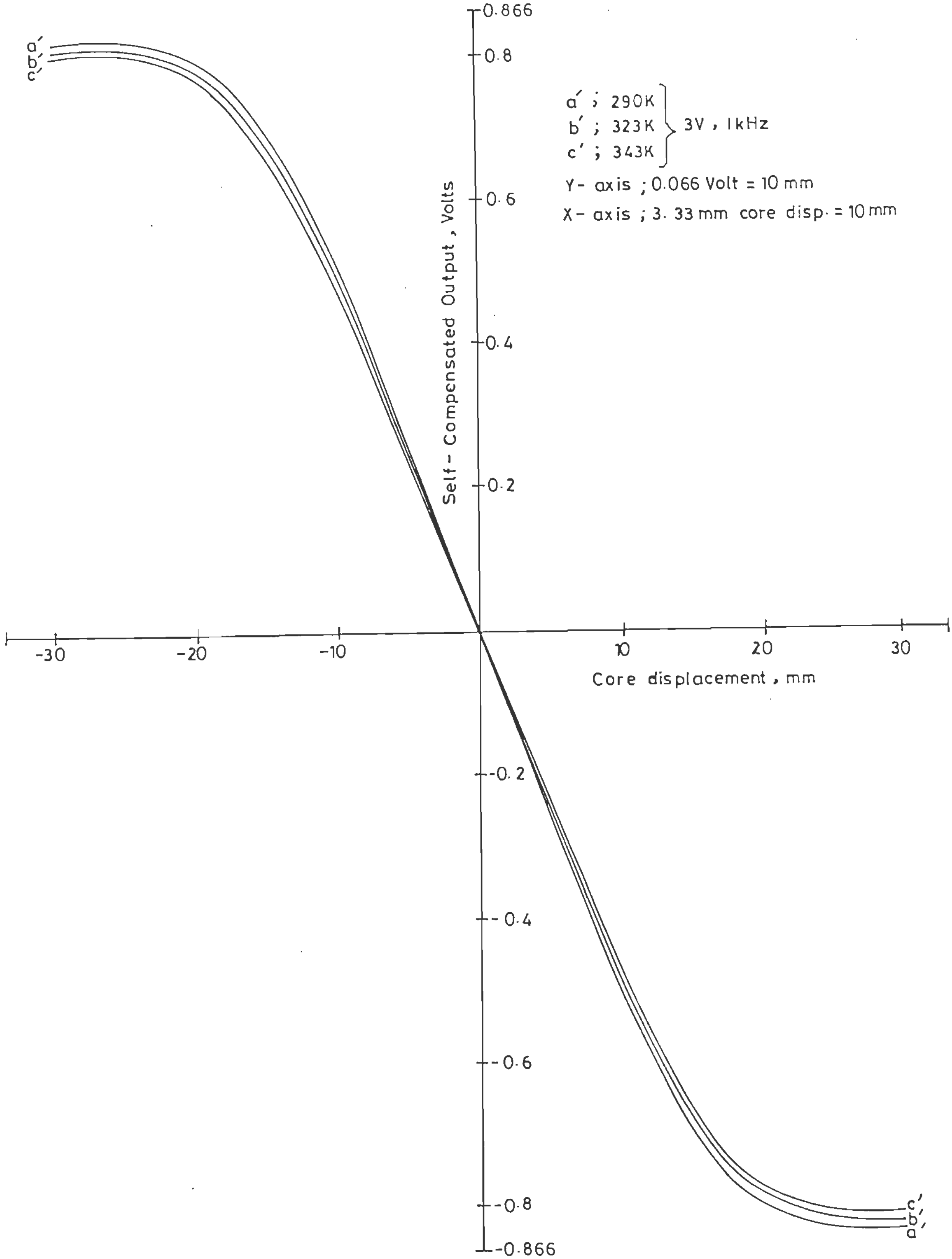



FIG.2.9 OUTPUT RESPONSE VS CORE DISPLACEMENT OF UNCOMPENSATED LVDT FOR DIFFERENT EXCITATION LEVELS AT 290K 323K AND 343K




FIG·2.10(a) OUTPUT RESPONSE VS CORE DISPLACENT OF SELF-COMPENSATED LVDT FOP 3VOLTS, 1 kHz AT 290K, 323K AND 343K

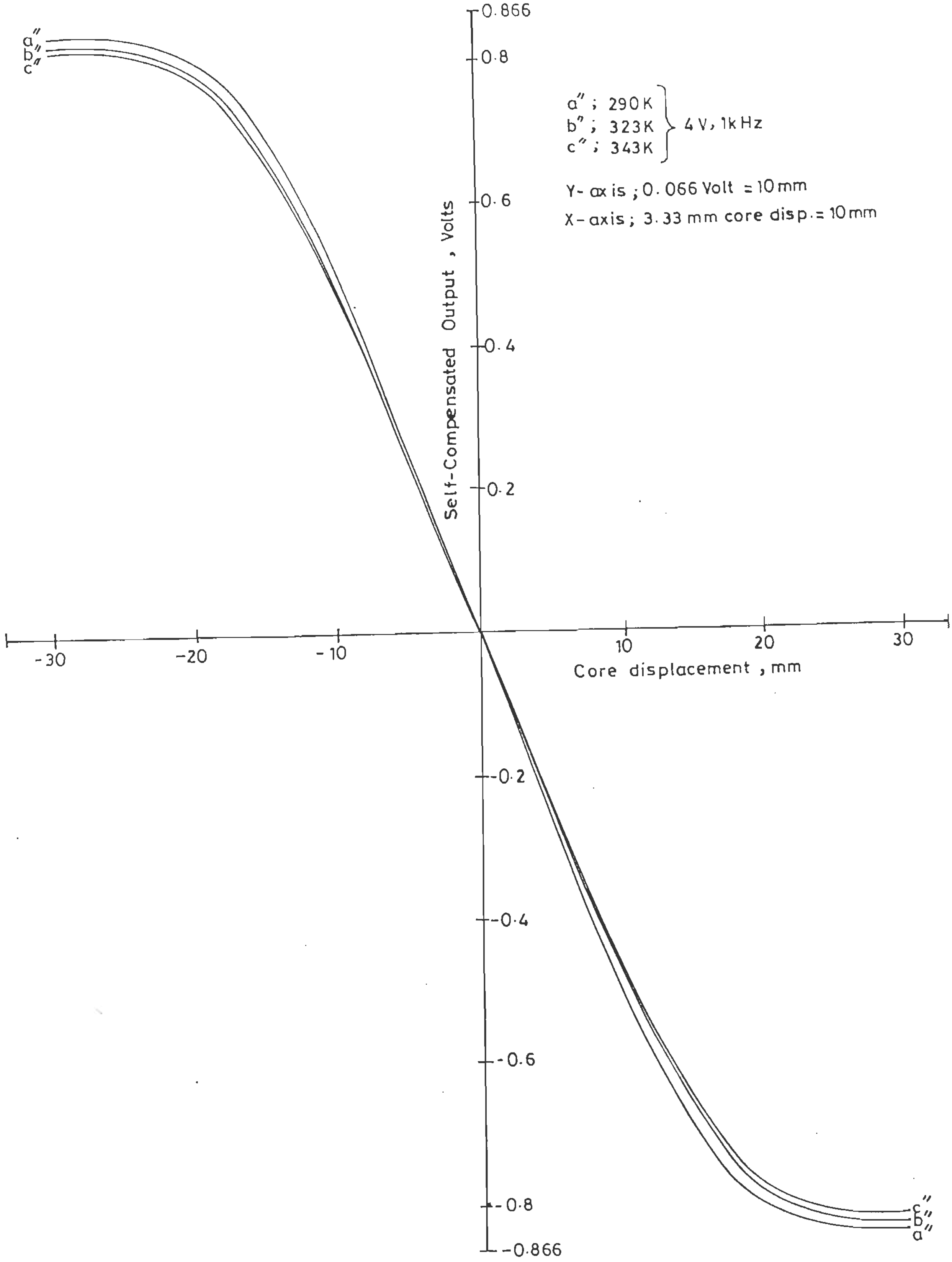



FIG. 2.10(b) OUTPUT RESPONSE VS CORE DISPLACEMENT OF SELF-COMPENSATED LVDT FOR 4 VOLTS, 1kHz AT 290K, 323K AND 343K

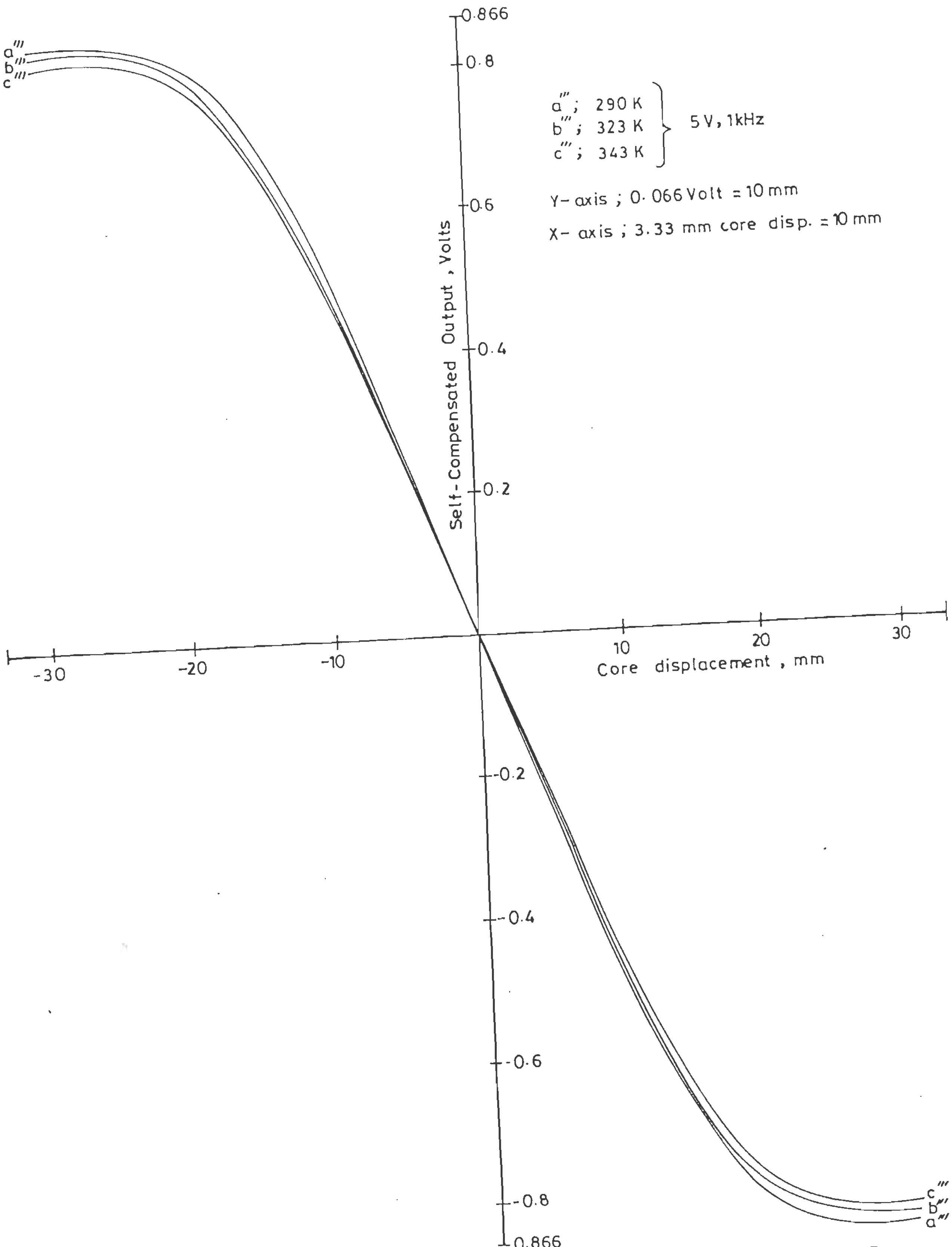



FIG. 2.10 (c) OUTPUT RESPONSE VS CORE DISPLACEMENT OF SELF-COMPENSATED LVDT FOR 5VOLTS, 1kHz AT 290K, 323K AND 343K

TABLE - I

| | | VARIATION IN SENSITIVITY; ($\frac{volt}{mm}$) | | | | | | | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | Uncompensated LVDT | | | | | | | | | Self-compensated LVDT | | | | | | | | |
| | | Temperature (K) | | | | | | | | | Temperature (K) | | | | | | | | |
| | | 290 | | | 323 | | | 343 | | | 290 | | | 323 | | | 343 | | |
| Sl. No. | Excitation level / Core displacement mm | 3V, 1kHz | 4V, 1kHz | 5V, 1kHz | 3V, 1kHz | 4V, 1kHz | 5V, 1kHz | 3V, 1kHz | 4V, 1kHz | 5V, 1kHz | 3V, 1kHz | 4V, 1kHz | 5V, 1kHz | 3V, 1kHz | 4V, 1kHz | 5V, 1kHz | 3V, 1kHz | 4V, 1kHz | 5V, 1kHz |
| 1. | 10.00 | 0.034 | 0.047 | 0.080 | 0.032 | 0.045 | 0.058 | 0.030 | 0.044 | 0.054 | 0.052 | 0.052 | 0.053 | 0.051 | 0.050 | 0.051 | 0.049 | 0.050 | 0.050 |
| 2. | 20.00 | 0.034 | 0.046 | 0.074 | 0.032 | 0.045 | 0.057 | 0.031 | 0.044 | 0.053 | 0.040 | 0.040 | 0.041 | 0.039 | 0.039 | 0.040 | 0.039 | 0.039 | 0.039 |
| 3. | 30.00 | 0.032 | 0.042 | 0.066 | 0.030 | 0.040 | 0.053 | 0.028 | 0.038 | 0.049 | 0.027 | 0.028 | 0.028 | 0.027 | 0.027 | 0.028 | 0.027 | 0.027 | 0.027 |
| 4. | -10.00 | 0.035 | 0.049 | 0.081 | 0.034 | 0.045 | 0.060 | 0.031 | 0.043 | 0.053 | 0.052 | 0.052 | 0.051 | 0.049 | 0.051 | 0.051 | 0.050 | 0.051 | 0.051 |
| 5. | -20.00 | 0.033 | 0.044 | 0.072 | 0.033 | 0.044 | 0.057 | 0.031 | 0.043 | 0.052 | 0.041 | 0.040 | 0.039 | 0.039 | 0.041 | 0.041 | 0.040 | 0.040 | 0.039 |
| 6. | -30.00 | 0.033 | 0.044 | 0.068 | 0.029 | 0.042 | 0.051 | 0.029 | 0.040 | 0.051 | 0.029 | 0.027 | 0.029 | 0.028 | 0.028 | 0.029 | 0.027 | 0.028 | 0.028 |

Fig. 2.11 (a), shows the sensitivity change for both types of transducers w.r.t. variations in excitation level for positive core displacement. The response is plotted for normalized sensitivity (S/So) vs excitation voltages. for a reference value of 3V at a frequency of 1 kHz at room temperature of (290 K). the normalized sensitivity of uncompensated type is within -0.469 to +1.062, whereas for compensated type is within -0.38 to +0.037 w.r.t. normalized axis for extreme core travel at +30mm displacement.

Fig.2.11(b) shows the sensitivity changes for both the transducer w.r.t. variations in excitation level for negative core displacement. The response is plotted for normalized sensitivity vs excitation voltages for a reference value of 3 volt at a frequency of 1 kHz and room temperture of 290 K. The normalized sensitivity of uncompensated type is within -0.0485 to +1.060, whereas for compensated LVDT, is only within -0.069 to 0.0 w.r.t. normalized axis for extreme core position at -30 mm displacement.

The influence of variation of excitation voltage from 1 to 5 volts, keeping the core displacement at a particular value, is observed, and the sensitivity and normalized sensitivity (S/So) values, their maximum variations for both types of the transducers are obtained and are shown in Table-II. It is clear from the table that self-compensated LVDT is highly insensitive to variations in excitation conditions at a constant excitation frequency.

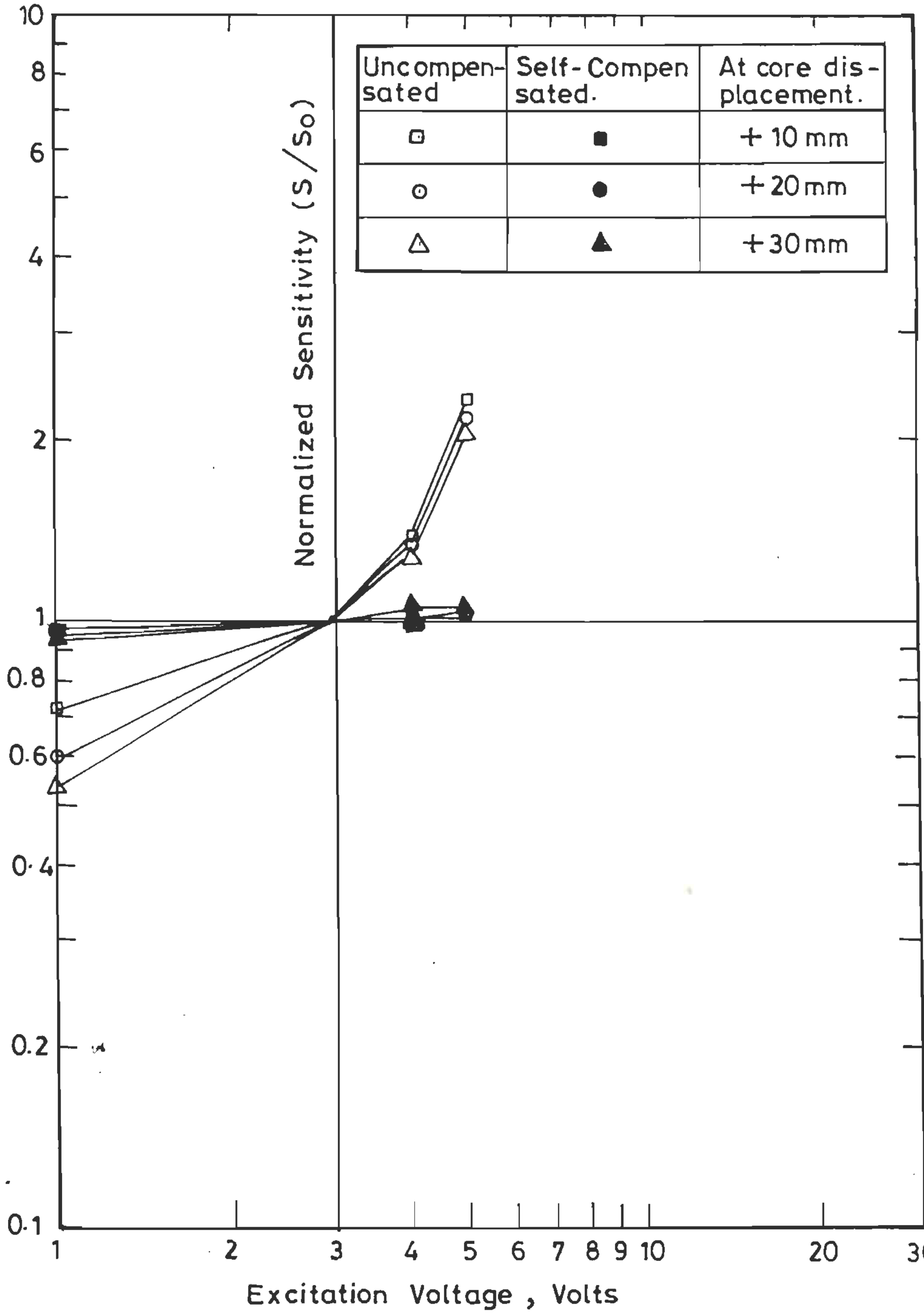



FIG.2.11(a) NORM. SENSITIVITY CHANGES IN UNCOMPENSATED AND SELF-COMPENSATED LVDT OUTPUTS DUE TO VARIATION IN EXCITATION VOLTAGE AT 1KHz AND ROOM TEMPERATURE 290 K FOR (+ve) CORE DISPLACEMENT (SENSITIVITY AT 3V IS TAKEN AS REFERENCE)

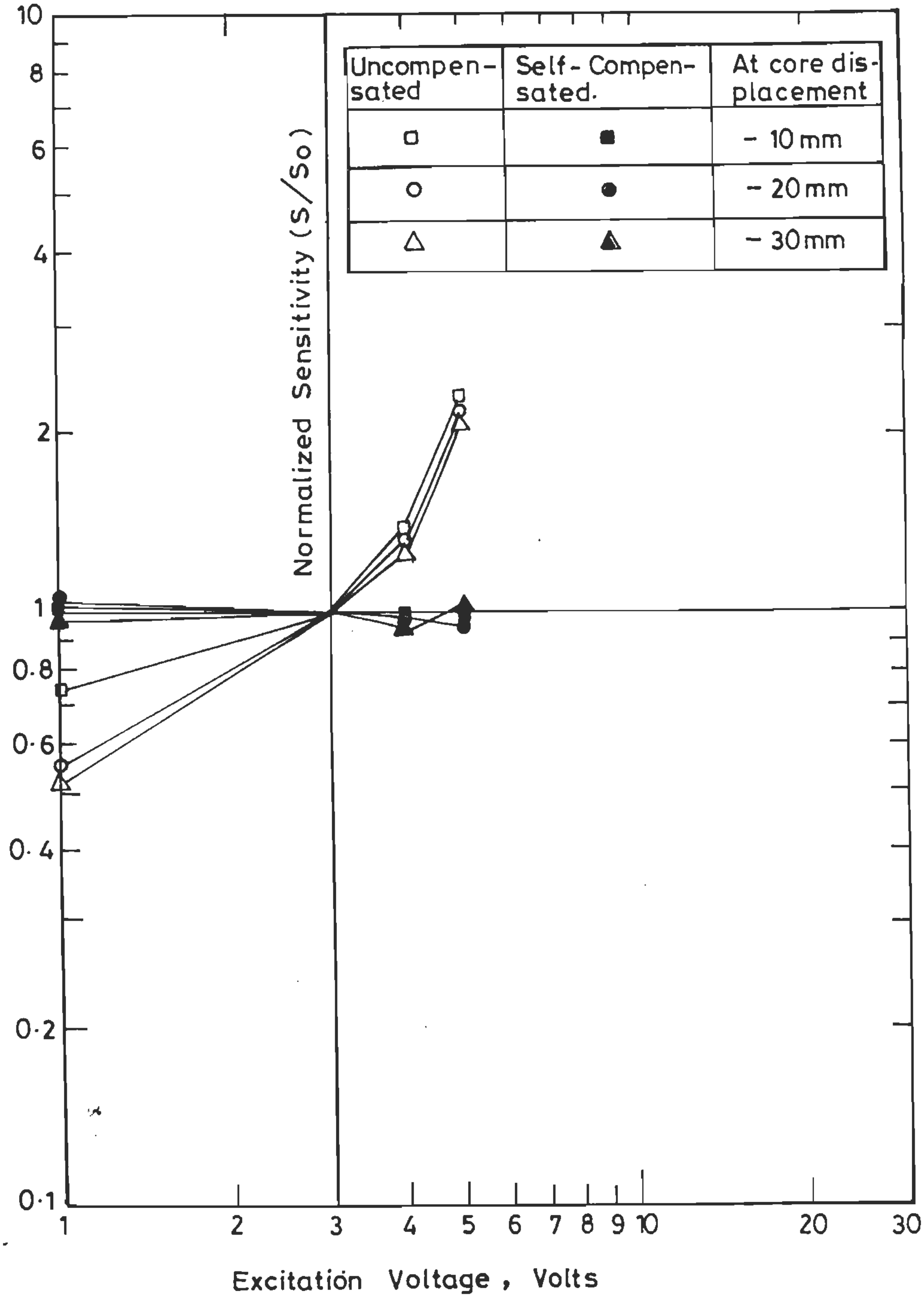



FIG. 2.11(b) NORM. SENSITIVITY CHANGES IN UNCOMPENSATED AND SELF-COMPENSATED LVDT OUTPUTS DUE TO VARIATION IN EXCITATION VOLTAGE AT 1 KHz AND ROOM TEMPERATURE 290 K FOR (-ve) CORE DISPLACEMENT (SENSITIVITY AT 3V IS TAKEN AS REFERENCE)

TABLE - II

| Sl. No., | | Core Displacement | Excitation Voltage (volts) at | Sensitivity for change in voltage from 1-5 volts, 1kHz, 290 K | | Normalized sensitivity (S/So) at 3 volts 1kHz, 290 K | | Max. variation in normalized sensitivity (S/So) w.r.t. reference value at 3 volts. | |
|---|---|---|---|---|---|---|---|---|---|
| | | | | Uncompensated LVDT (volt/mm) | Self-compented LVDT (volt/mm) | Uncompensated LVDT | Self-compensated LVDT | Uncompensated LVDT | Self-compensated LVDT |
| 1. | a | 10.00 | 1.0 | 0.024 | 0.051 | 0.705 | 0.980 | 1.352 | 0.020 |
| | b | | 3.0 | 0.034 | 0.052 | 1.000 | 1.000 | | |
| | c | | 4.0 | 0.047 | 0.052 | 1.382 | 1.000 | | |
| | d | | 5.0 | 0.080 | 0.053 | 2.352 | 1.019 | | |
| 2. | a | 20.00 | 1.0 | 0.020 | 0.039 | 0.588 | 0.975 | 1.176 | 0.025 |
| | b | | 3.0 | 0.034 | 0.040 | 1.000 | 1.000 | | |
| | c | | 4.0 | 0.046 | 0.040 | 1.352 | 1.000 | | |
| | d | | 5.0 | 0.074 | 0.041 | 2.176 | 1.025 | | |
| 3. | a | 30.00 | 1.0 | 0.017 | 0.026 | 0.531 | 0.962 | 1.062 | 0.038 |
| | b | | 3.0 | 0.032 | 0.027 | 1.000 | 1.000 | | |
| | c | | 4.0 | 0.042 | 0.028 | 1.312 | 1.037 | | |
| | d | | 5.0 | 0.066 | 0.028 | 2.062 | 1.037 | | |
| 4. | a | -10.00 | 1.0 | 0.026 | 0.053 | 0.742 | 1.019 | 1.314 | -0.02 |
| | b | | 3.0 | 0.035 | 0.052 | 1.000 | 1.000 | | |
| | c | | 4.0 | 0.049 | 0.052 | 1.400 | 1.000 | | |
| | d | | 5.0 | 0.081 | 0.051 | 2.314 | 0.0980 | | |
| 5. | a | -20.00 | 1.0 | 0.018 | 0.042 | 0.545 | 1.024 | 1.181 | -0.024 |
| | b | | 3.0 | 0.033 | 0.041 | 1.000 | 1.000 | | |
| | c | | 4.0 | 0.044 | 0.040 | 1.333 | 0.975 | | |
| | d | | 5.0 | 0.072 | 0.039 | 2.181 | 0.951 | | |
| 6. | a | -30.00 | 1.0 | 0.017 | 0.028 | 0.515 | 0.965 | 1.060 | 0.035 |
| | b | | 3.0 | 0.033 | 0.029 | 1.000 | 1.000 | | |
| | c | | 4.0 | 0.044 | 0.027 | 1.333 | 0.931 | | |
| | d | | 5.0 | 0.068 | 0.029 | 2.060 | 1.000 | | |

Fig.2.12(a) illustrates the plot between normalized sensitivity and variation in excitation frequency from 100 Hz to 2.5 kHz for a reference value of 1 kHz at 3 volt and room temperature of 290 K. The variation in normalized sensitivity of the uncompensated type is within -0.219 to +0.343, whereas of self-compensated is only within +0.111 to -0.112 w.r.t. normalized axis for an extreme core position of +30 mm. The plot for normalized sensitivity vs variation in excitation frequency from 100 Hz to 2.5 kHz for a reference value of 1 kHz at 3 volt and room temperture 290 K is shown in Fig.2.12(b). The variation in (S/So) value of the uncompensated LVDT is within -0.243 to 0.363, whereas of compensated type is only within -0.207 to 0.103, w.r.t. normalized axis for the extreme core position of -30 mm displacement.

Table-III shows the results of sensitivity and normalized sensitivity for variation in excitation frequency from 100-2500 Hz for a fixed value of excitation voltage at room temperature. It also shows the maximum variations of sensitivity and normalized sensitivity values for these transducers. Evidently and without reservation it also supports that self-compensated LVDT transducer has better performance stability compared to uncompensated type for the changes in influencing parameter.

Fig. 2.13, shows the variation in normalized sensitivity with a change in ambient temperature from 290 to 343 K for fixed excitation voltage of 5V and frequency of 1 kHz for

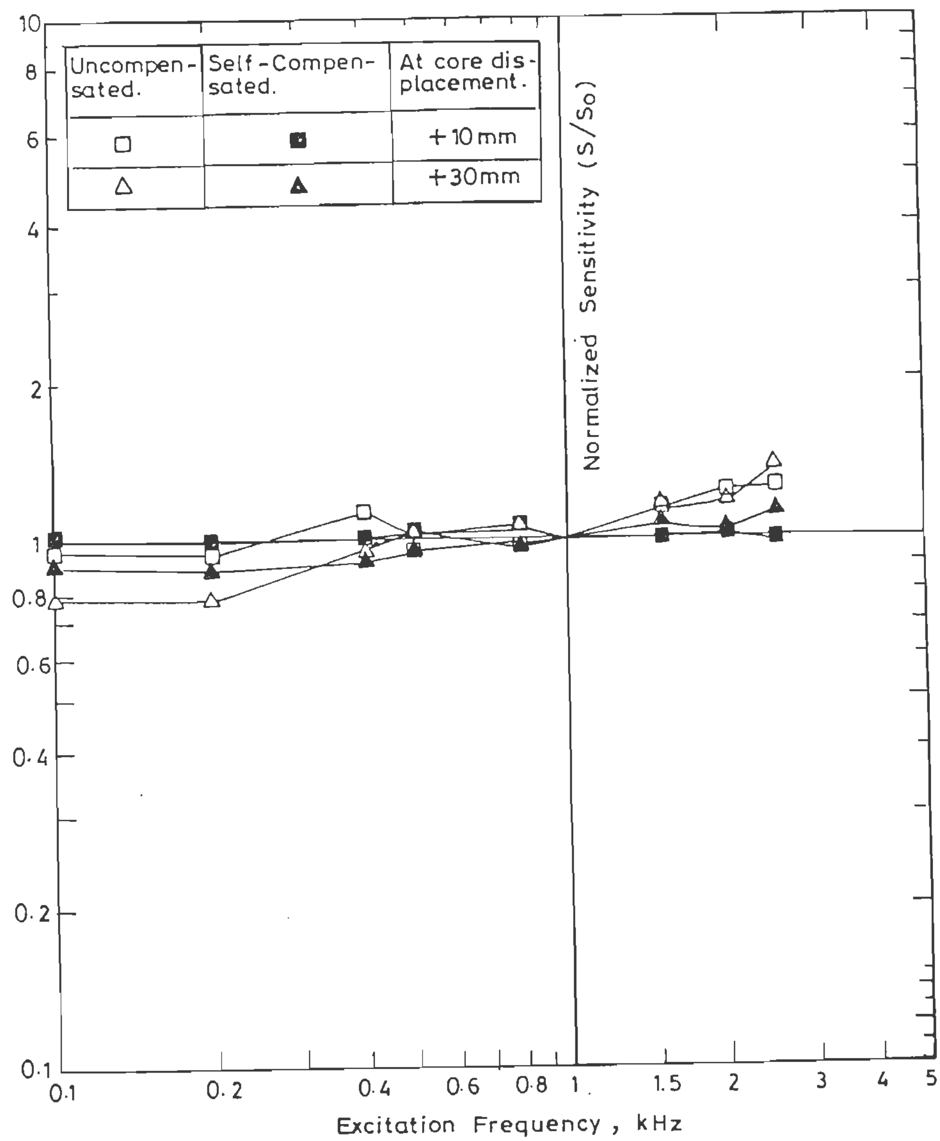



FIG. 2.12(a) NORM. SENSITIVITY CHANGES IN UNCOMPENSATED AND SELF-COMPENSATED LVDT OUTPUTS DUE TO VARIATION IN EXCITATION FREQUENCY AT 3V AND ROOM TEMPERATURE 290 K FOR (+ve) CORE DISPLACEMENT (SENSITIVITY AT 1kHz IS TAKEN AS REFERENCE)

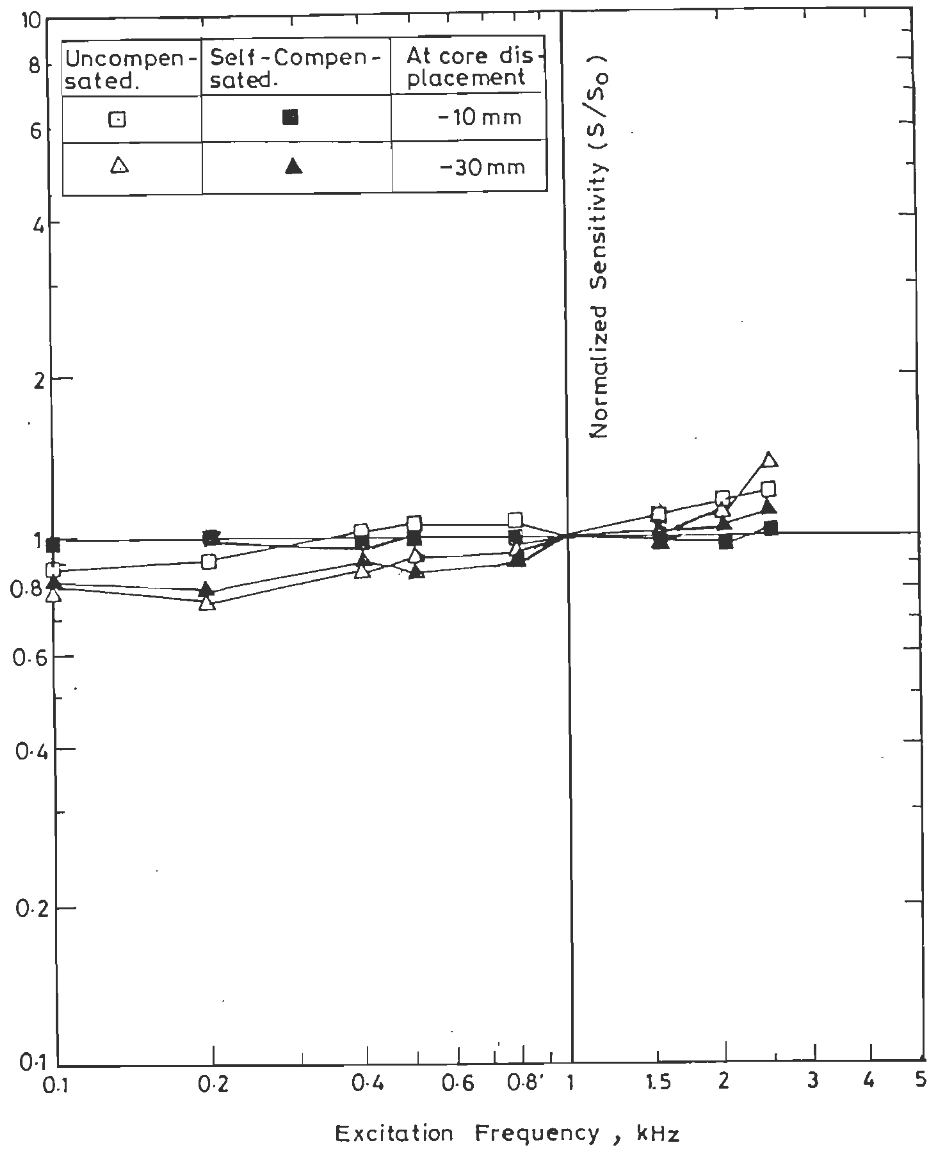



FIG. 2.12 (b) NORM. SENSITIVITY CHANGES IN UNCOMPENSATED AND SELF-COMPENSATED LVDT OUTPUTS DUE TO VARIATION IN EXCITATION FREQUENCY AT 3V AND ROOM TEMPERATURE 290K FOR (-ve) CORE DISPLACEMENT (SENSITIVITY AT 1kHz IS TAKEN AS REFERENCE)

TABLE - III

| Sl. No. | | Core Displacement | Excitation Frequency (Hz), at 3V,290K | Sensitivity for change in frequency from 100-1500 Hz, at 3 volt, 290 K | | Normalized sensitivity (S/So) at 1 kHz, 3 volts 290 K | | Max. variation in Normalized Sensitivity (S/So) w.r.t. reference value at 1 kHz 3v, 290 K | |
|---|---|---|---|---|---|---|---|---|---|
| | | | | Uncompensated LVDT (volt/mm) | Self-compensated LVDT (volt/mm) | Uncompensated LVDT (volt/mm) | Self-uncompensated (volt/mm) | Uncompensated LVDT | Self-compensated LVDT |
| | a | | 100 | 0.033 | 0.053 | 0.970 | 1.019 | | |
| | b | | 200 | 0.032 | 0.051 | 0.941 | 0.980 | | |
| | c | | 400 | 0.038 | 0.052 | 1.117 | 1.000 | | |
| | d | | 500 | 0.035 | 0.054 | 1.029 | 1.038 | | |
| 1. | e | +10.00 | 800 | 0.037 | 0.050 | 1.088 | 0.961 | 0.264 | 0.039 |
| | f | | 1000 | 0.034 | 0.052 | 1.000 | 1.000 | | |
| | g | | 1500 | 0.040 | 0.052 | 1.126 | 1.000 | | |
| | h | | 2000 | 0.042 | 0.053 | 1.235 | 1.019 | | |
| | i | | 2500 | 0.043 | 0.051 | 1.264 | 0.980 | | |
| | a | | 100 | 0.025 | 0.025 | 0.781 | 0.925 | | |
| | b | | 200 | 0.025 | 0.024 | 0.781 | 0.888 | | |
| | c | | 400 | 0.031 | 0.025 | 0.968 | 0.925 | | |
| | d | | 500 | 0.033 | 0.026 | 1.031 | 0.962 | | |
| 2. | e | +30.00 | 800 | 0.034 | 0.027 | 1.062 | 1.000 | 0.343 | 0.112 |
| | f | | 1000 | 0.032 | 0.027 | 1.000 | 1.000 | | |
| | g | | 1500 | 0.036 | 0.029 | 1.125 | 1.074 | | |
| | h | | 2000 | 0.039 | 0.028 | 1.218 | 1.037 | | |
| | i | | 2500 | 0.043 | 0.030 | 1.343 | 1.111 | | |
| | a | | 100 | 0.031 | 0.051 | 0.885 | 0.980 | | |
| | b | | 200 | 0.032 | 0.052 | 0.914 | 1.000 | | |
| | c | | 400 | 0.036 | 0.051 | 1.028 | 0.980 | | |
| | d | | 500 | 0.037 | 0.053 | 1.057 | 1.019 | | |
| 3. | e | -10.00 | 800 | 0.037 | 0.052 | 1.057 | 1.000 | 0.200 | 0.038 |
| | f | | 1000 | 0.036 | 0.052 | 1.00 | 1.000 | | |
| | g | | 1500 | 0.038 | 0.051 | 1.085 | 0.980 | | |
| | h | | 2000 | 0.040 | 0.051 | 1.142 | 0.980 | | |
| | i | | 2500 | 0.042 | 0.054 | 1.200 | 1.038 | | |
| | a | | 100 | 0.027 | 0.025 | 0.818 | 0.862 | | |
| | b | | 200 | 0.025 | 0.023 | 0.757 | 0.793 | | |
| | c | | 400 | 0.029 | 0.026 | 0.878 | 0.896 | | |
| | d | | 500 | 0.031 | 0.024 | 0.939 | 0.827 | | |
| 4. | e | -30.00 | 800 | 0.032 | 0.026 | 0.969 | 0.896 | 0.363 | 0.103 |
| | f | | 1000 | 0.033 | 0.029 | 1.000 | 1.000 | | |
| | g | | 1500 | 0.036 | 0.030 | 1.090 | 1.034 | | |
| | h | | 2000 | 0.037 | 0.031 | 1.121 | 1.068 | | |
| | i | | 2500 | 0.045 | 0.032 | 1.363 | 1.103 | | |

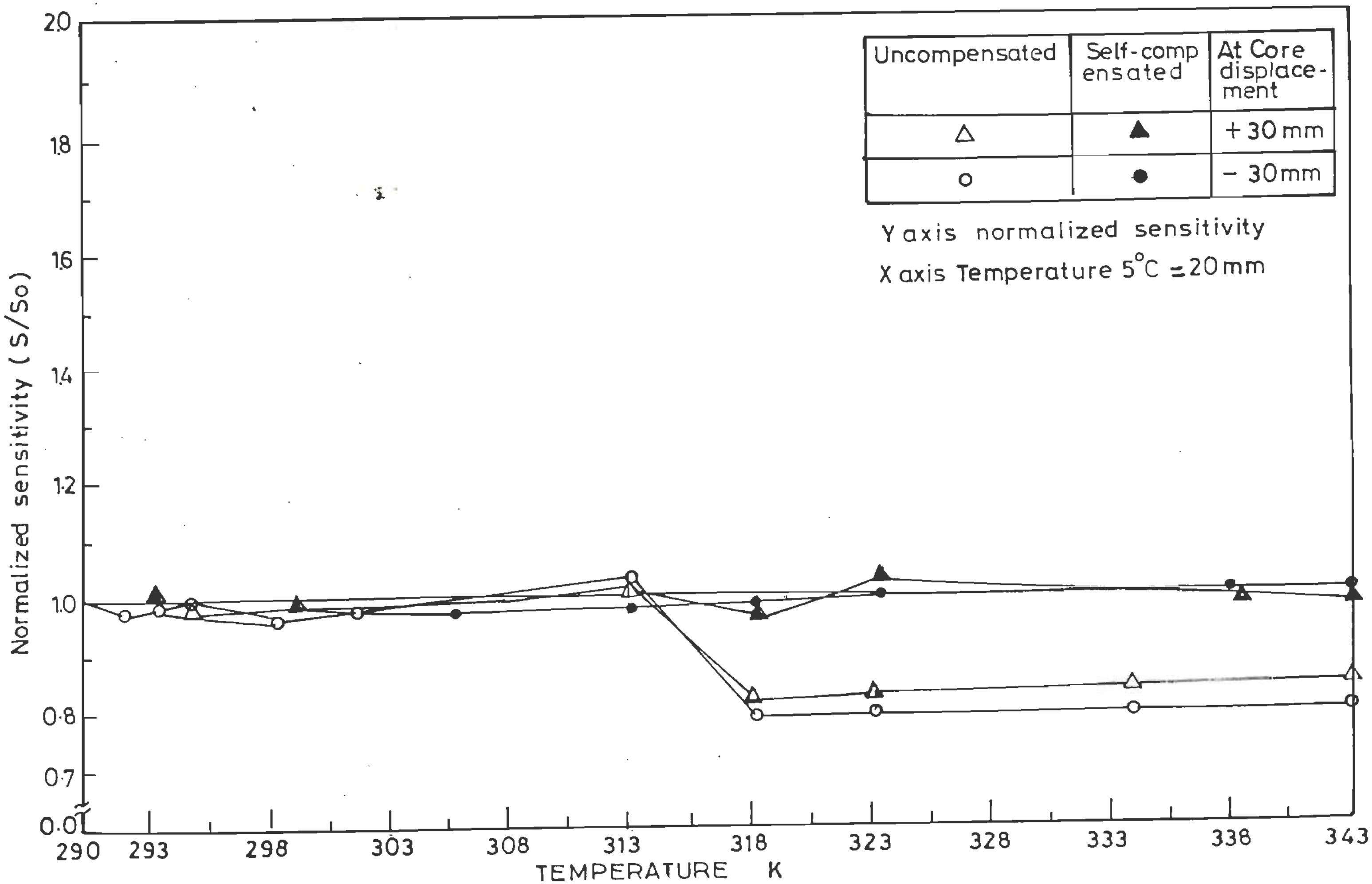



FIG. 2.13 NORM. SENSITIVITY CHANGES IN ORDINARY DIFFERENTIAL AND SELF-COMPENSATED LVDT OUTPUTS AT 5V AND 1KHz DUE TO VARIATION IN AMBIENT TEMPERATURE.
(THE SENSITIVITY AT ROOM TEMPERATURE IS TAKEN AS REFERENCE)

core positions of 30 mm in both directions w.r.t. null (central) position of the core travel. The variation of $(S/S_o)$ for uncompensated LVDT is within -0.19 to 0.01, whereas of compensated is only within -0.04 to 0.01 w.r.t. the normalized axis for a reference temperature of 290 K for the extrement core position of +30 mm. The variation of $S/S_o$ for uncompensated type is in the range of -0.21 to 0.02 while for self-compensated, is only within -0.02 to 0.006 w.r.t. normalized axis for the extreme core position of -30 mm. The reference ambient temperature is choosen as 290 K. Thus, the influence of the variation in ambient temperature upon the performance of self-compensated type is negligible.

It is evident from theoretical relationships and duly supported experimental results that compensated dual set secondary winding LVDT transducer has better and improved performance compared to uncompensated LVDT. As such, self-compensated LVDT is highly immune to variations in excitation voltage, frequency and changes in environmental temperature. The sensitivity of compensated type for a change in measured (linear core displacement) is high. It has good linearity of input-output characteristic in normal operating span of ±30 mm. The dual set winding compensation technique introduces a marginal increase in coil dimensions, weight, and cost of transducer assembly. This technique makes the transducer 'smart' enough to provide self-compensation without making use of additional temperature compensation element (e.g., NTC-thermistor) and inhousing of another (reference) transducer with the measuring transducer [91]. The data acquisition and handling from the compensated LVDT do not put any (additional) constraint on the signal processing and conditioning circuits, or on transducer assembly and also its placement even in adverse conditions. This approach is a novel in design and original in concept for self-compensated smart LVDT transducers.

This approach as such has resulted in the development of a smart self-compensated LVDT with minimal modifications and additions in the over all transducer assembly. Due to high performance stability w.r.t. variations in ambient temperature, the transducer is suitable for all indoor and outdoor applications including hostile environment. The obtained performance characteristics show that the compensated LVDT offers more stable operation than uncompensated under identical variations in test conditions i.e., input influencing parameters and changes in environmental temperature in and around transducer assembly.

## 2.3 RATIO AND DIFFERENTIAL INDUCTIVE TRANSDUCERS

The performance of inductive ratio transducer and inductive differential transducer depend, by and large, upon their dimensional features, material of core, excitation conditions of voltage and frequency, number of turns in the windings and environmental and winding temperatures in and around the transducer assembly [46]. In the present work, efforts have been made, to make the inductive transducers immune to variations in excitation conditions and environmental and winding temperatures. As such, two linear variable inductive transducers, using ratio and difference of inductances offered by the two identical coils of transducer assembly have been analyzed and developed for precise measurement of displacement in range of ±15 mm.

### 2.3.1 Basic Principle of Ratio and Differential Transducers

By inductive displacement transducers, linear displacement to be measured as input is made to vary the inductances of coils. The inductances of coil depend on the length of that part of the core which has penetrated into the coil. The varying inductance changes the parameters of an electronic circuit which is connected for processing and conditioning of signals.

The principle of transduction of transducer is shown in Fig.2.14(a). Two inductive coil windings ($C_1$,$C_2$) are connected in output circuits of constant current sources, $A_1$ and $A_2$. The coils are perfectly identical in geometry, dimensions, and also in their inductance and resistance values. Both the coils exhibit similar values of inductance (L) and effective resistance (r) for identical conditions of the core. With these instrumentation systems, the coils are arranged in push-pull combination and the input magnitude (as a result of ferromagnetic core displacement), simultaneously causes an

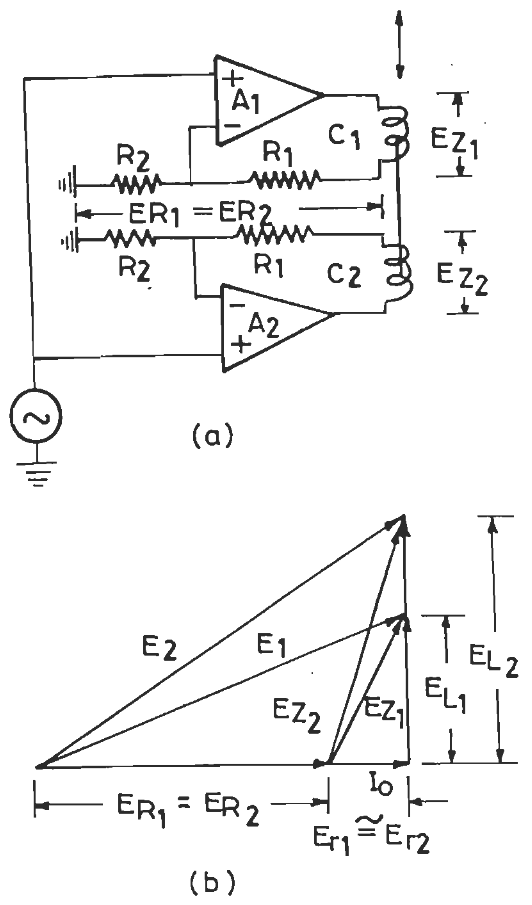



FIG. 2.14 . PRINCIPLE OF TRANSDUCTION
(a) BASIC CIRCUIT (b) VECTOR DIAGRAM

increase of one inductance and a decrease of another. For two different configuration of the coils (in ratio or differential mode), the subsequent stage measures either the ratio or difference of two inductance values.

2.3.1.1 Case I: Inductive Ratio Transducer

The coils are arranged in ratio mode. Fig.2.14(b) shows the vector diagram of voltage drops across the coils ($C_1$,$C_2$) and the fixed value resistors ($R_1$ and $R_2$). These resistors are connected in output circuits of current sources. The sources supply constant current, ($I_o$) at constant frequency ($\omega$). $E_{R1}$ and $E_{R2}$ are constant voltage drops across $R_1$ and $R_2$. From the vector diagram:

$E_{L1}$ is the reactive component of voltage drop across $C_1$,
$E_{L2}$ is the reactive component of voltage drop across $C_2$,
$Er_1$ is the resistive component of voltage drop across $C_1$,
$Er_2$ is the resistive component of voltage drop across $C_2$,
$Ez_1$ is the resultant voltage drop across $C_1$,
$Ez_2$ is the resultant voltage drop across $C_2$,
$E_{R1}$ is the fixed voltage drop across $R_1$, and
$E_{R2}$ is the fixed voltage drop across $R_2$.

Also, $E_{L1} >> Er_1$

$E_{L2} >> Er_2$

and $Er_1 = Er_2$

From the figure,

$$Ez_1 = Er_1 + E_{L1} \text{ and } E_{L1} \doteq I_o\omega L_1 \qquad \ldots (2.16)$$

$$Ez_2 = Er_2 + E_{L2} \text{ and } E_{L2} = I_o\omega L_2 \qquad \ldots (2.17)$$

The output voltage, $E_{oir} = \frac{E_{z1}}{E_{z2}} = \frac{I_o\omega L_1}{I_o\omega L_2} = \frac{L_1}{L_2}$ ... (2.18)

For linear displacement of core in normal operating range, the output voltage is the ratio of two inductances $L_1$ and $L_2$. When the core is exactly held stationary in the centre of coil windings, the output is unity. The excitation parameters $I_o$ and $\omega$ do not appear in the final output expression given by equation (2.18). Therefore, there is no influence of the excitation parameters on the output response of transducer. Also the influence of temperature on transducer response is negligible.

### 2.3.1.2 Transducer Assembly and Instrumentation System

The cross-sectional view and constructional details of inductive ratio transducer are shown in Figs.2.15 and 2.16 respectively. The coaxial coils are wound uniformly close to each other on a nonmagnetic and insulating former made of paper conduit with equal number of turns of 36 S.W.G., S.E.C. wire. The winding assembly is within a cylindrical case and the connecting leads are brought out from the assembly. The core is made of ferromagnetic material so that it acts as a magnetic shield against strong magnetic fields. Guide rings are provided

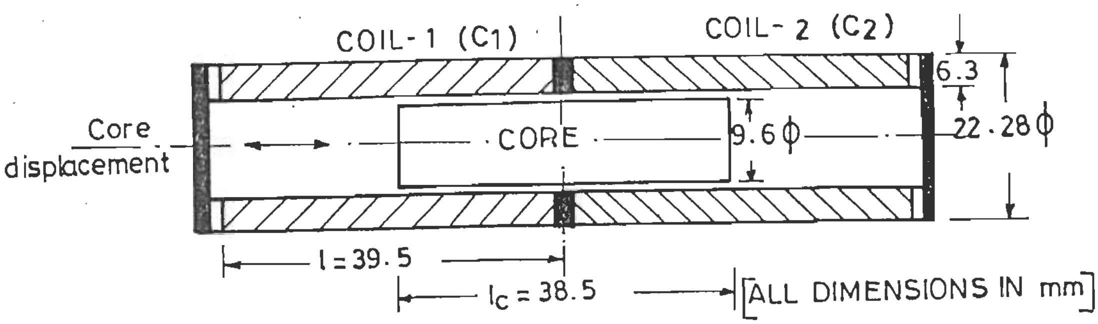



FIG. 2.15 VARIABLE INDUCTANCE TRANSDUCER
[$L_1$, $L_2$ = 40mH with air and 230 mH with ferrite-core, Number of turns of 36 S.W.G, S.E.C conductor in each coil = 3500]

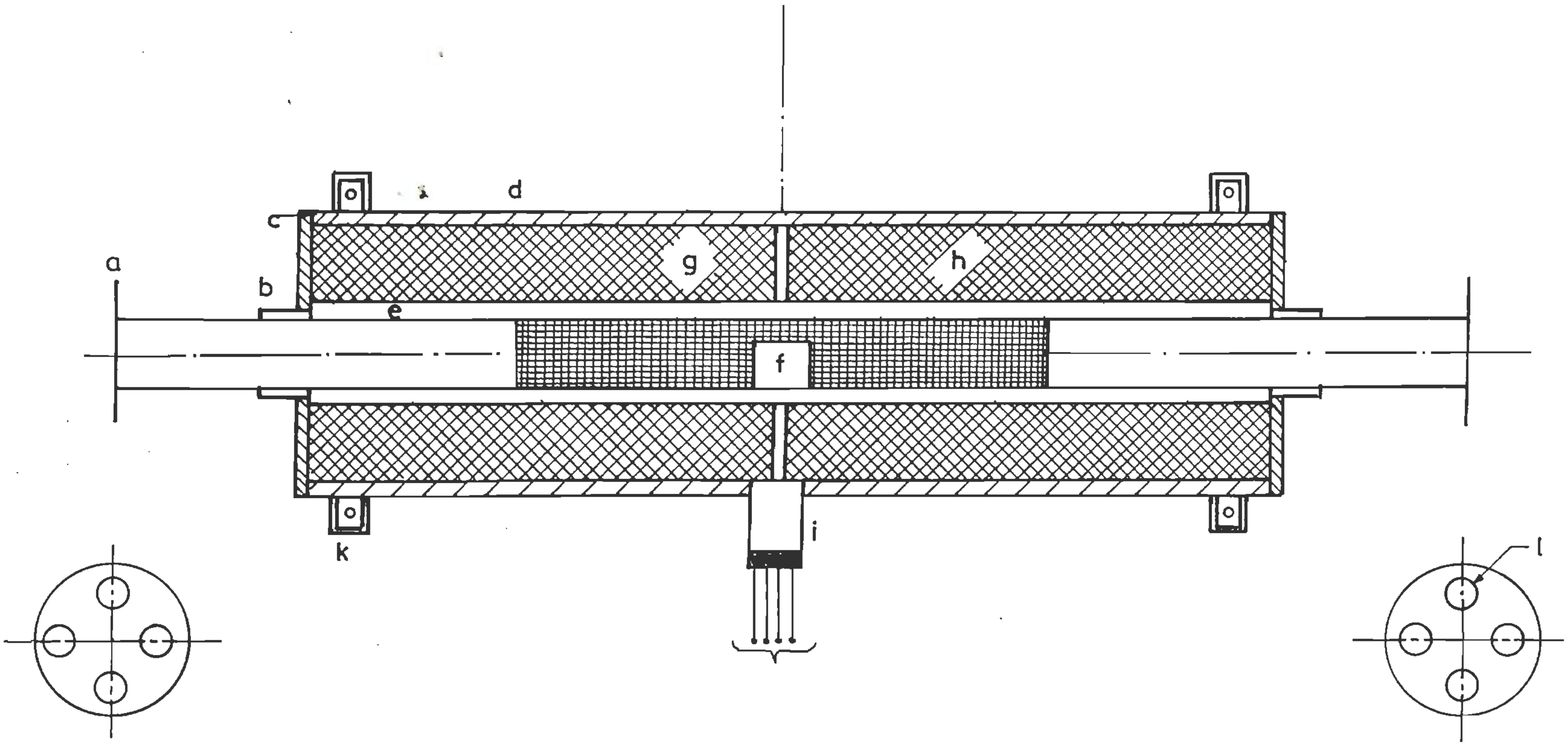



FIG. 2.16 INDUCTIVE RATIO TRANSDUCER CONSTRUCTION

(a) FLANGE FOR COUPLING ; (b) GUIDE SLEEVE ; (c) END COVER ; (d) OUTER COVER ; (e) NON - MAGNETIC METALLIC EXTENSION ; (f) FERROMAGNETIC CORE ; (g) COIL - 1 ; (h) COIL - 2 ; (i) HOLES TO TAKE OUT WIRES ; (j) WIRES ; (k) FOR FIXING THE TRANSDUCER ; (l) HOLES FOR COUPLING in the end covers on both extremes of casing for friction-free and axial movement of the core within the bobbin. The length of the armature (core) is extended on both sides with non-magnetic pieces of similar diamter to facilitate the movement of core inside the bobbin. This extension is used here as sensing shaft.

The value of inductance in a single layer and uniformly wound coil is given by the relationship,

$$L = \frac{4\pi^2 \; n^2 \; r^2 10^{-7}}{\ell} \; (H) \qquad \ldots (2.19)$$

where,

n is the length of turns in coil,

$\ell$ is the length of coil (m), and

r is the radius of coil (m).

If a ferromagnetic core of the same length as the coil and radius $r_c$ is introduced, the inductance will increase with the increased total flux to a value given by the expression

$$\frac{4\pi^2 n^2 [r^2 + (\mu_m - 1) r_c^2]}{\ell} \; 10^{-7} \; (H) \qquad \ldots (2.20)$$

where $\mu_m$ is the 'effective' permeability of the ferromagnetic core in the push-pull arrangement. With a core length $\ell_c$ smaller than coil length $\ell$, the inductance becomes

$$L = \left[\frac{4\pi^2 n^2}{\ell_c} \cdot \left(\frac{\ell_c}{\ell}\right)^2 \{ r^2 + (\mu_m - 1) r_c^2 \} + \frac{4\pi^2 n^2}{\ell - \ell_c} \cdot \left(\frac{\ell - \ell_c}{\ell}\right)^2 r^2 \right] . 10^{-7}$$

$$= 4\pi^2 n^2 \{\ell r^2 + (\mu_m - 1) \ell_c r_c^2\} / 10^7 \ell^2 \; (H) \qquad \ldots (2.21)$$

It is seen from equation (2.21) that for an increase in $\ell_c$ by $\delta\ell_c$, i.e., if the core is pushed farther into coil $C_1$ by an amount of $\delta\ell_c$, the inductance (L) increases by ($\delta$L). The new value of inductance is

$$(L+\delta L) = 4\pi^2 n^2 \{\ell r^2+(\mu_m-1)r_c^2(\ell+\delta\ell_c)\}/10^7.\ell^2 (H) \qquad \ldots (2.22)$$

By substracting equation (2.21) from (2.22), the change in inductance is obtained as given below

$$\delta L = 4\pi^2 n^2\ r_c^2(\mu_m-1)\delta\ell_c/10^7.\ell^2 (H) \qquad \ldots (2.23)$$

The fractional change of inductance, representing sensitivity of transducer (of this type), is given by the expression as under

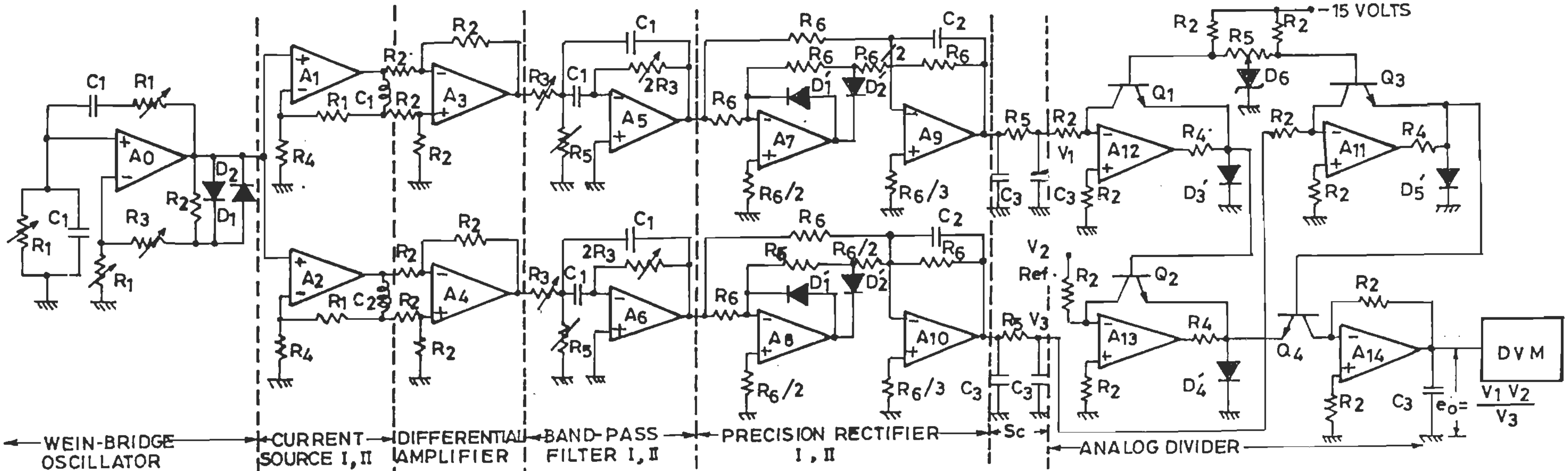
$$\frac{\delta L}{L} = \frac{\delta\ell_c}{\ell_c}.\ \frac{1}{1+(\frac{\ell}{\ell_c}).(\frac{r}{r_c})^2\{\frac{1}{(\mu_m-1)}\}} \qquad \ldots (2.24)$$

In a similar way, coil $C_2$ experiences an identical change in inductance of opposite sign.

It is clear from equation (2.24) that the fractional change of inductance ($\delta$L/L) is proportional to the fractional change in core length ($\delta\ell_c/\ell_c$) (i.e., core displacement) multiplied by a factor smaller than unity. To make the factor as large as possible, i.e., to obtain maximum sensitivity, the ratios $\ell/\ell_c$ and $r/r_c$ should approach unity and effective permeability $\mu_m$ of the core should be as large as possible.

The identical coils ($C_1$ and $C_2$) are designed by making use of relationships given by equations (2.19) and (2.20). The inductance value of each coil with-air as medium is 40 mH and with ferrite-core 230 mH. These values have been verified by the actual measurements on an inductance bridge. The ratios of $\ell/\ell_c$ and $r/r_c$ are very cloase to unity and a very high value of effective permeability of the core, approximately equal to 10,000 $Hm^{-1}$, is chosen for the design to meet the condition of maximum sensitivity of inductive ratio transducer.

The schematic details of instrumentation scheme are shown in Fig.2.17. The output of Wein-Bridge oscillator is fed to constant current sources $A_1$ and $A_2$. The output voltage and frequency of the oscillator is varied from 0.5 to 5.0 volt and 2.33 kHz to 8.33 kHz, respectively. The output current of the sources is 353 $\mu$Amp. The output voltages of the coils after processing through differential amplifiers, band-pass filters, and precision rectifiers, is fed to monolithic IC analog divider unit (ref. section 2.2.2 for analog divider unit). The output from the divider unit is finally given to a digial voltmeter (DVM) which is calibrated directly in terms of linear displacement. High quality narrow band-pass filter is designed at resonant frequency ($f_r$), of 5 kHz for the quality factor (Q) to be equal to 16 and bandwidth (B) equals to be 2000 rad/sec ($<0.1\ \omega r$). The closed-loop gain of filter is taken as unity. Fig.2.18 shows the frequency response of the filter. The filter circuit components Rs and Cs are choosen of high quality with very low tolerance limit to achieve




[$R_1$ = 4.7K, $R_2$ = 10K, $R_3$ = 50K, $R_4$ = 1K, $R_5$ = 120Ω, $R_6$ = 22K, $C_1$ = 0.01μF, $C_2$ = 10uF, $C_3$ = 50uF, ($D_1$-$D_2$) = 1N914, $D_6$ = LM103 2.4V, ($Q_1$-$Q_4$) = 2N3728 matched pairs, ($A_1$-$A_{14}$) = LM324, ($C_1$-$C_2$) = Inductive transducer coils, SC = Smoothing circuits, ($D_1'$-$D_5'$) = IN4148]

FIG.2.17 CIRCUIT DETAILS OF INSTRUMENTATION SYSTEM

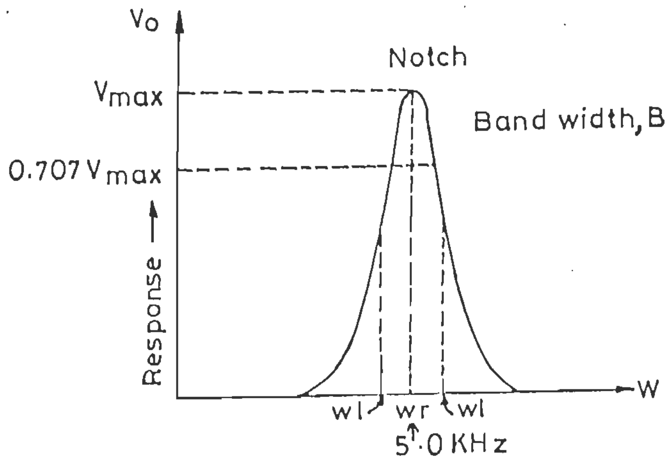



Figure 2.18 RESPONSE OF BAND-PASS FILTER

sharp response characteristic. The details of filter circuit are shown in Fig.2.17. The resistance, $R_5$ is used as variable component to adjust the notch of filter response to match the frequency output of Wein-Bridge oscillator The filter circuit is disconnected from the rest of the instrumentation scheme, when the transducer is tested for its performance under the effect of variation of frequency from 2.32 to 8.33 kHz. The instrumentation scheme in entity is used for the measurement of linear displacement under constant input excitation voltage (5V) and frequency (5 kHz).

### 2.3.1.3 Experimental Results and Discussions

Fig. 2.19 , shows the output response w.r.t. linear displacement upto ±15 mm with a sensitivity of 0.10 volts/mm. The variation in normalized sensitivity is within +0.62 for a change in excitation voltage from 0.5 to 5.0 volts at 8.33 kHz w.r.t. reference voltage at 2.5 volts; within +0.08 for a change in excitation frequency from 2.32 to 8.33 kHz; and within ±0.01 for a change in ambient temperature from 298 to 373 K at 5 volt and 8.33 kHz w.r.t reference temperature at 313 K as shown in Fig.2.20,2.21,2.22, respectively. The variations in response characteristics are quite small. It shows that the transducer can be used for accurate and precise measurement without being much influenced by the considerable changes in excitation conditions and environmental temperatures in and around the transducer assembly.

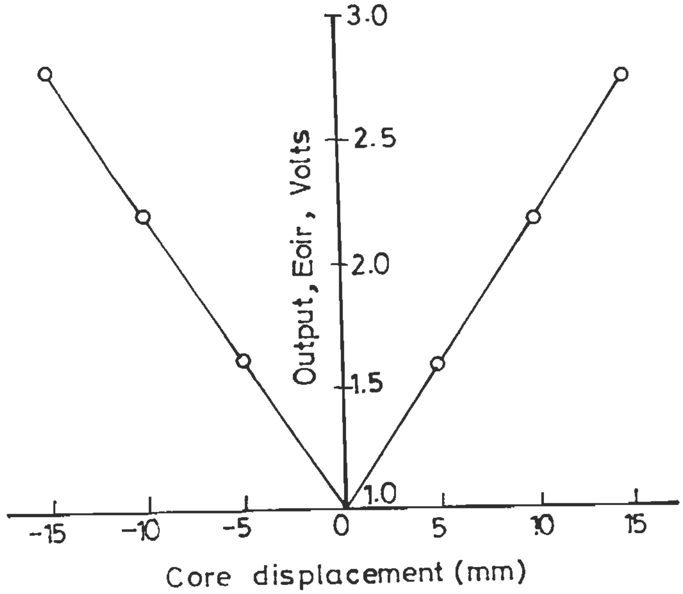



FIG. 2.19 CORE DISPLACEMENT VS OUTPUT RESPONSE FOR IR TRANSDUCER AT 4.0 VOLT, 8.33 kHz

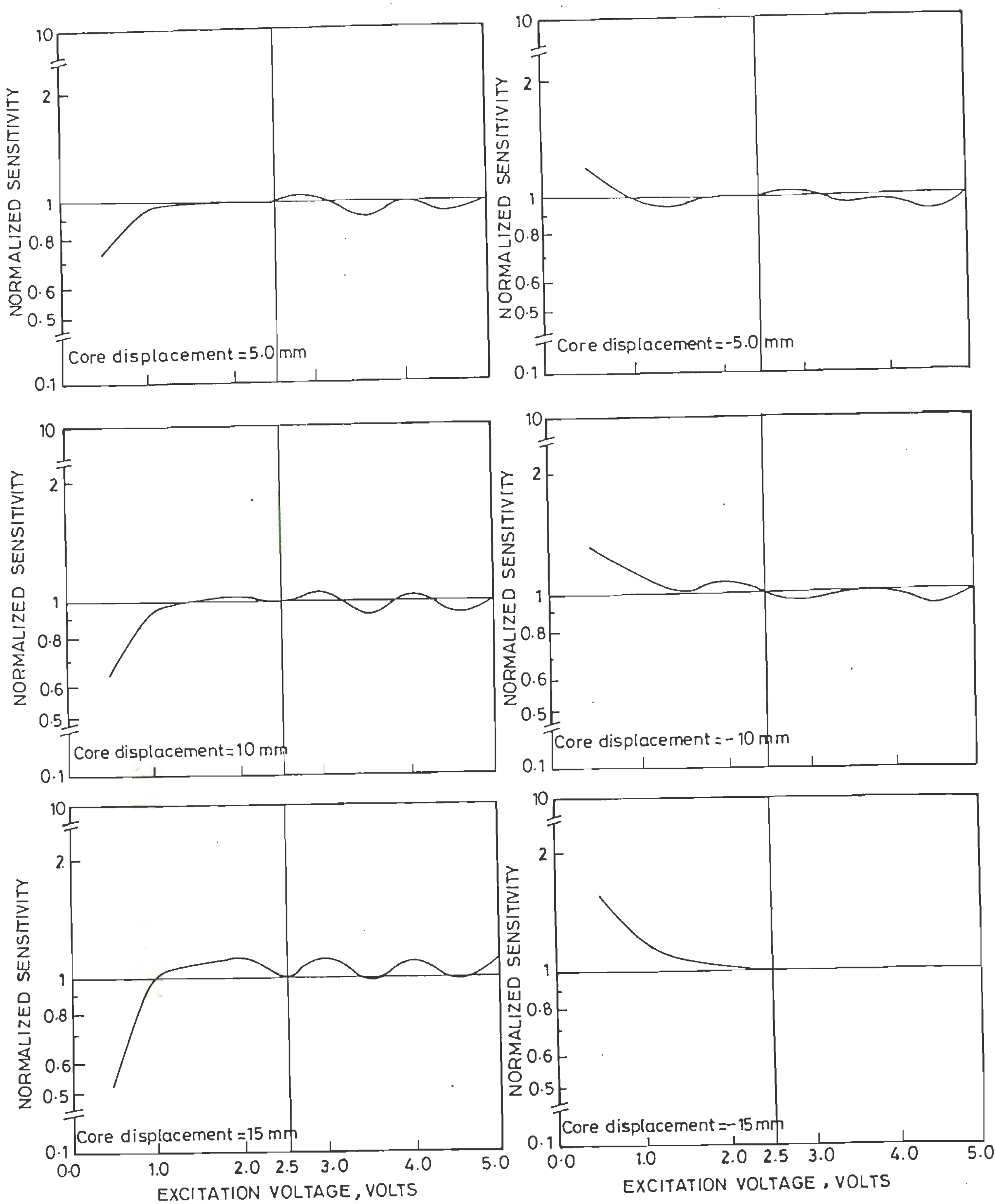



FIG. 2.20 NORM. SENSITIVITY CHANGES IN IR TRANSDUCER DUE TO VARIATION IN EXCITATION VOLTAGE AT 8.33 kHz AT ROOM TEMPERATURE (SENSITIVITY AT 2.5 VOLTS IS TAKEN AS REFERENCE)

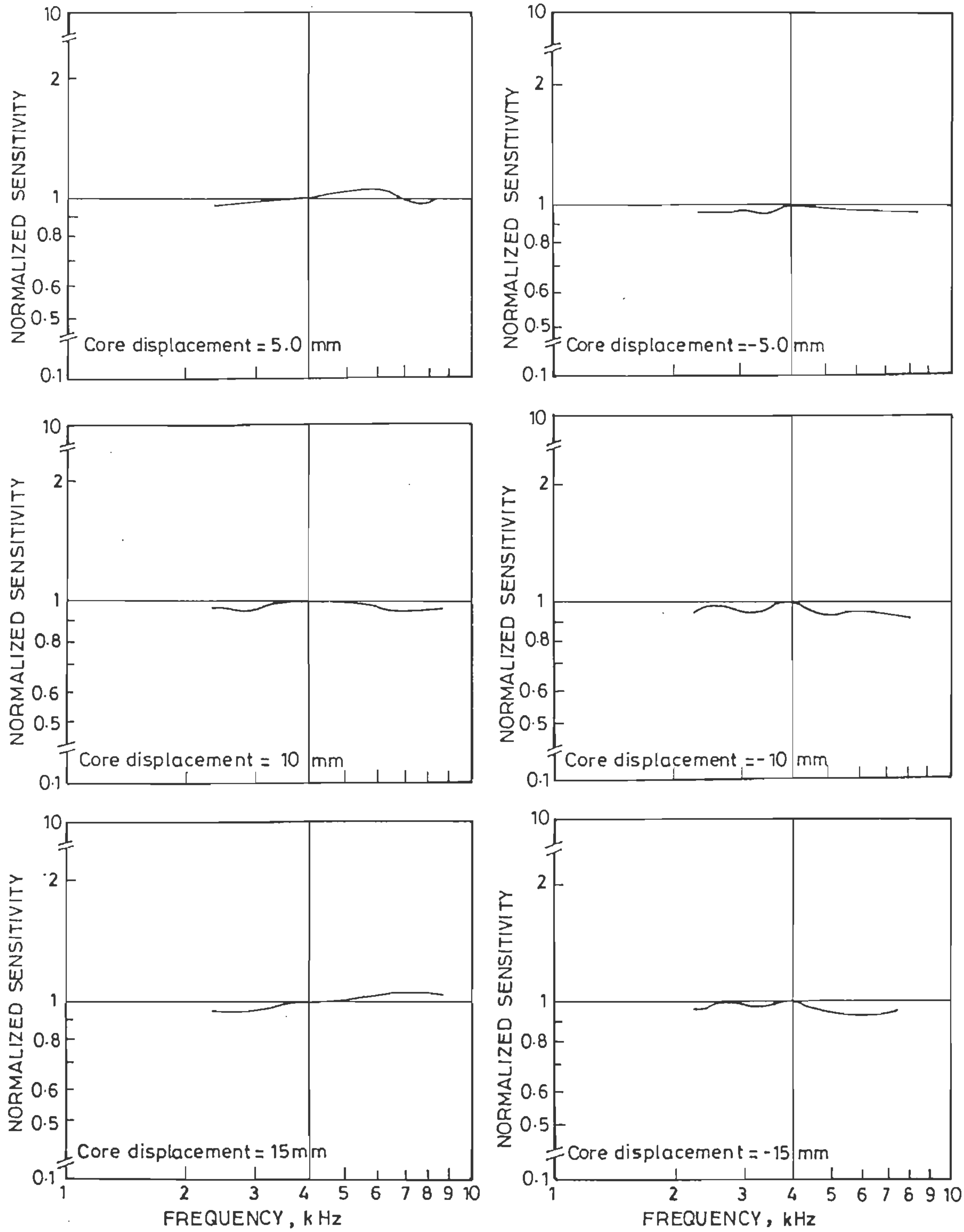



FIG. 2.21 NORM. SENSITIVITY CHANGES IN IR TRANSDUCER DUE TO VARIATION IN EXCITATION FREQUENCY AT 2 VOLTS AT ROOM TEMPERATURE (SENSITIVITY AT 4 kHz IS TAKEN AS REFERENCE.)

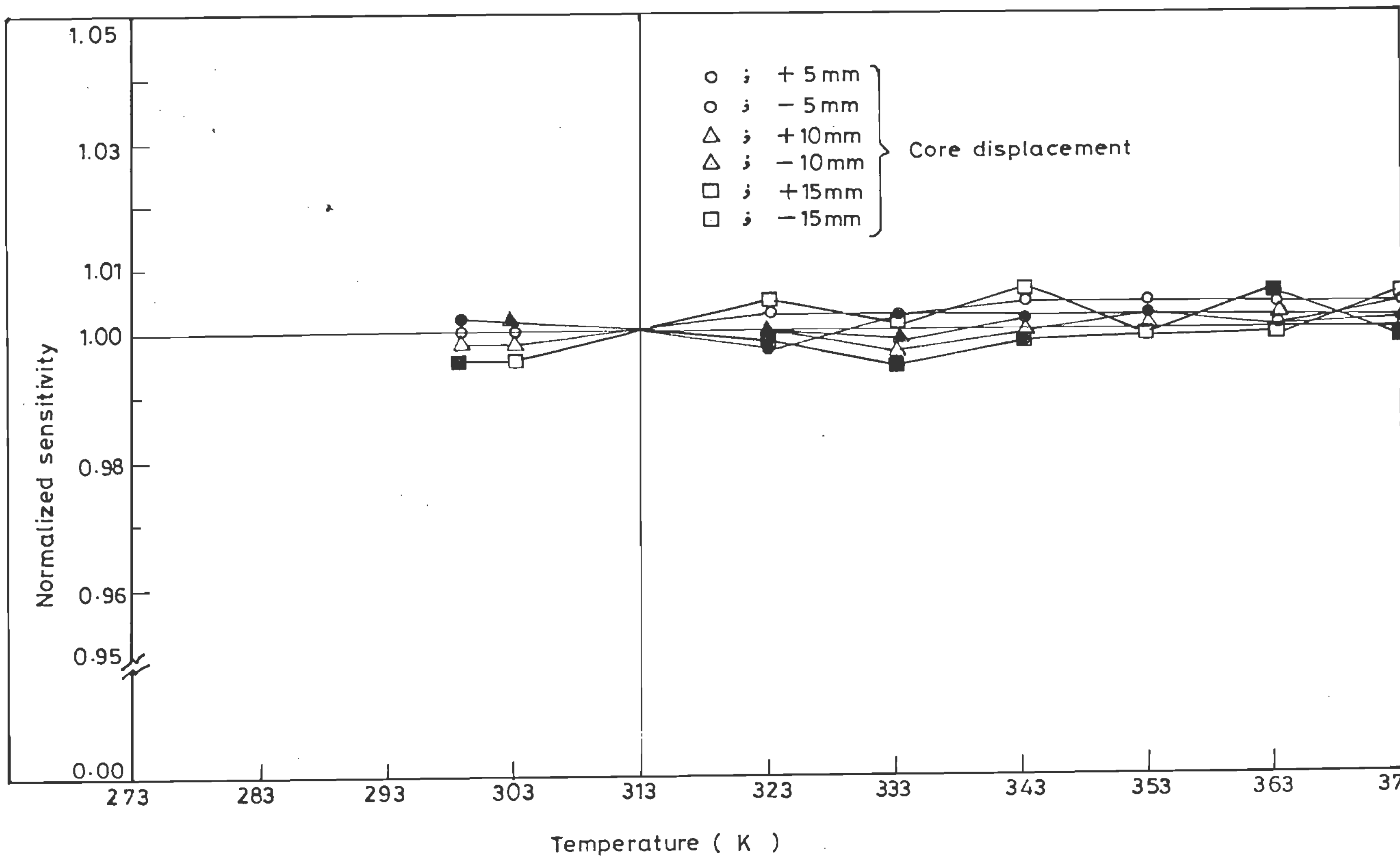



FIG.2.22 NORM. SENSITIVITY CHANGES IN IR TRANSDUCER AT 5V, 8.33 kHz DUE TO VARIATION IN AMBIENT TEMPERATURE (THE SENSITIVITY AT 313K IS TAKEN AS REFERENCE)

IRT based instrumentation system has been successfully developed and tested for precise linear displacement measurement. It has following advantages:

i) Although, the response is similar to that obtained by compensated LVDT transducer, it has distinct advantage over LVDT transducer. It has only two coil windings compared to three in uncompensated and five in compensated LVDT transducers. There is reduction in size, weight, and cost of the complete transducer assembly due to lesser number of windings.

ii) The transducer response is not influenced by variations in undesirable excitation parameters and environmental and windings temperatures.

iii) As the response of the instrumentation system is highly linear and stable, it can be used for accurate measurement of linear displacement for different applications in Industrial, Process, Aerospace, and Biomedical fields in all type of environments including hostile conditions in indoor and outdoor applications.

iv) Due to nonavailability of analog divider chip at the time of development of this instrumentation scheme, it has been tested using monolithic analog divider. The response of the system will further improve by the use of analog divider IC chips (AD533, AD534) in the output stage of instrumentation scheme.

v) This is suitable for all practical purposes, as it is smaller in size, light in weight and does not come in contact with the measurand directly.

vi) Thus, in its present modified form, it is an improved transducer in the family of the inductive transducers for linear displacement measurements.

2.3.2.1 Case II: Inductive Differential Transducer

The identical coils ($C_1$,$C_2$) are connected in differential mode in this system. Figs.2.23(a) and (b) show the arrangements of coils in this mode and vector diagram for different voltage drop components, respectively. Again coils are connected in the output circuits of the constant current sources, $A_1$ and $A_2$. Both the coils have same values of inductances (L) and effective resistance (r) for identical condition of the core. Vector diagram shows the drop across fixed components ($R_1$ and $R_2$) and other elements of the circuit. $E_R$ is voltage drop across $R_1$ and $R_2$ and remains same all throughout the experimentation due to constant current ($I_o$) and constant frequency ($\omega$) flowing through the elements. $Ez_1$ and $Ez_2$ are voltage drops across the coils $C_1$ and $C_2$. $Ez_1$ and $Ez_2$ are resultant voltage drops of reactive components $E_{L1}$ and $E_{L2}$ and resistive components $Er_1$ and $Er_2$.

At reference (null) position, when the core is stationed in the middle of coil assembly, $Ez_1$=$Ez_2$, $E_{L1}$=$E_{L2}$, $Er_1$=$Er_2$ and output voltage $E_0$ equals to

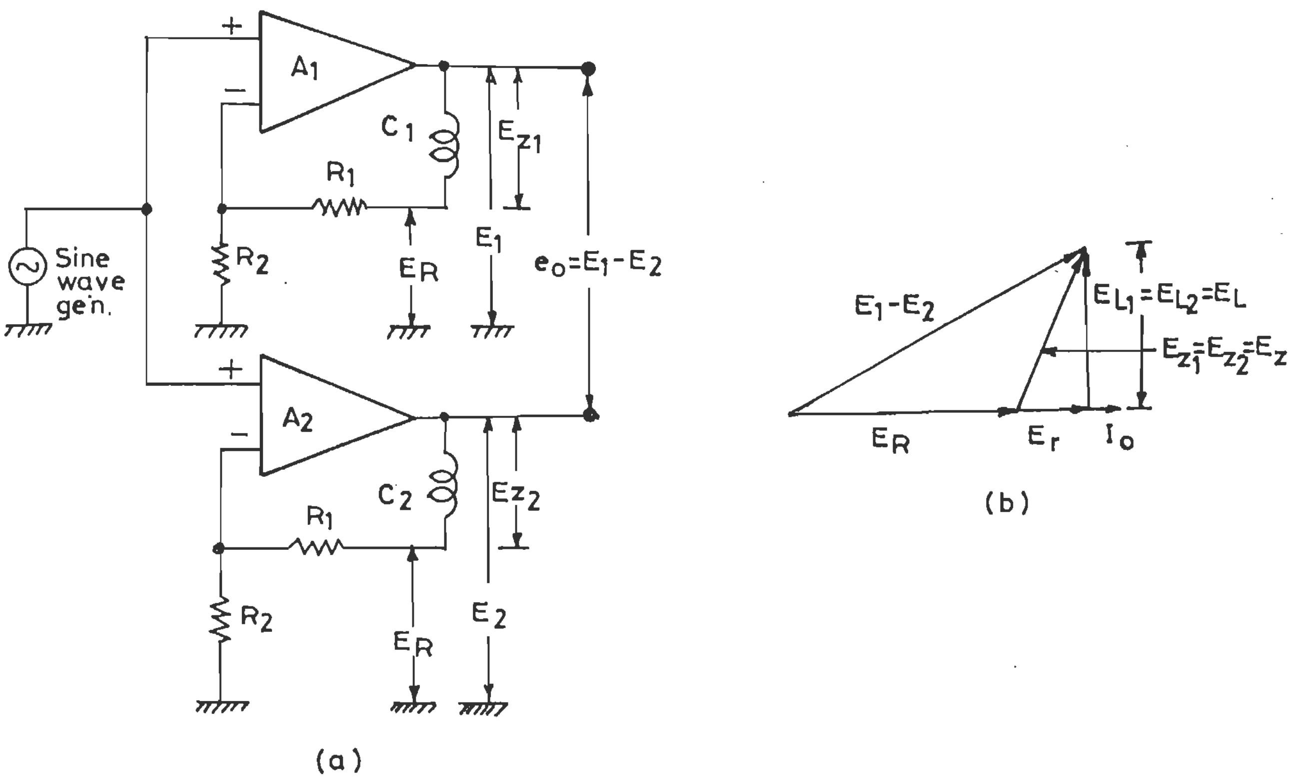



FIG. 2.23 PRINCIPLE OF TRANSDUCTION
(a) BASIC CIRCUIT (b) VECTOR DIAGRAM

$$E_0 = (Ez_1+E_R)-(Ez_2+E_R) = 0 \qquad \ldots (2.25)$$

When the core is pushed in positive (upward) direction in coil $C_1$, w.r.t. null position, there is a magnetic asymmetry in both the coils and $Ez_1$ becomes greater than $Ez_2$. The net output voltage equals to

$$E_0(+) = (Ez_1+E_R) - (Ez_2+E_R) = Ez_1-Ez_2 \qquad \ldots (2.26)$$

For the similar displacement of the core in coil $C_2$ in negative direction w.r.t. null position, $Ez_2$ becomes greater than $Ez_1$ and net output voltage equals to

$$E_0(-) = -(Ez_1-Ez_2) = Ez_2-Ez_1 \qquad \ldots (2.27)$$

For identical displacement of the core on either direction:

$$|Ez_1-Ez_2| = |Ez_2-Ez_1| \qquad \ldots (2.28)$$

For a coil of high quality (for Q > 100), the resistive component of the coil is negligible compared to its inductive component, [115] and



$$Ez = E_L = I_o \omega L \qquad \ldots (2.29)$$

For a particular level of excitation with fixed $I_o$ and $\omega$, $E_z$ is proportional to the value of inductance. For differential mode (of coils $C_1$,$C_2$), the final output voltage equals to

$$E_{old} = I_o \omega \, (L_1-L_2) \qquad \ldots (2.30)$$

This output voltage is proportional to differential change in inductance values for fixed values of $I_o$ and $\omega$. The variation in excitation conditions changes the output voltage without any change in inductance.

This type of instrumentation system is suitable for the measurement of linear displacement with constant value of excitation conditions. Even with these limitations, IDT is economical in use and cheaper in cost due to being developed only with two coil windings compared to uncompensated LVDT transducer which has three coil windings.

2.3.2.2 Transducer Assembly and Instrumentation System

The similar transducer assembly as discussed in case of the IRT, is employed for IDT, except that the coils are connected in differential mode.

Fig. 2.24, shows the basic block-diagram of the instrumentation system. It consists of Wein-Bridge oscillator, inducive differential transducers, constant current sources, differential amplifier, narrow band-pass filter, precision rectifier, smoothing circuit and digital volt meter (DVM).

Signal Processing Unit:

Fig. 2.25, shows the details of instrumentation scheme. The constant current source $A_1$ and $A_2$ are being fed at 5.0 volt, 8.33 kHz from Wein-Bridge oscillator. Larger value of $R_1$ and

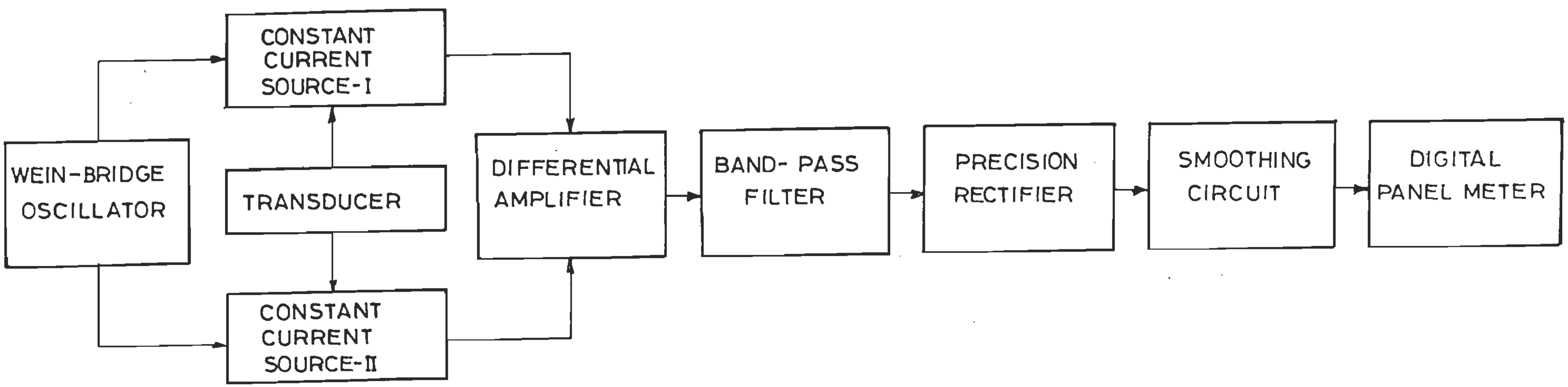
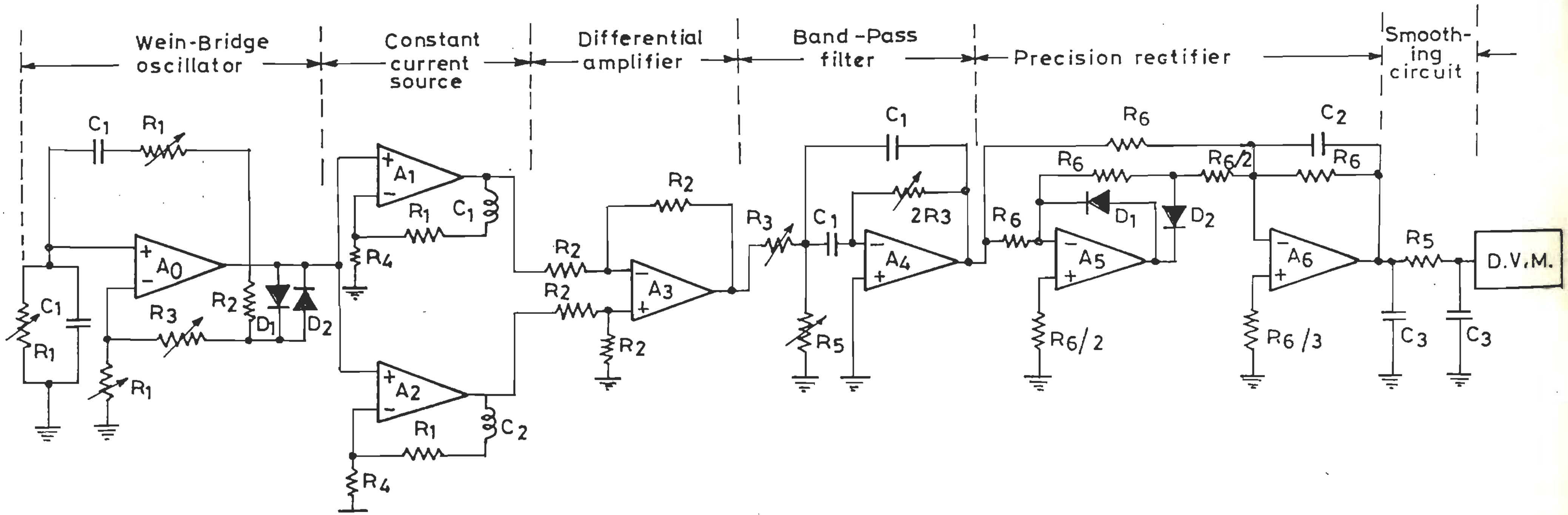



Figure 2.24 BASIC INSTRUMENTATION SYSTEM



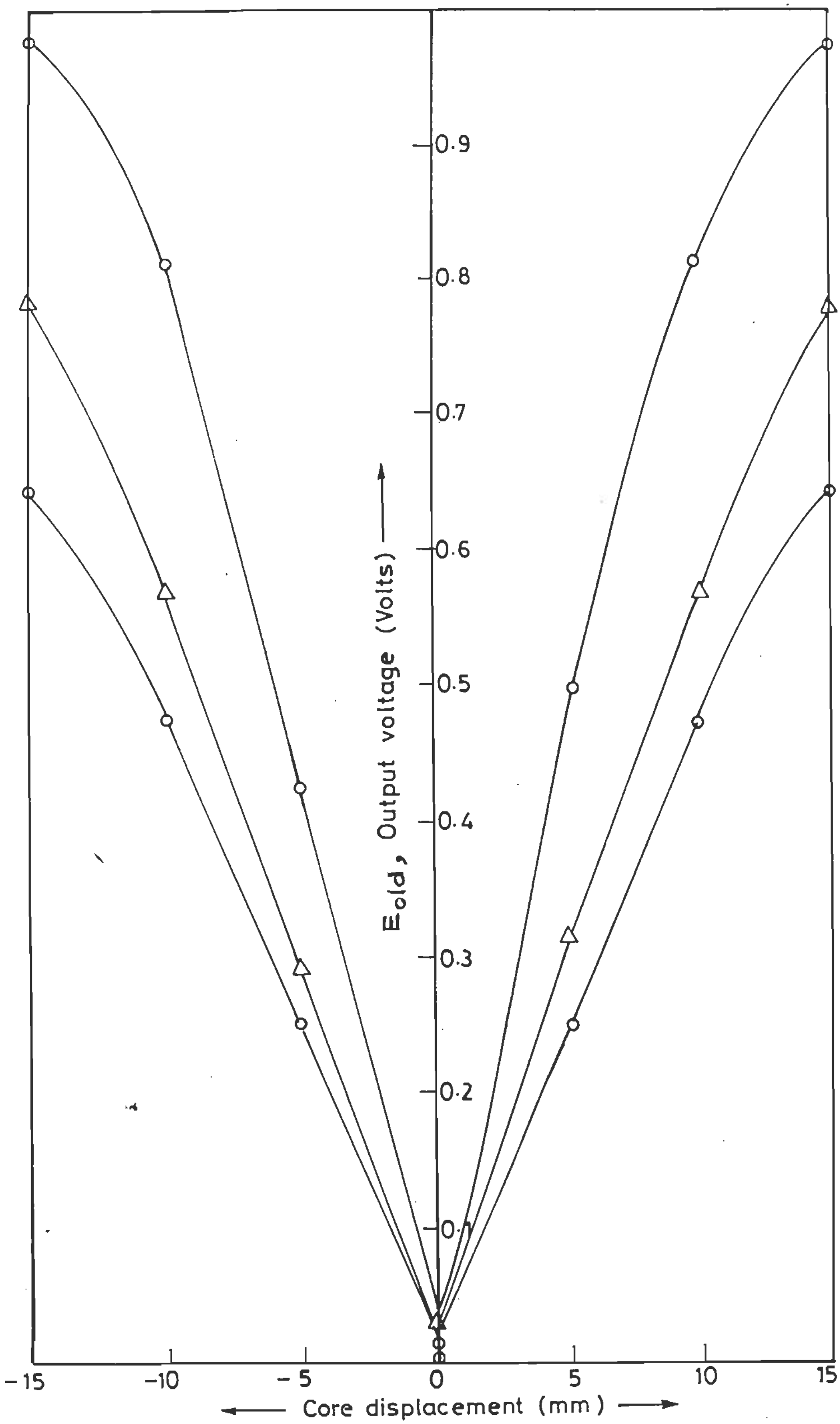

[ $R_1$=4.7K, $R_2$=10K, $R_3$=50K, $R_4$=1K, $R_5$=220 Ω, $R_6$= 22K, $C_1$= 0.01μ, $C_2$=10μ, $C_3$=50μ, ($D_1$-$D_2$) = 1N914, ($A_0$-$A_6$) = LM-324 ]

Figure 2.25 CIRCUIT DIAGRAM

$R_2$ compared to $Z_1$ and $Z_2$ helps in maintaining the constant value of $I_o$. Inductances $L_1$ and $L_2$ are exactly equal for matched coils $C_1$ and $C_2$. The output voltage drop across the transducer coils, which is proportional to $(L_1-L_2)$, is fed to differential amplifier. The output of previous stage is passed through the filter which has been designed for a resonant frequency of 8.33 kHz. It has unity (closed loop) gain with Q of 26 and bandwidth of 2000 rad/sec. Resistor $R_5$ is used to adjust the notch of frequency response of filter to match the output frequency of Wein-Bridge oscillator. The output of filter is fed to a precision rectifier and then subsequently to a smoothing circuit to get dc output proportional to $(L_1-L_2)$ in response to displacement of the core. This stage provides ripple free dc output and is measurable by digital volt meter which is calibrated directly in terms of linear displacement.

#### 2.3.2.3 Experimental Results and Discussions

The transducer has been tested and its performance has been evaluated for different types for variations in excitation level and temperature. Fig.2.26, shows the response curve of differential transducer at excitation voltages of 4,6, and 8 volts at 8.33 kHz for linear core displacement in the operating range of ±15 mm w.r.t. null position. The sensitivity is within ±0.08 volt/mm for 4 volts excitation; ±0.05 volts/mm for 6 volts excitation; and ±0.04 volt/mm for 8 volts excitation at a fixed value of excitation frequency




(Y-axis : 0.1V = 20 mm ; X-axis : 5 mm core disp = 20 mm ; Scale : Y-axis-1:1; X-axis-4:1; O-4V; 8.33KHz ; Δ- 6V; 8.33KHz ; O- 8V; 8.33KHz)

Figure 2.26 RESPONSE OF LVLD TRANSDUCER FOR DIFFERENT LEVELS OF EXCITATION VOLTAGES AT ROOM TEMPERATURE

(8.33 kHz). The response characteristics are perfectly linear upto about 80 per cent of the core travel. Further, the system has been tested for temperature variation from 298 to 373 K at 5.0 volts, 8.33 kHz. The normalized temperature sensitivity is within ±0.004 unit w.r.t. reference temperature at 313 K as shown in Fig.2.27. This IDT system exhibits no hystresis effect during the testing for number of repeated cycles for increasing and decreasing values of excitation voltage.

IDT with associated electronic instrument is suitable for measurement of linear displacement in response to displacement itself or any other physical parameter which produces displacement in response to change in its physical behaviour. It has high degree of accuracy, precision and stability with hystresis free output response. The interfering and influencing signals to both coils are eliminated due to differential mode of operation. The testing of system has proved beyond doubt the performance consistency of transducer.

## 2.4 CONCLUSIONS

All three transducer system have been tested successfully. The input-output response curves of the transducers are not direction sensitive, but can be made so, by using a phase sensitive detector circuit [87]. The detector consists of a analog multiplier unit which has two input; one is

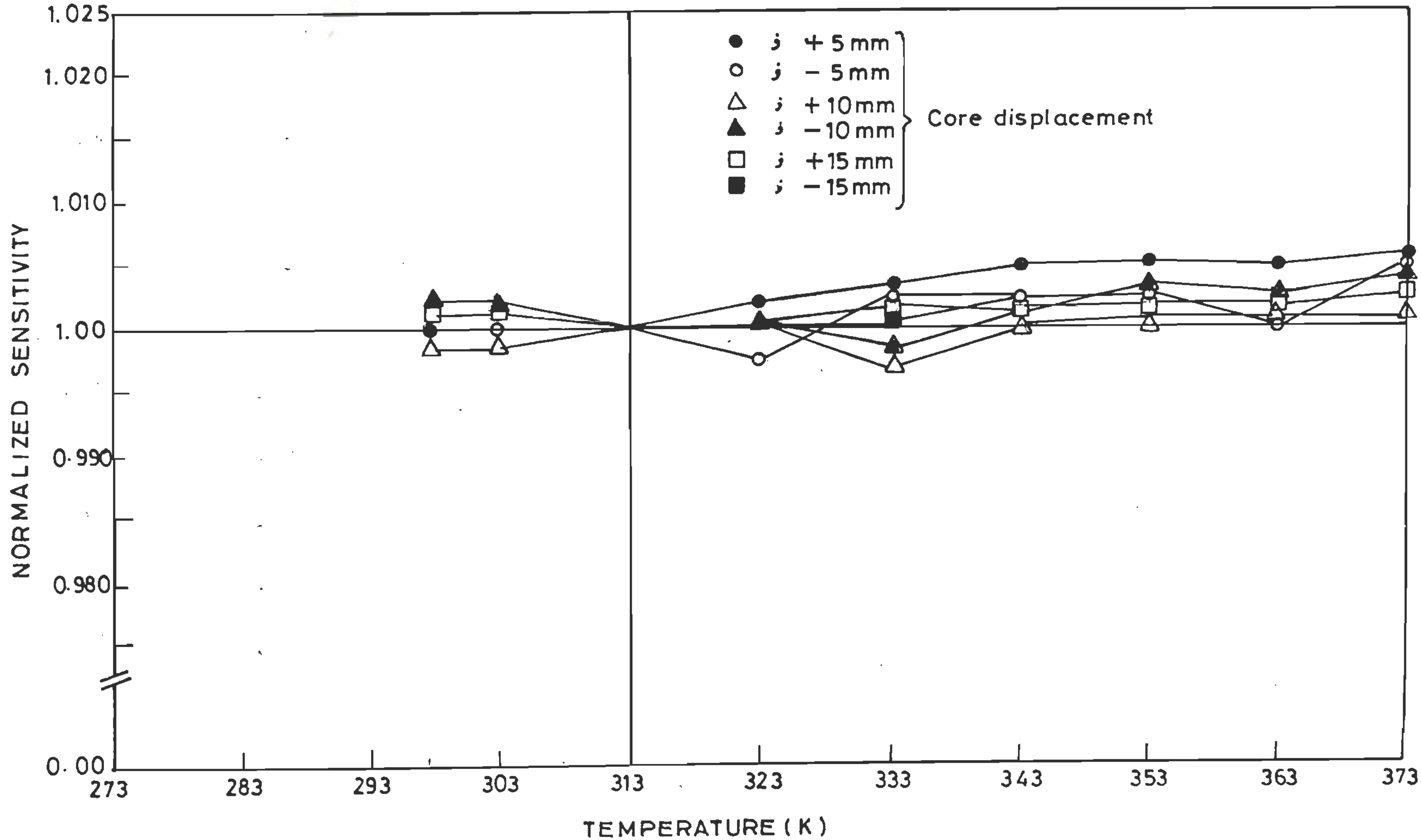



FIG. 2.27 NORM. SENSITIVITY CHANGES IN ID TRANSDUCER AT 5V, 8.33 kHz DUE TO VARIATION IN AMBIENT TEMPERATURE (THE SENSITIVITY AT 313K IS TAKEN AS REFERENCE)

the output of the transducer signal to be measured and the other is the reference sinusoid with fixed amplitude and with a phase identical to that of the ac excitation. The transducers output is available after amplification in one direction for position core displacement and in opposite direction for negative displacement. Thus the direction of the displacement is indirected by the polarity of the output voltage signal.

As such these transducers are new addition in the existing family of inductive transducers. These are stable in operation and show consistency in results in normal as well hostile environments with hystresis free response. Due to stable and consistency in responses, these are suitable for use in normal as well hostile environment for both indoor and outdoor applications.

# CHAPTER - III

# LINEARIZATION OF THERMISTOR RESPONSE USING ANALOG CIRCUITS

## 3.1 INTRODUCTION

From its development, thermistor has been in wide use for quick and precise measurement of temperature. Among the common temperature transducers, thirmistor has some advantages, like high temperature coefficient, small dimension, fast response, high sensitivity, easy matching with various electrical instruments, and ability to withstand electrical as well mechanical stresses. Typical commercially available thermistors provide usable outputs from -100 to +450°C. Some of these can be used upto 1000°C with limited accuracy, stability, and sensitivity. These are manufactured in resistance values ranging from tens of ohms to megohms at 25°C (room temperature) with a high manufacturing tolerance in the range of 5 to 25 per cent. Thermistors are semiconductor materials whose resistance changes with temperature. Thermistors are basically of two types due to difference in their constituent materials- Negative temperature coefficient (NTC) and Positive temperature coefficient (PTC) type. Out of these, NTC thermistors are extensively used for the measurement and control applications as they exhibit large variation of resistance over wide temperature range. Semiconductors usually have a high negative temperature coefficients of resistance. As the temperature is increased from 0 to 300°C, thermistor resistance decreases by a factor of thousand. The most common type of NTC thermistors are manufactured by mixing oxides of the iron group of transition metal elements, like chromium, maganese, iron, cobalt, and nickel. The resistance of NTC thermistor decreases exponentially with increasing temperature.

## 3.2 FACTORS AFFECTING LINEARITY

The resistance of NTC thermistor decays exponentially with its temperature (like discharging of a capacitor). The relationship of resistance against temperature is expressed mathematically as

$$R_{(T_1)} = R_{(To)} \exp b\left(\frac{1}{T_1} - \frac{1}{To}\right) \qquad \ldots (3.1)$$

where, $R_{(T_1)}$ is the cold resistance at measured absolute temperature $T_1$,

$R_{(To)}$ is the known cold resistance at known absolute temperature To, usually stated at ice point or at 25°C, and

b is the constant of thermistor material in K.

Typical values of b are between 3000K and 4500K and determined from resistance measurements at the ice point and at 50°C. The cold resistance values from various types of thermistors range from 500 ohms to over 10 megohms at 25°C. Thermistors are made in the form of beads, rods, disks, wafers, and flakes by different manufacturing processes. Time constant for a common bead unit is small and differ from medium to medium.

For example, in still air it is about 1 to 2 sec. while in agitated water (3 ft/sec.) it is less than 50 msec. Dissipation constant, which is the power required to raise thermistor temperature 1°C due to self-heating, lies between 0.1 and 2mW. Neither the resistance nor the current through the thermistor placed across a battery varies linearly over a wide range of temperature due to inherent property of its material.

## 3.3 LINEARIZING SCHEMES

NTC Thermistors are often used in wide variety of temperture measuring applications due to their foremost inherent properties of high resistivities and high negative temperature coefficient. The input-output response of the thermistor is highly nonlinear throughout their operating range of temperature. As a result, the measurement is usually restricted to narrow range of temperature for which the relation [eq.(3.1)] is approximately linear. One way to expand the region of operation is to 'linearize' the relationship with the aid of electronic circuitry.

Most of the approaches produce qualitative results while a few are loaded with quantitative information [41]. Each one out of these claims advantage over the previously developed approach. Here, two analog schemes, namely single thermistor active bridge and thermistor-logarithmic converter in conjunction with analog divider circuit have been analysed and developed for gain linearization in the range from 298 to 368 K.

### 3.3.1 Single-Thermistor-Active-Bridge Analysis

In the year 1974, Broughton [15] analysed six circuits using single thermistor for response linearization. Out of these circuits, three are around operational amplifier bridges, where thermistors are placed in input branch of the bridges. For gain-linearization, the response of the circuits is transformed in linear fraction and the resulting transfer function is expanded in Taylor's infinite series about reference temperature. The second and all higher coefficients in the series are equated to zero in the operating temperature range to obtain a necessary condition. On the basis of obtained conditions, the elements of the circuit are selected to obtain linear response in the desired range. The approach [15] has limitation, as the condition for linearity is dependent upon the thermistor used in the individual circuit. In the present work, single-thermistor has been used in feed-back path of operational amplifier instead of in input branch of the circuit like Broughton [15], and analysis has been carried out for obtaining optimum condition for the gain-linearization. This condition is independent of thermistor specifications.

The analysis has been carried under few assumptions. Self-heating of thermistor and effects of circuit output loading are neglected. Further, the operational amplifier is considered to be ideal one exhibiting zero output current, zero output impedence, and infinite gain.

The Resistance-temperature relation for thermistor given in equation (3.1) is expressed as

$$R_{(T_1)} = R_{(To)} \exp b\,(T_1^{-1} - To^{-1}) \qquad \ldots (3.2)$$

By defining, (Fig 3.1)

$$\frac{R_{(T_1)}}{R_4} = r(x) \qquad \ldots (3.3)$$

$$\frac{R_{(To)}}{R_4} = ro \qquad \ldots (3.4)$$

$$\frac{T_1}{To} = x \qquad \ldots (3.5)$$

$$\frac{b}{To} = \beta \qquad \ldots (3.6)$$

$$\frac{R_1}{R_3} = \alpha \qquad \ldots (3.7)$$

and $$\frac{e_o}{e_i} = K; \qquad \ldots (3.8)$$

the standard relation of equation (3.1) can be written in normalized form as below

$$r(x) = r_o \exp\,[\beta(\frac{1}{x} - 1)] \qquad \ldots (3.9)$$

The output-input voltage ratio (gain) of the circuit shown in Fig. 3.1 can be written as

$$\frac{e_o}{e_i} = K = \frac{R_4 R_1 - R_{(T_1)} R_3}{R_4 (R_3 + R_1)}$$

$$K = \frac{\dfrac{R_1 R_4}{R_1 R_{(T_1)}} - \dfrac{R_{(T)} R_3}{R_1 R_{(T_1)}}}{\dfrac{R_4 R_3}{R_1 R_{(T_1)}} + \dfrac{R_4 R_1}{R_{T_1} R_1}} = \frac{\dfrac{R_4}{R_{(T_1)}} - \dfrac{R_3}{R_1}}{\dfrac{R_4 R_3}{R_1 R_{(T_1)}} + \dfrac{R_4}{R_{(T_1)}}} \qquad \ldots (3.10)$$

By substitution of normalized values from equations (3.3), (3.4), and (3.7) in equation (3.10), we have

$$K = \frac{\dfrac{1}{r(x)} - \dfrac{1}{\alpha}}{\dfrac{1}{r(x)} \cdot \dfrac{1}{\alpha} + \dfrac{1}{r(x)}} \qquad \ldots (3.10\text{-a})$$

Normalized gain expression of equation (3.10-a) after linear fractional transformation can be rewritten as

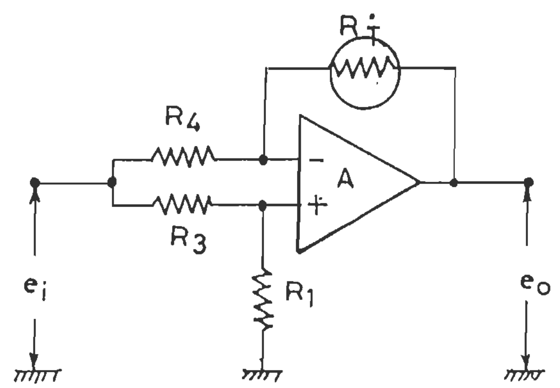
$$K = \frac{\alpha - r}{\alpha + 1} \qquad \ldots (3.11)$$

After linear fractional transformation in r, the K can be expressed in the form

$$K = \frac{a_o + a_1 r}{b_o + b_1 r} \qquad \ldots (3.11\text{-a})$$

where, $a_o$, $a_1$, $b_o$, $b_1$ are gain coefficients and given as;

$a_o = \alpha$, $a_1 = -1$, $b_o = \alpha$, and $b_1 = \frac{1}{r}$




[$e_i$ = 1.13V, $R_1$ = 6.2K, $R_T$ = THERMISTOR(EUP 4.7K), $R_3$ = 9.4K, $R_4$ = 7.12K]

FIG. 3.1 SINGLE THERMISTOR ACTIVE BRIDGE AMPLIFIER

Now for gain-linearization, the gain K, given by equation (3.11) is expanded in terms of Taylor's infinite series about x = 1, i.e., $\frac{T_1}{T_o}$

$$K(x) = K(1) + \left.\frac{\partial K}{\partial x}\right|_{x=1}(x-1) + \frac{1}{2}\left.\frac{\partial^2 K}{\partial x^2}\right|_{x=1}(x-1)^2 + \frac{1}{6}\left.\frac{\partial^3 K}{\partial x^3}\right|_{x=1}(x-1)^3 + \ldots \qquad \ldots (3.12)$$

$$= K_{(o)} + K_1(x-1) + K_2(x-1)^2 + K_3(x-1)^3 + \ldots \qquad \ldots (3.13)$$

where,

$$K_{(o)} = K_{(1)}, K_1 = \left.\frac{\partial K}{\partial x}\right|_{x=1}, \ K_2 = \frac{1}{2}\left.\frac{\partial^2 K}{\partial x^2}\right|_{x=1}, \ K_3 = \frac{1}{6}\left.\frac{\partial^3 K}{\partial x^3}\right|_{x=1}$$

Linear fractional transformation of gain, K given by equation (3.11-a), is partially differentiated w.r.t. r,

$$\frac{\partial K}{\partial r} = \frac{(b_o+b_1 r)\frac{\partial}{\partial r}(a_o+a_1 r) - (a_o+a_1 r)\frac{\partial}{\partial r}(b_o+b_1 r)}{(b_o + b_1 r)^2}$$

$$= \frac{a_1(b_o+b_1 r)-b_1(a_o+a_1 r)}{(b_o + b_1 r)^2} = \frac{a_1 b_o - b_1 a_o}{(b_o+b_1 r)^2} \qquad (3.14\text{-a})$$

Normalized relation of r, given by equation (3.9) is partially differentiated w.r.t. x to get $(\frac{\partial r}{\partial x})$ as

$$r = r_o e^{\beta(\frac{1}{x} - 1)}$$

$$\frac{\partial r}{\partial x} = r_o e^{\beta(\frac{1}{x} - 1)} \cdot \beta(-\frac{1}{x^2} - 0) = -\frac{r\beta}{x^2} \qquad (3.14\text{-b})$$

Now first partial derivative of K[eq.(3.11-a)] w.r.t. x, is obtained as

$$(\frac{\partial K}{\partial x}) = (\frac{\partial K}{\partial r}) \cdot (\frac{\partial r}{\partial x})$$

on substitution from equations (3.14-a) and (3.14-b), respectively

$$\frac{\partial K}{\partial x} = \frac{a_1 b_o - b_1 a_o}{(b_o + b_1 r)^2} \cdot \left(-\frac{\beta r}{x^2}\right) \qquad \ldots (3.14)$$

For second derivative of K, $\left(\frac{\partial K}{\partial x}\right)$ is partially differentiated w.r.t. x.

$$\frac{\partial^2 K}{\partial x^2} = \frac{\partial}{\partial x} \cdot \left(\frac{\partial K}{\partial x}\right)$$

$$= \frac{\partial}{\partial x}\left[\left(-\frac{\beta r}{x^2}\right) \cdot \left(\frac{a_1 b_o - b_1 a_o}{(b_o + b_1 r)^2}\right)\right]$$

$$= -\beta(a_1 b_o - b_1 a_o) \frac{\partial}{\partial x}\left[\frac{1}{x^2} \cdot r \cdot \frac{1}{(b_o + b_1 r)^2}\right]$$

$$\left(\frac{\partial^2 K}{\partial x^2}\right) = \frac{\beta r}{x^3} \cdot \frac{a_1 b_o - b_1 a_o}{(b_o + b_1 r)^2} \cdot \left[2 + \frac{\beta}{x} \cdot \left(\frac{b_o - b_1 r}{b_o + b_1 r}\right)\right] \qquad \ldots (3.15)$$

and

$$\left(\frac{\partial^3 K}{\partial x^3}\right) = -\frac{\beta r}{x^4} \cdot \frac{a_1 b_o - b_1 a_o}{(b_o + b_1 r)^2} \cdot \left[\frac{\beta^2}{x^2} \cdot \frac{b_o^2 - 4 b_o b_1 r - b_1^2 r^2}{(b_o + b_1 r)^2}\right.$$

$$\left. + 6\frac{\beta}{x} \cdot \frac{b_o - b_1 r}{b_o + b_1 r} + 6\right] \qquad \ldots (3.16)$$

Now by making use of equations (3.9), (3.11-a), (3.14), (3.15), and (3.16), at $x = 1$, $r = r_o$, the following relations are obtained.

$$K_o = \frac{a_o + a_1 r_o}{b_o + b_1 r_o} \qquad \ldots (3.17)$$

$$K_1 = \beta r_o \frac{a_o b_1 - b_o a_1}{(b_o + b_1 r_o)^2} \qquad \ldots (3.18)$$

$$K_2 = -\beta r_o \frac{a_o b_1 - a_1 b_o}{(b_o + b_1 r_o)^2} \left[\frac{\beta}{2} \cdot \frac{b_o - b_1 r_o}{b_o + b_1 r_o} + 1\right]$$

$$= -K_1 \ (A + 1) \qquad \ldots (3.19)$$

$$K_3 = \beta r_o \frac{a_o b_1 - a_1 b_o}{(b_o + b_1 r_o)^2} \left[ \frac{\beta^2}{6} \cdot \frac{b_o^2 - 4 b_o b_1 r_o + b_1^2 r_o^2}{(b_o + b_1 r_o)^2} + \beta \cdot \frac{b_o - b_1 r_o}{b_o + b_1 r_o} + 1 \right]$$

$$= K_1 (B + 2A + 1) \qquad \ldots (3.20)$$

where,

$$A = \frac{\beta}{2} \cdot \frac{b_o - b_1 r_o}{b_o + b_1 r_o} \qquad \ldots (3.21)$$

and

$$B = \frac{\beta^2}{6} \cdot \frac{b_o^2 - 4 b_o b_1 r_o + b_1^2 r_o^2}{(b_o + b_1 r_o)^2} \qquad \ldots (3.22)$$

On substituting $a_o, a_1, b_o, b_1$ values and for $r = r_o$

$$A = \frac{\beta}{2} \cdot \frac{\alpha - 1}{\alpha + 1} \qquad \ldots (3.23)$$

and

$$B = \frac{\beta^2}{6} \; \frac{\alpha^2 - 4\alpha \cdot \frac{r_o}{r_o} + \frac{r_o^2}{r_o^2}}{(\alpha + \frac{1}{r_o} \cdot r_o)^2} = \frac{\beta^2}{6} \cdot \frac{\alpha^2 - 4\alpha + 1}{(\alpha + 1)^2} \qquad \ldots (3.24)$$

For gain-linearization, second and all higher coefficients of the Taylor's series are made equal to zero in the operating temperature range.

From Equation (3.19),

$$K_2 = - K_1 (A + 1) = 0$$

$$K_2 = 0, \text{ if } (A+1) = 0$$

$$A = -1$$

$$A+1 = \frac{\beta}{2} \cdot \frac{b_o - b_1 r_o}{b_o + b_1 r_o} + 1 = 0$$

$$\frac{\beta}{2} \cdot \frac{b_o - b_1 r_o}{b_o + b_1 r_o} = -1$$

$$\frac{\beta}{2} \cdot \frac{\alpha - 1}{\alpha + 1} = -1 \text{ (on substitution of values of } b_o \text{ and } b_1)$$

$$\frac{\alpha - 1}{\alpha + 1} = -\frac{2}{\beta}$$

$$\alpha = \frac{\beta - 2}{\beta + 2} \qquad \ldots (3.25)$$

The value of $\alpha$ from equation (3.25) is substituted in equation (3.24) to obtain B in terms of $\beta$ as

$$B = \frac{\beta^2}{6} \cdot \frac{b_o^2 - 4b_1 b_o r_o + b_1^2 r_o^2}{(b_o + b_1 r_o)^2} = \frac{\beta^2}{6} \cdot \frac{(\alpha-1)^2 - 2\alpha}{(\alpha+1)^2}$$

$$= \frac{\beta^2}{6} \left| \frac{\{(\frac{\beta-2}{\beta+2}) - 1\}^2 - 2\,(\frac{\beta-2}{\beta+2})}{\{(\frac{\beta-2}{\beta+2}) + 1\}^2} \right|$$

$$B = \frac{-\beta^2 + 12}{12} \qquad \ldots (3.26)$$

For the circuit shown in Fig.(3.1), for $K_o = 0$ and $\alpha$ given by equation (3.25), equation (3.11) becomes as

$$\frac{e_o}{e_i} = K(x) = [\frac{\alpha - r}{\alpha + 1}] \text{ and } K_o = \frac{a_o + a_1 r}{b_o + b_1 r} = 0; \text{ i.e. } a_o = -a_1 r_o$$

$$r_o = (-\frac{a_o}{a_1}) = -(\frac{\alpha}{-1})$$

So, $r_o = \alpha \Big|_{K_o=0}$

When the bridge is under balance condition, for $r_o = \alpha$

i.e., $\frac{R_{(To)}}{R_4} = \frac{R_1}{R_3}$

$$K(x) = \frac{r_o - r}{r_o + 1}\Bigg|_{r_o=\alpha \text{ for } K_o=0}$$

$$= \frac{r_o - [r_o \exp \beta(\frac{1}{x} - 1)]}{r_o + 1} = \frac{\alpha - \alpha.\exp \beta(\frac{1}{x} - 1)}{\alpha + 1}$$

$$K(x) = \frac{(\frac{\beta-2}{\beta+2}) - (\frac{\beta-2}{\beta+2})\exp \quad (\frac{1}{x} - 1)}{(\frac{\beta-2}{\beta+2}) + 1}$$

$$K(x) = \frac{(\beta-2)\{1-\exp \beta(\frac{1}{x} - 1)\}}{2\beta} \qquad \ldots (3.27)$$

Again from equation (3.20); $K_3 = K_1(B+2A+1)$

$$K_3 = K_1[-\frac{\beta^2+12}{12} + 1-2] = (\frac{-\beta^2}{12})K_1 \qquad \ldots (3.28)$$

Fractional deviation, D from linearity is defined by expression

$$|D| = \left|\frac{K(x)-K_1(x-1)}{K_1(x-1)}\right| \qquad \ldots (3.29)$$

K(x) is obtained approximately by equation (3.13) as

$$K(x) - K_1(x-1) = K_3(x-1)^3$$

$$|D| = \left|\frac{K_3}{K_1}\right| \frac{(x-1)^3}{(x-1)} = \left|\frac{K_3}{K_1}\right| (x-1)^2$$

$$|D| = \left|\frac{K_3}{K_1}\right| (x-1)^2 \qquad \ldots (3.30)$$

$$|D| = \left|\frac{\beta^2}{12}\right| \frac{K_1}{K_1}(x-1)^2 = \left|\frac{\beta^2}{12}\right| (x-1)^2 \qquad \ldots (3.31)$$

Equation (3.31) relates fractional deviation D, to normalized thermistor parameter, $\beta$ and normalized temperature deviation $(x-1)$.

If $(x_D-1)$ is the maximum normalized temperature change to produce desired deviation, D, then from equation (3.31)

$$(x_D-1) = \frac{2\sqrt{3D}}{\beta} \qquad \ldots (3.32)$$

For balance condition of the bridge, linear gain coefficient $K_1$, is given as

$$K_1 = \beta r_o \frac{a_o a_1 - a_1 b_o}{(b_o + b_1 r_o)^2} = \beta r_o \frac{\alpha \cdot \frac{1}{r_o} - (-1)\alpha}{\left(\alpha + \frac{r_o}{r_o}\right)^2}$$

$$= \frac{\beta\alpha(1+r_o)}{(1+\alpha)^2}$$

$$= \beta\alpha(1+r_o)(1+\alpha)^{-2} \qquad \ldots (3.33)$$

3.3.1.1 Design Example

Condition for linearity has been obtained in equation (3.25) and accordingly the circuit parameters are selected. Thermistor (EUP 4.7K) having $R_{(To)}$ equal to 4.7K at 25°C and b equal to 3000K is selected as sensor to measure temperature in the range from 298 to 368 K with maximum output voltage magnitude of about 1.0 volt and maximum gain-slope deviation, D equals to 0.4

$$x_{min} = \frac{298}{298} = 1 \qquad (x = \frac{T_1}{To}) \qquad \ldots (3.34)$$

$$x_{max} = \frac{368}{298} = 1.2181208 \qquad \ldots (3.35)$$

$$\beta = \frac{b}{To} = \frac{3000}{298} = 10.067114 \approx 10.06$$

Again from equation (3.32) the value of β is calculated for D = 0.4

$$\beta = \frac{2\sqrt{3D}}{x_D - 1} = \frac{2\sqrt{3x0.4}}{0.2181208} = 10.040375$$

$$\simeq 10.04$$

Under balanced condition of the bridge as given by equation (3.25),

$$r_o = \alpha = \frac{\beta - 2}{\beta + 2} = \frac{10.06-2}{10.06+2} = 0.668325$$

$$\simeq 0.66$$

From equation (3.33), gain coefficient $K_1$ is calculated as

$$K_1 = \beta\alpha(1+\alpha)^{-1}\Big|_{\alpha = r_o}$$

$$= \frac{\beta-2}{2} = \frac{10.06-2}{2} = 4.03 \qquad \ldots (3.36)$$

From equation (3.32), gain-slope deviation, D is calculated as,

$$|D| = \left|\frac{\beta^2}{12}\right| . (x_D - 1)^2$$

$$= \frac{10.06 x 10.06}{12} \text{ x } (0.2181208)^2$$

$$= 0.4012443 \simeq 0.4$$

The bridge output is zero at reference temperature To = 298K and 1.0 volt (approximately) at 368K. From equations (3.8) and (3.35), the excitation voltage to bridge is calculated as

$$\frac{e_o}{e_i} = K = K_1(x_D - 1) \qquad \text{... (3.37)}$$

$$e_o = 4.03(0.2181208)e_i$$

$$e_i = \frac{1}{0.8790268} = 1.3171671 \text{ volt}$$

$$\simeq 1.31 \text{ volt (approximately)}$$

After deciding the input voltage, resistors $R_1$,$R_3$, and $R_4$ (of the bridge) are selected using equations (3.3) and (3.4).

$$r_o = \frac{R_{(To)}}{R_4} = \frac{4.7K}{R_4} = 0.66$$

$$R_4 = 7.1212121K \text{ ohms}$$

$$\simeq 7.12 \text{ K ohms}$$

For $R_3 = 9.4$ K ohms

$$\frac{R_1}{R_3} = \alpha = 0.66$$

$$R_1 = 0.66 \times 9.4 \text{ K ohms} = 6.204 \text{ K ohms} \simeq 6.2 \text{ K ohms}$$

A circuit for this temperature sensor, $R_T^{\cdot}$ with resistors $R_1$, $R_3$, and $R_4$ is shown in Fig. 3.1. Complete details of the circuit are shown in Fig. 3.2. A constant dc supply is exciting the bridge $A_1$, at +1.13 volt. The bridge output $+e_o$ is fed to an unity gain amplifier $A_2$, to isolate linear temperature dependent voltage from digital voltmeter (DVM) which is connected at output stage of this circuit. DVM is directly calibrated in terms of temperature from 298 to 368 K.

3.3.1.2 Results and Discussions

The plot between the bridge output voltage and temperature is shown in Fig. 3.3.. The maximum output actually obtained from the circuit is 0.365 volt. This is due to higher value of b and larger value of $(x_D-1)$. The input-output relationship is linear in two segments; one segment is from 298 K (reference temperature) to 323 K and the other one is from 323 K to 368 K. There is a major change in the slope of the curve around 323 K as shown in Fig. 3.3.. The nature of normalized transducer characteristic for the active bridge is shown in Fig. 3.4.. Thus, the linearization is obtained in two segments. Obviously, it is due to the fact that resistance versus temperature characteristic of thermistor may be regarded

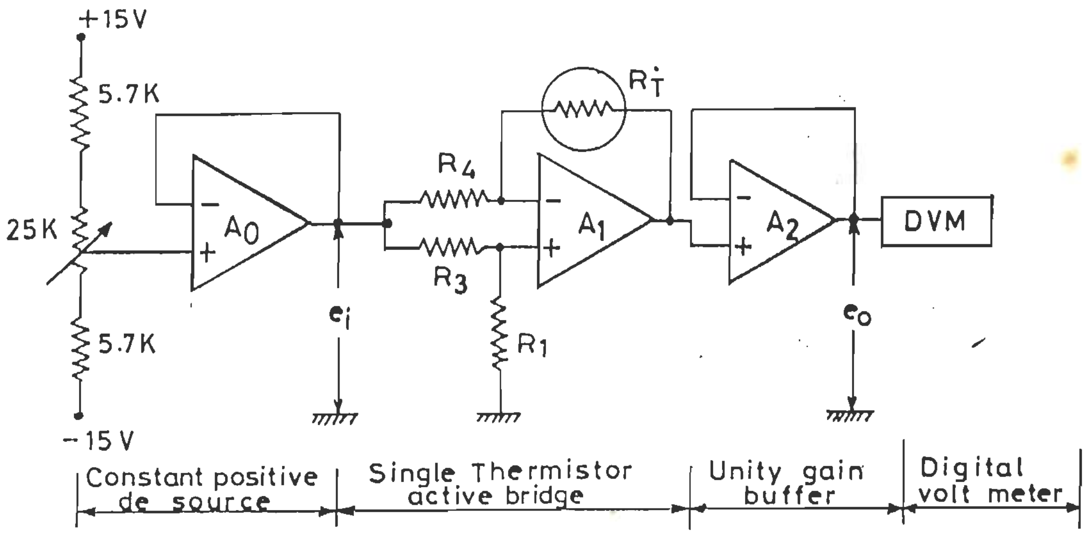



FIG. 3.2 DETAILS OF CIRCUIT

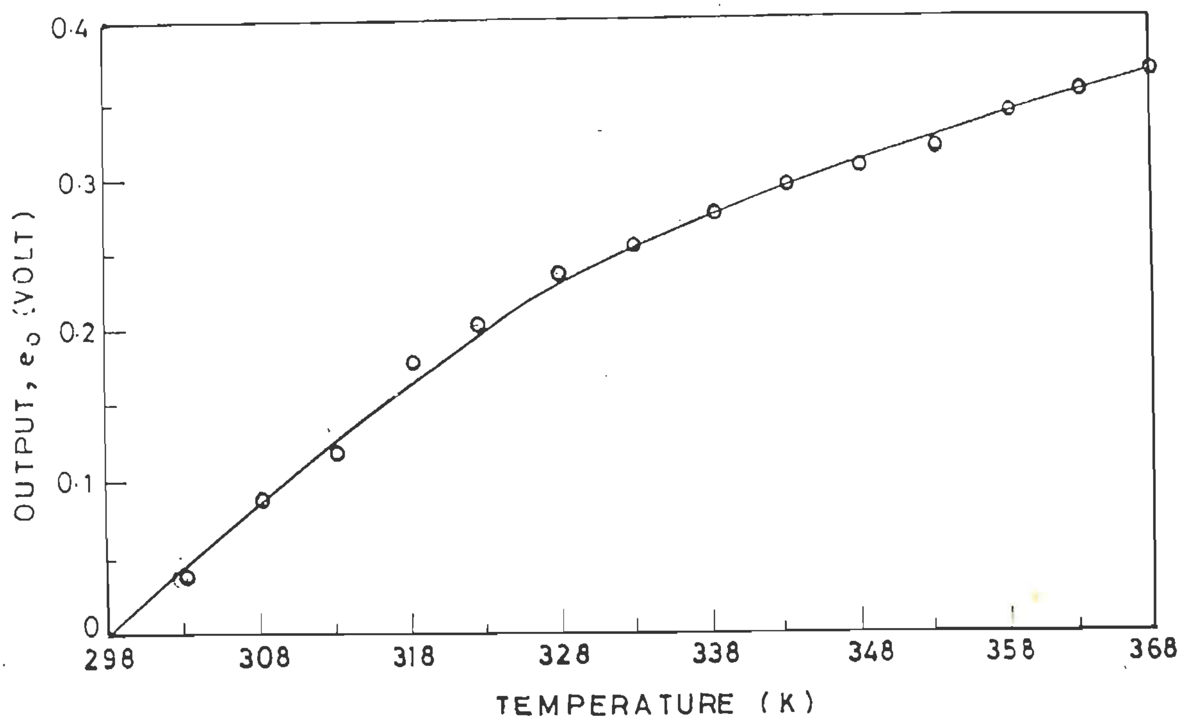



FIG. 3.3 PLOT BETWEEN BRIDGE OUTPUT AND TEMPERATURE

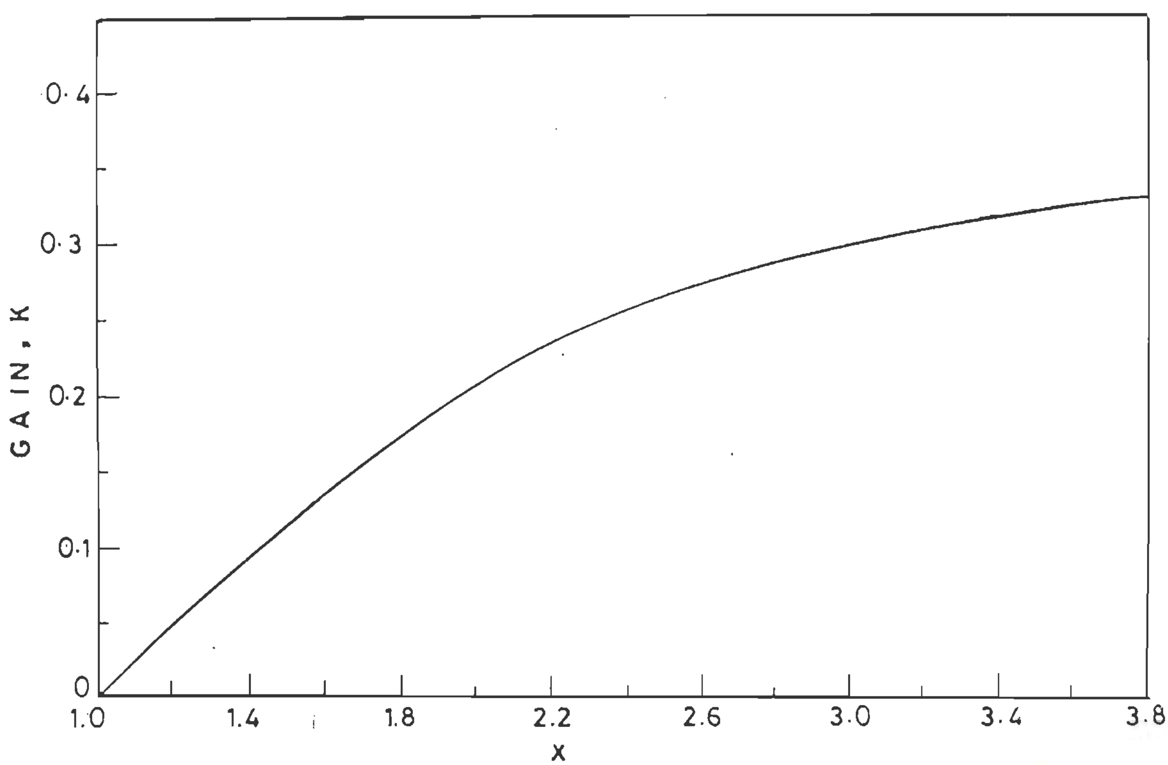



FIG. 3.4 NORM. TRANSDUCER CHARACTERISTICS, GAIN Vs X

as exhibiting a large decay (rapid) in its resistance value over a temperature range from 0 to 323 K (approximately) and then slow decay from 323 K to 373 K, Fig.3.5. The input-output relationship achieved here is linear for 80 per cent of its full span.

The optimum condition of linearity is obtained which is independent of the thermistor specification. The bridge is excited at a low level to avoid self-heating in the thermistor. Due to the limitations of temperature controlling facilities the experiment has been carried out from room temperature upto 368 K. The results are in agreement with those obtained in earlier works [15].

### 3.3.2 Thermistor-Logarithmic Network Coupled with Analog Divider Unit

Logarithmic and antilogarithmic amplifiers are promising circuits for obtaining linear input-output response of thermistors in addition to their use as multiplier, divider and square-root circuits. There are many phenomena in physics which have exponential decay with time. Such results can also be conveniently displayed as a straight line decay by using log converter. Several analog linearizing schemes for the measurement of temperature by making use of thermistors as a temperature sensor in log network have been developed in the past and there is continual effort to improve their performance [20,51,53,54,55,78]. Recently in 1988, Patranabis, et al. [78] described a general technique for linearizing

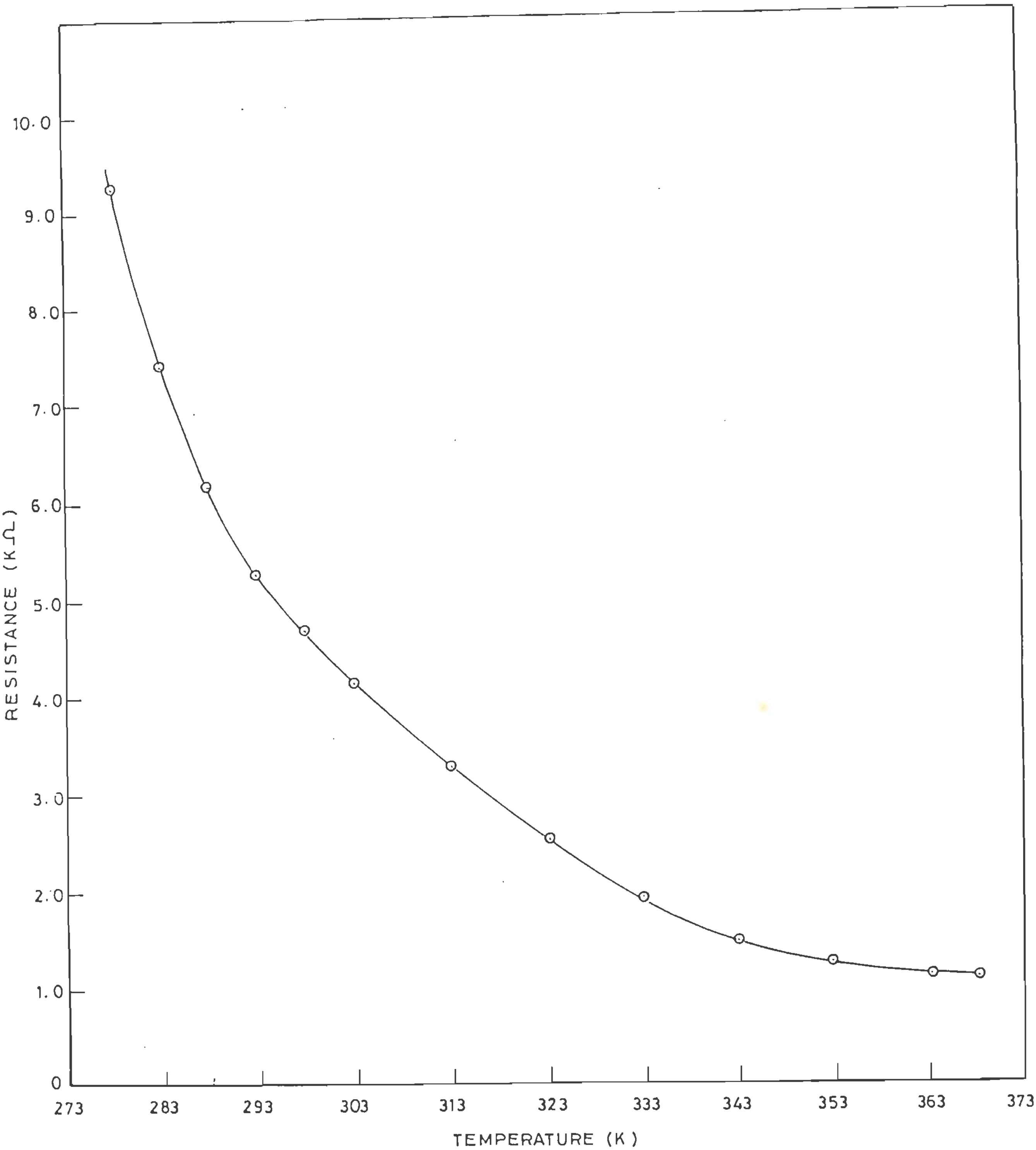



FIG. 3.5 RESISTANCE VS TEMPERATURE PLOT FOR THERMISTOR EUP(4.7K)

transducer characteristics falling in exponentially rising and decaying categories with a particular emphasis on an analog scheme for thermistor gain-linearization. In this technique, thermistor in conjunction with log converter and field effect transistor (FET) has been used to obtain linear relationship between temperature and its electrical equivalent output in the range from 30 to 95°C. The linearization could be obtained either by inverting the converted variable or by compensating the non-linearity. In their work, they have used the compensation technique using FET as circuit component. It was also indicated that the inverter is not a commonly available module to be easily coupled with log amplifier for inverting the (converted) variable.

In the present work, a scheme employing thermistor-logarithmic network has been developed to obtain gain-linearization over a wider temperature range from 298 to 363 K. A monolithic analog divider circuit has been used as it facilitates to 'invert' the converted variable for obtaining a linear relationship between input and output parameters. The general analysis has been carried out in the next section to establish the linear relationship between temperature and output of the scheme.

3.3.2.1 Analysis

Instrumentation scheme involves single thermistor inverting amplifier (A) in conjunction with log amplifier

(LA) and analog divider unit (D). Excitation level to the inverting amplifier is kept constant at $+e_i$ volt from a power supply source (P). The block diagram of the scheme is shown in Fig. 3.6 . The scheme has been developed for the temperature measurement from To (298 K) to $T_1$ (368 K).

The inverting amplifier produces a temperature dependent voltage which is given as

$$V_1 = e_i \frac{R_T^{\cdot}}{R_o} \text{ volt} \qquad \ldots (3.25)$$

where, $e_i$ is the excitation voltage,

$R_T^{\cdot}$ is thermistor modelled by equation (3.1), and

$R_o$ is the fixed value resistor equivalent to resistance of the thermistor at absolute temperature, $T_1$.

The voltage $V_1$ forms input to log amplifier and its output is expressed as

$$V_{out} = \frac{kTo}{q} \ell_n\left(\frac{V_1}{R_i \cdot I'_o}\right) \text{ volt} \qquad \ldots (3.26)$$

where, k is the Boltzmen's constant = $1.38 \times 10^{-23}$ J/K.

To is the temperature in K (at room temperature, 298 K),

q is the charge on electron = $1.60 \times 10^{-19}$ C,

$\ell_n$ is the natural log (base = 2.718),

$V_1$ is the input voltage,

$R_i$ is the input resistance of log amp, and

$I'_o$ is the reverse leakage current of transistor (= 1 μ amp for SL 100 transistor).

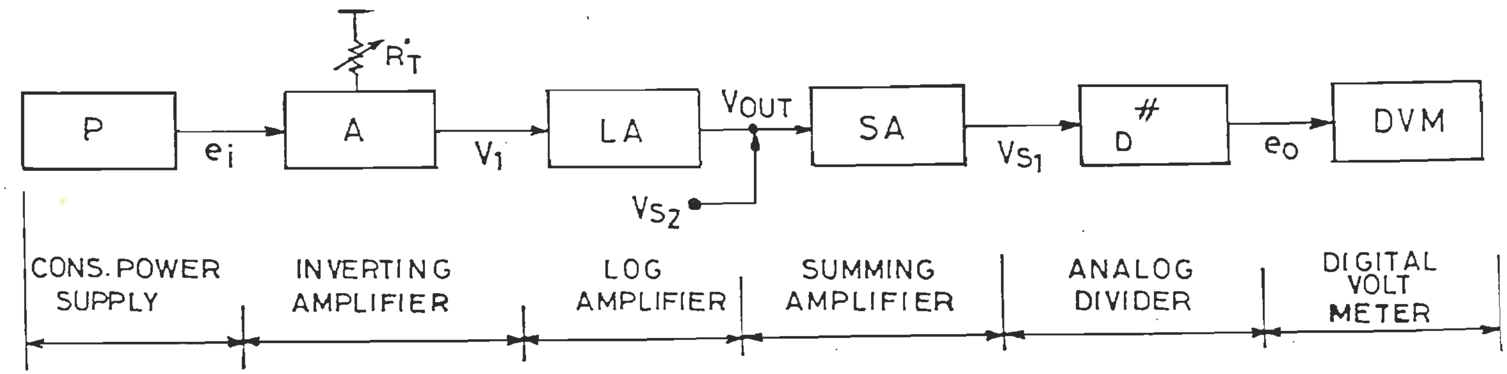



Figure-3.6 BLOCK DIAGRAM OF THE SCHEME

On substitution of $V_1$ from equation (3.25) in equation (3.26),

$$V_{out} = \frac{kTo}{q} \ln\left[ \frac{e_i \frac{R^{\bullet}_T}{R_o}}{R_i \cdot I'_o} \right]$$

$$= \frac{kTo}{q}\left[ \ln\left(e_i \frac{R^{\bullet}_T}{R_o}\right) - \ln(R_i \cdot I'_o)\right]$$

$$= \frac{kTo}{q} \left[ \ln e_i + \ln R^{\bullet}_T - \ln R_o - \ln R_i - \ln I'_o \right]$$

$$= \frac{kTo}{q}\left[ \ln e_i + \ln\{R_{To} \exp b(T_1^{-1} - To^{-1})\} - \ln R_o - \ln I'_o - \ln R_i \right] \text{ (on substitution of } R^{\bullet}_T \text{ from equation (3.1))}$$

$$= \frac{kTo}{q} \left[ \ln e_i + \ln R_{To} + \ln \exp b(T_1^{-1} - To^{-1}) - \ln R_o - \ln I'_o - \ln R_i \right]$$

$$= \left[ \frac{kTo}{q} \cdot \left(\frac{b}{T_1}\right) - \frac{kTo}{q}\left(\frac{b}{To}\right) + \frac{kTo}{q}(\ln e_i - \ln I'_o - \ln R_i - \ln R_o + \ln R_{To}) \right]$$

$$= \left[ \frac{kTo}{q} \cdot \left(\frac{b}{T_1}\right) \right] - \left[ \frac{kTo}{q}\left\{ \left(\frac{b}{To}\right) - (\ln e_i + \ln R_{To} - \ln R_o - \ln I'_o - \ln R_i)\right\} \right]$$

Thus $V_{out}$ can be written as

$$V_{out} = [V_{s1} - V_{s2}] \qquad \ldots (3.27)$$

$$\text{where, } V_{s1} = \frac{kTo}{q}.\left(\frac{b}{T_1}\right) \qquad \ldots (3.28)$$

$$\text{and } V_{s2} = \frac{kTo}{q}\left[\left(\frac{b}{To}\right) - \ell n\left(\frac{e_i . R_{To}}{R_o . R_i . I_o'}\right)\right] \qquad \ldots (3.29)$$

The $V_{out}$ consists of two voltage components; $V_{s1}$ and $V_{s2}$. $V_{s1}$ is temperature dependent while $V_{s2}$ is constant. $V_{s1}$ is inversely proportional to temperature $T_1$. The temperature dependent voltage, $V_{s1}$ is extracted out of $V_{out}$ by cancelling $V_{s2}$ by making use of summing amplifier (SA). The output of this amplifier is $V_{s1}$ only.

This temperature dependent voltage forms one out of three inputs $V_1$, $V_2$, and $V_3$ of analog divider unit (ref. Sec. 2.2.2, Ch-II). The output of the unit is

$$e_o = \left(\frac{V_1 . V_2}{V_3}\right) \text{ volt}$$

For, $V_1 = V_2 = +1.0$ volt (constant ref.), and $V_3 = V_{s1}$, the output voltage equals to

$$e_o = \frac{1.0 \times 1.0}{V_{s1}}$$

$$e_o = \frac{1.0 \times 1.0}{\frac{kTo}{q} . \frac{b}{T_1}}$$

$$= \frac{T_1}{\frac{kTo}{q} . b}$$

$$e_o = m\, T_1 \qquad \ldots (3.30)$$

where, $m = \dfrac{1}{\frac{kTo}{q} \cdot b}$ ; slope of the line presented by equation (3.30).

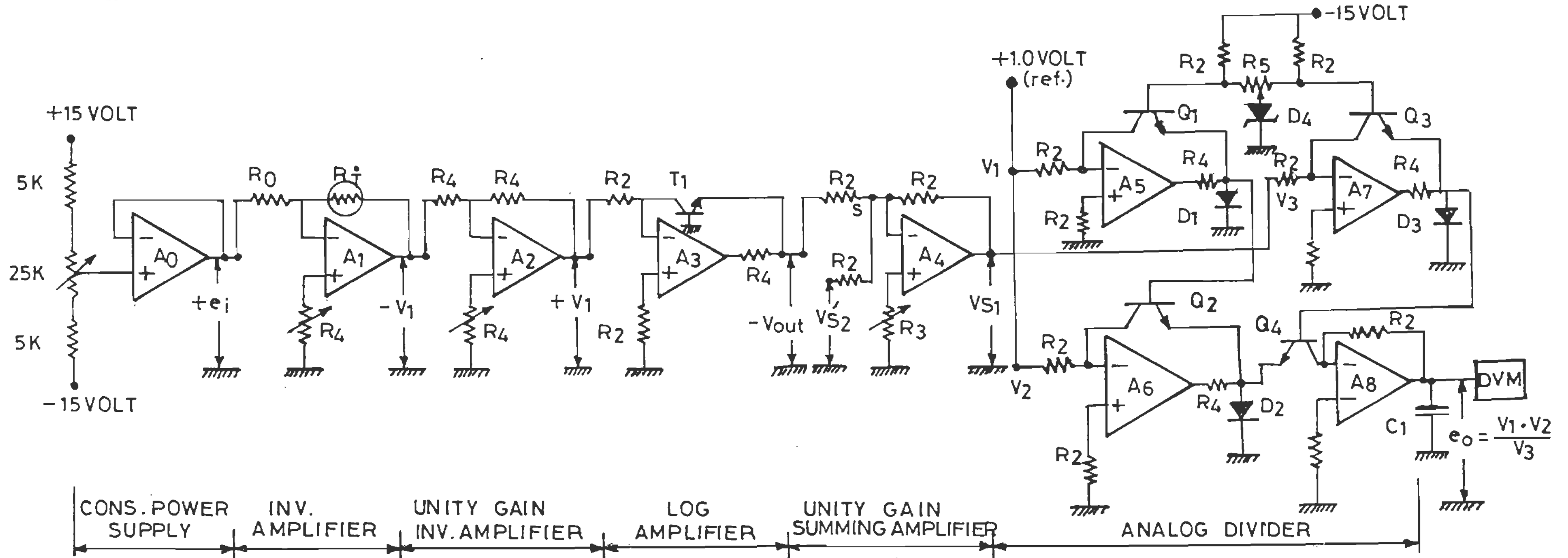
and $e_o \propto T_1$

Thus, output is linearly dependent upon temperature $T_1$ for a wide range and follows a straight line (y = mx) characteristic.

3.3.2.2 Hardware Implementation

The complete circuit diagram of the scheme is shown in Fig. 3.7. Thermistor (EUP 4.7 K) with $R_{(To)}$ equals to 4.7 K at 298 K and b equals to 3000 K is selected as temperature sensor. A voltage divider unit provides constant dc potential $e_i$ at +1.0 volt to single thermistor inverting amplifier $A_1$. The output voltage ($-V_1$) from $A_1$, is fed to unity gain amplifier $A_2$ to obtain $+V_1$. This is necessary as input to log amplifier should be a positive voltage. The voltage $V_1$ forms input to log amplifier $A_3$ and output of this amplifier is $V_{out}$. As discussed in preceding section, $V_{out}$ has two components: $V_{s1}$ and $V_{s2}$.

The output of log amplifier is an input to unity gain summing amplifier $A_4$. The other input $V'_{s2}$ to its summing node S is +0.353 volt, which in addition to $V_{s2}$ also accounts for offset voltage of the circuit.




[ $R_0$ = 1.1K, $R_T^{\circ}$; THERMISTOR (EUP4.7K), $R_2$ = 10K, $R_3$ = 47K, $R_4$ = 1K, $R_5$ = 220Ω, $C_3$ = 50μF, $T_1$ = SL100, ($Q_1$ − $Q_4$) = 2N3728 (matched pairs), ($A_0$ − $A_8$) = LM-324, $D_4$ = LM103, 2.4V.]

Figure 3.7 CIRCUIT DETAILS OF INSTRUMENTATION SCHEME

From equation (3.28);

$$V_{s1} = 0.0258(\frac{3000}{298}) \text{ volt} = 0.259 \text{ volt at } 298 \text{ K}$$

$$= 0.0258(\frac{3000}{368}) \text{ volt} = 0.210 \text{ volt at } 368 \text{ K}$$

and

From equation (3.27);

$$V_{s2} = \frac{kTo}{q} [(\frac{b}{To}) - \ell_n(\frac{e_i.R_{To}}{I'_o.R_o.R_i})] \text{ volt}$$

On substitution of values of different parameters.

$$V_{s2} = 0.0258[(\frac{3000}{298}) - (\frac{\log_{10}\frac{\{(1.0)x(4.7x10^3)\}}{\{(1x10^{-6})x(1.1x10^3)x(1x10^3)\}}}{\log_{10}e})]$$

$$= 0.0440433$$

$$\simeq 0.044 \text{ volt (approximately)}$$

From equation (3.27);

$$V_{out} = [V_{s1} - V_{s2}] \text{ (analytically)}$$

$$= (0.259-0.044) = 0.215 \text{ volt (approximately)}$$

But $V_{out}$ from practical circuit at 298 K = - 0.612 volt, which is due to circuit configuration. By applying one input as -0.612 volt (i.e. $V_{out}$) and the other as +0.353 volt (i.e. $V'_{s2}$) to the summing amplifier $A_4$, the output of +0.259 volt is achieved which is in agreement with its calculated value [from equation (3.28)]. $V'_{s2}$ is constant at +0.353 volt throughout the experimentation as its constituent components remain constant. The output voltage from $A_4$ forms one input to analog divider unit where as remaining two inputs $V_1$ and $V_2$ are held constant at +1.0 volt (reference).

The output $e_o$ from analog divider unit is given as

$$e_o = \frac{V_1 \cdot V_2}{V_3} \text{ volt}$$

where, $V_1 = V_2 = +1.0$ volt (ref.), and

$$V_3 = V_{s1} = 0.259 \text{ volt}$$

Thus $$e_o = \frac{1.0 \times 1.0}{0.259}$$

$$= 3.86 \text{ volt. (at 298 K)}$$

and, $$= 4.76 \text{ volt (at 368 K)}$$

3.3.2.3 Results and Discussions

Table-I shows output voltages response from log amplifier and analog divider unit from 298 to 368 K. Fig.3.8 and 3.9 show variation of $V_{out}$ versus temperature and $e_o$ versus temperature, respectively. It is clear that $V_{out}$ is inversely varying where as variation of $e_o$ versus temperature is linear. The nature of the plot, shown in Fig. 3.9, , is in conformity with equation (3.30).

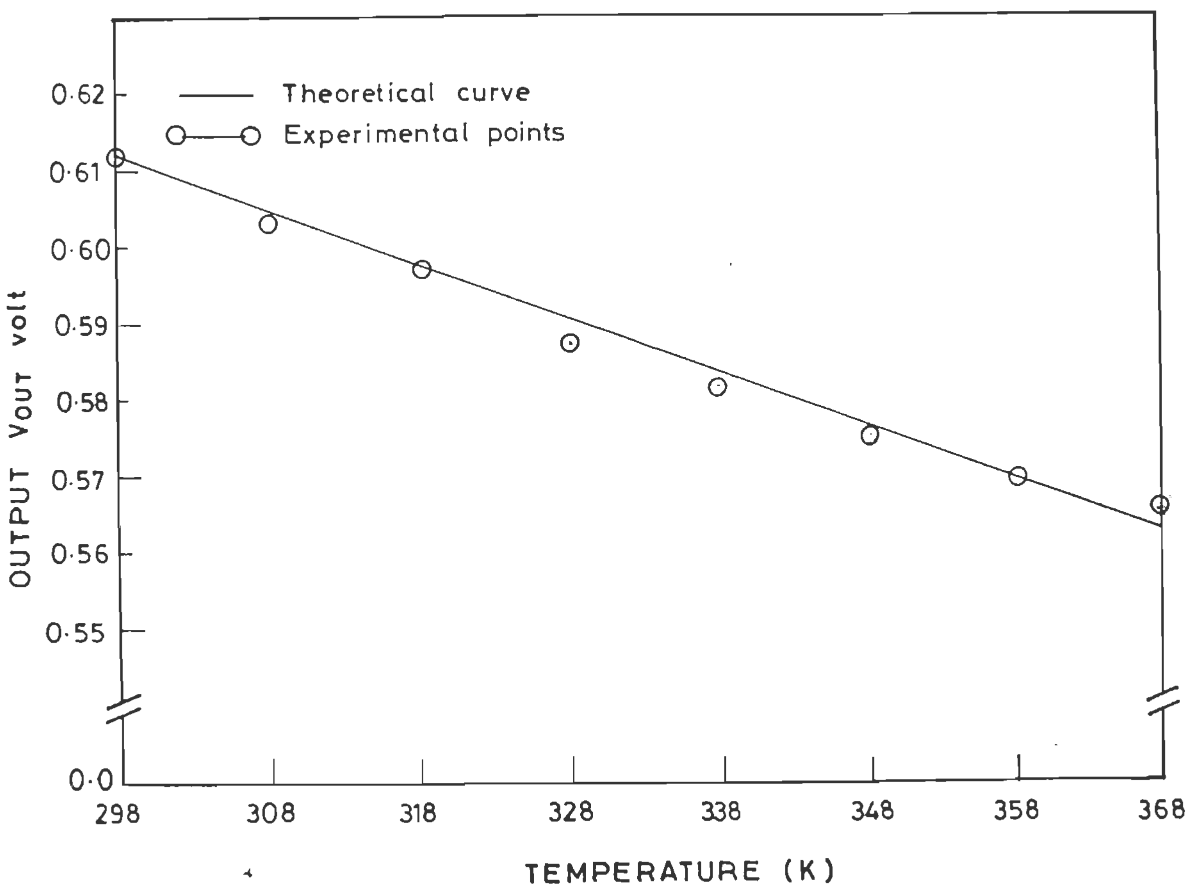



FIG. 3.8 PLOT OF Vout VERSUS TEMPERATURE

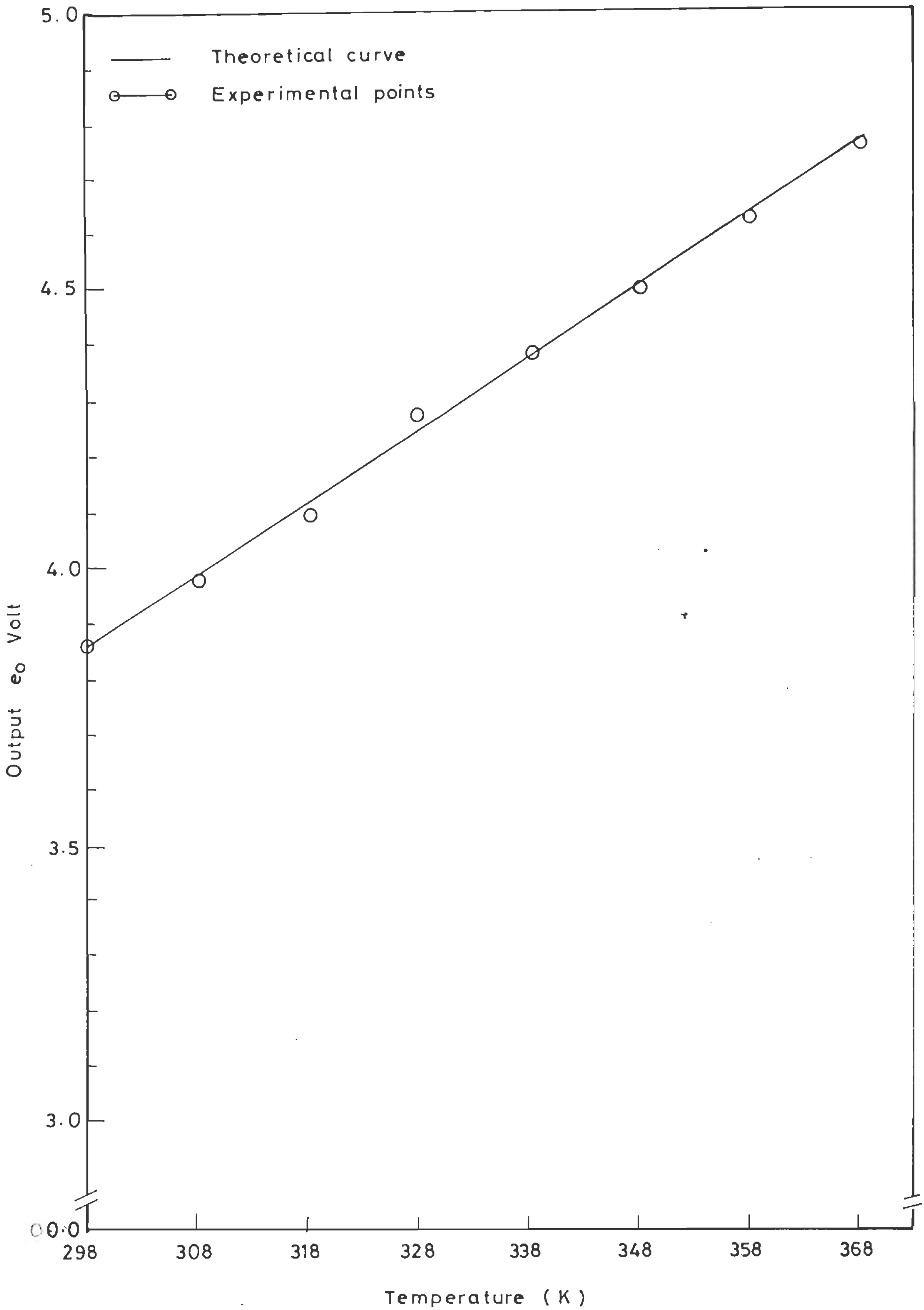



FIG.3.9 PLOT OF $e_o$ VERSUS TEMPERATURE

TABLE - I

| S.No. | Temperature, $T_1$ (K) | $V_{out}$; (log.amp. o/p) (volt) | $e_o$ (Analog divider (o/p) (volt) |
|---|---|---|---|
| 1 | 298 | -0.612 | 3.84 |
| 2 | 308 | -0.603 | 3.98 |
| 3 | 318 | -0.597 | 4.09 |
| 4 | 328 | -0.587 | 4.27 |
| 5 | 338 | -0.581 | 4.38 |
| 6 | 348 | -0.574 | 4.48 |
| 7 | 358 | -0.570 | 4.62 |
| 8 | 368 | -0.565 | 4.74 |

## 3.4 CONCLUSIONS

Two analog schemes for gain-linearization of NTC thermistor and of other transdancers exhibiting similar response characteristics have been developed and successfully tested in temperature range 298 to 368K. These circuits have been fabricated by making use of easily available electronic components, keeping their cost in main consideration. The hardware implementation may be further modified by using log and analog divider modules.

## CHAPTER - IV

# MICROPROCESSOR-BASED MULTIPOINT TEMPERATURE MEASUREMENT

## 4.1 INTRODUCTION

Temperature is one of the most important and vital parameters for measurement in large rating electrical apparatus and equipments, power plants, atomic plants, steel and fertilizer plants, ceramic and semiconductor manufacturing industries, and in many other industries. Temperature measurement is carried out in many ways and has been in general classified into two categories: Electrical and Electronic methods & nonelectrical methods. In nonelectrical methods, a change in basic properties of the object is observed due to variation in temperature, e.g., change in volume of a liquid, change in pressure of gas, change in vapour pressure, or change in dimensions of a solid on account of change in temperature. In electrical methods, the measurement is carried out measuring the voltage generated at the common junction of two dissimilar metals in thermocouples by measuring change in resistance of conducting materials in RTDs and by measuring change in resistance of semiconducting materials in thermistors. In the year 1970, Johnson [47] developed methods for continuous measurement and recording of temperature rise in motors and transformers. The test results were obtained by him on split-phase induction motor (½ hp) and two transformers (each of 3 kVA and 100 kVA ratings). Rele and Palmer [88] developed an automated data acquisition equipment for obtaining temperature rise at large number of points in 49 MVA single - phase transformer. Thermocouples were employed as sensors in the windings for temperature measurement of hot spots. The complete analysis was done for this technique on IBM 370/160 computer. MacMartin and Iwanusiw [65] described a method of direct current comparator for measuring the temperature rise of rotor of large (500 MW) generator. The test results were obtained for change in resistance of field winding within ±0.1 per cent to calculate the teperature rise within 0.25 per cent. Pietsch, et al. [81] investigated a family of organic compounds contained in generator insulation to provide an early warning of local overheating within the generator. Later in 1979, Gupta and Dewal [38] presented a digital technique for measuremnt and monitoring of average temperature of rotor field winding of brushless synchronous machine. In this technique, the temperature is measured as the function of change in resistance of the field winding due to rise in temperature. This (scheme) also includes a simple telemetry system. An eight channel multipoint temperature monitoring systme around Intel 8085 microprocessor was developed by Bhardwaj, et al. [8] to cover a temperature range from -100 to 1760°C by making use of various thermocouples. The output of these thermocouples has been linearized employing analog and digital method.

In another effort, McNutt, et al. [69] measured directly hot spot temperature of power transformer winding by making use of fibre optic sensors. In this technique, the average winding temperature rise over average oil temperature has the same slope versus winding current as the hot spot sensor temperature rise over top oil measurement. It was determined by resistance measurement technique. Thus it confirmed validity of the sensor in use. Microprocessor-based adaptive and prediction control algorithms were proposed by Sato, et al. [90] for steam temperature measurement in thermal power plants. In 1986, Mukosiej [70] described a method for the measurement of average value of temperature rise and distribution along a cylindrical surface with emphasis on rotor frame surface of an induction motor. Thermocouples as temperature sensors were used in this method. Yang and Lue [120] developed an electronic temperature controller by making use of analog-to-digital converter and a digital-to-analog converter around Apple II microcomputer. The nonlinearity of thermocouple, which was used as sensor, has been corrected employing a digital method. Recently in 1987, Rameshu and Shivaprasad [85,86] presented two microprocessor-based temperature indicating schemes using thermistor and platinum resistance thermistor (PRT) as temperature sensors. The linear performances were obtained by them for thermistors in the temperature range from 0 to 100°C with an accuracy of ±0.1°C and for PRT, in the temperature range from -183 to 200°C within accuracy limit of ±0.05°C.

It has been experienced that temperature measurement in electrical apparatus by detecting 'hot-spots' require installation of many sensors [88,8]. Further, this necessitates the development of specific type of signal conditioning and processing units, multiplexers and microprocessor and microcomputer-based systems for temperature detection and telemetry. Unfortunately, number of problems are encountered in these systems like non-linearity of transducer response, number of scanning sequence from different locations and necessity of reduction of number of link lines between the sensing and receiving ends. Simple and efficient computational algorithms for converting the received electrical signals into temperature at microprocessor end are not available. In this work, a microprocessor-based multichannel temperature scanning and telemetry scheme has been developed. The output voltage of earlier developed analog linearizing scheme in this work in Chapter-III, section 3.3 has been used for quick and continuous detection and monitoring of the temperature from eight different locations in a sequential order. Pulse-width-modulated (PWM) technique has been used for telemetry. The details of the scheme are given in the following section.

## 4.2 BASIC SCHEME

A multichannel temperature scanning and telemetry scheme, under the broad-based classification of temperature measurement by electronic methods, has been developed around Intel 8085 microprocessor. It measures and transmits the temperature signal from eight different locations in a sequential order. An effort has been made to minimize the snags of the earlier existing techniques [47,88,8,65]. The linearized output from single thermistor-active bridge circuit has been utilized for continuous detection and monitoring of the temperature. The number of links are reduced by making use of PWM technique for data transmission in conjunction with 8-channel multiplexer. A fast and efficient computational algorithm has been developed for electrical signal conversion into equivalent temperature at microprocessor end. The instrumentation scheme consists of a constant positive voltage supply source $A_0$,(NTC) thermistor-active-bridge amplifier $A_1$,ramp generator, comparator $A_3$, multiplexer M, I/O peripheral device and Intel-8085 microprocessor system. Fig. 4.1 shows the block diagram representation of the scheme.

## 4.3 DIGITAL DATA COMMUNICATIONS

The noise level affects the analog communication systems. Hence, instead of continuous-level-continous-time message signals, digital transmission of information (bearing signals) is efficient and convenient from the sources which produce symbols for communication purpose over long distance. To add to this, it has compatability with computers. Thus, conversion of the value of physical

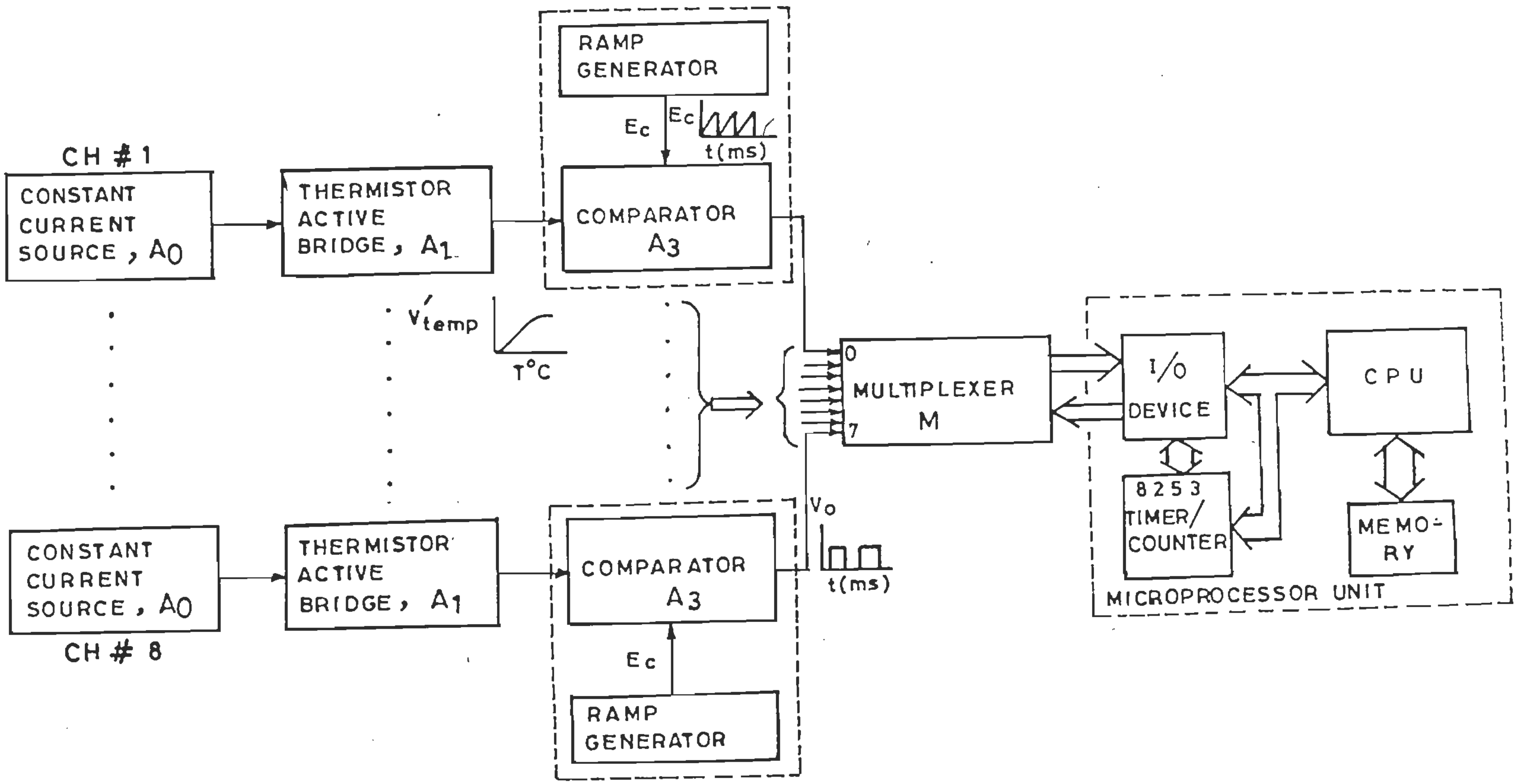



FIG. 4.1 BLOCK DIAGRAM OF MICROPROCESSOR - BASED MULTIPOINT TEMPERATURE SCANNING SCHEME

quantity into digital signal has some principal advantages and shows an upward trend. The sharp increase in digital communications increasingly at the expanse of analog communication is due to two interworking factors. Firstly, a lot of information to be transmitted, is in pulse form and secondly, the advent of LSI has permitted the use of complex coding systems and utilizes the channel capacities to the best possible extent. The PWM technique has the folloiwng advantages for telemetry [121]:

i) The information regarding signal is contained in the width of the pulse i.e., information signal proportional width of the pulse.

ii) The amplitude and total time period of carrier wave remain constant.

iii) Demodulation of PWM is simple. PWM is fed to an integrating circuit which produces a signal whose amplitude at any time is proportional to the width of pulse.

iv) The width of the pulse can not be negative.

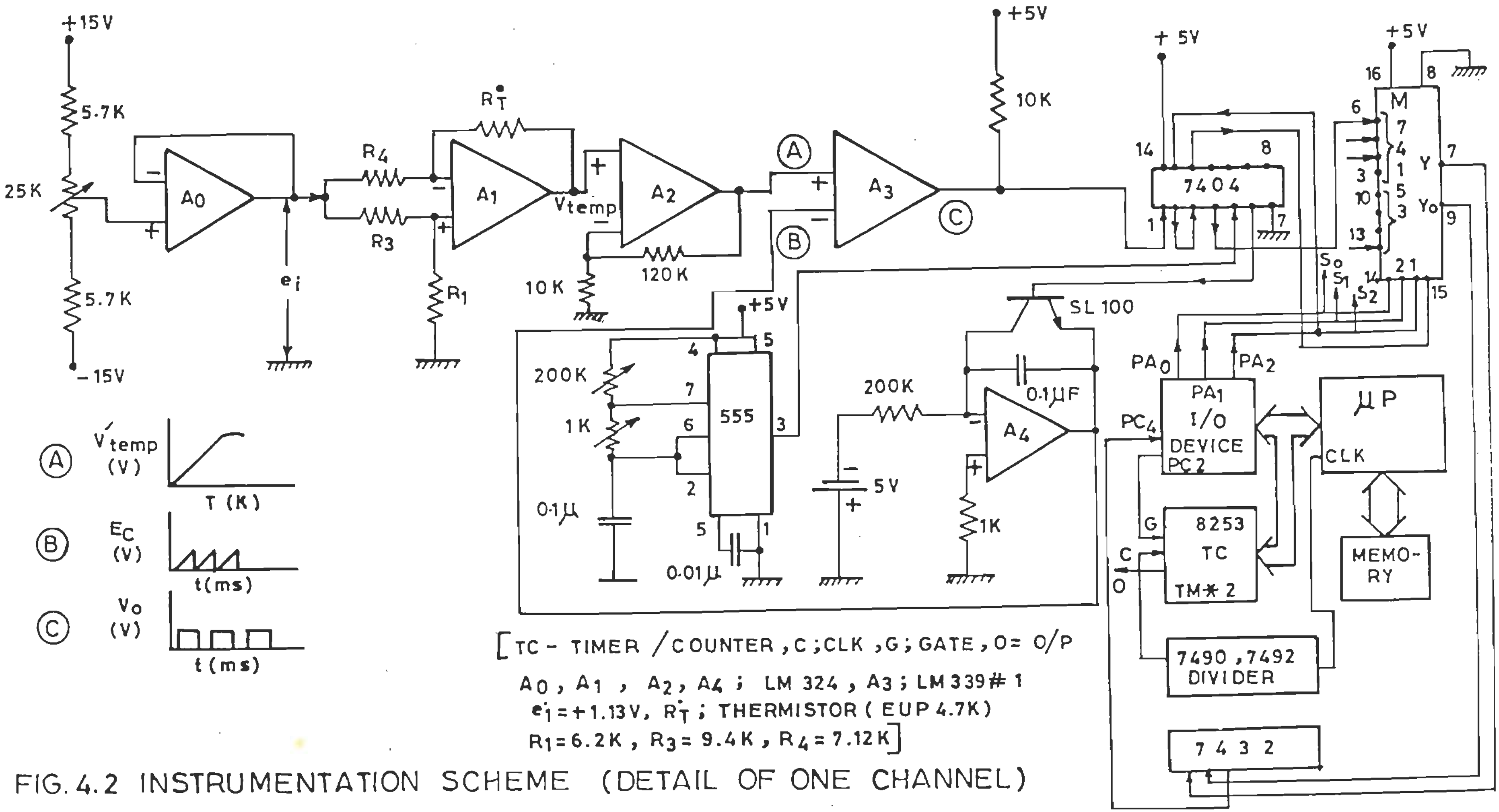
v) PWM works still when synchronization between transmitter and receiver fails.

## 4.4 MULTIPLEXING

The simultaneous transmission of signals without interference using a single channel from different data sources is accomplished by multiplexing. These are classified into two types- time-division multiplexing (TDM), and frequency-division multiplexing (FDM). TDM is an extension of pulse modulation. In TDM, use is made of the fact that narrow pulses with wide spaces between them are generated in pulse modulation systems. The wide space between pulses is then used by signal from other sources. Each of the signal is assumed to have been sampled at the Nyquist or at some higher rate. The samples are interleaved and a single composite signal consisting of all interleaved pulses is transmitted over the channel. At receiving end, the interleaved samples are separated by a synchronous switch or demultiplexer and each signal is reconstructed from the approximate set of samples. Proper operation of this system depends upon proper synchronization between transmitting and receiving end. If the message signals have equal bandwidths, the samples are then transmitted sequentially. If the sampled signals have unequal bandwidths, more samples must be transmitted per unit from the wideband channels. FDM concerns itself with combining continuous (or analog) signals. It is a technique whereby several message signals are translated using modulation to different spectral locations and added to form a baseband signal. The carriers used to form the baseband are usually referred to as subcarriers. Then if desired, the baseband signal are transmitted over a single channel using a single modulation process.

## 4.5 HARDWARE IMPLEMENTATION

In the present work, Pulse-width modulated technique with time-division multiplexing has been used for transmission of temperature data from eight different locations at a common point uwing a multiplexer in a microprocessor-based system as shown in Fig. 4.1. The circuitry has been developed for single channel and the details are shown in Fig.4.2. The positive dc supply ($e_i$) at constant potential of 1.13 volt is fed to a single-thermistor-active-bridge amplifier, $A_1$.The output $V_{temp}$ from this amplifier is linearly dependent upon temperature. This output is fed to, $A_2$, a non-inverting amplifier whose closed-loop gain is set at 12 and in turn its output is $V'_{temp}$. A saw tooth (voltage) wave (Ec) with constant amplitude and frequency (5V, 100 Hz) is obtained from a ramp-generator. Voltage comparator, $A_3$,(LM-339) compares two input voltages, Ec and $V'_{temp}$ as shown in Fig.4.3(a). Ec is connected to negative input terminal of comparator. $V'_{temp}$ is connected to positive input terminal of the comparator (LM-339#1) in non-inverting (pulse-width modulator) mode. The inputs are interchanged for inverting mode of operation. The output $V_o$ of the comparator is high (TTL level) and is in shape of rectangular positive going pulses of time interval $T_H$ (width of the pulse). The magnitude of $V_o$ remaining same for time interval whereas latter is varying in accordance with $V'_{temp}$. The output stays high for 0.98 msec when $V'_{temp}$ equals to 0.49 volt (at 303 K) and increases to 9.0 msec for corresponding increase




[TC - TIMER /COUNTER, C; CLK, G; GATE, O= O/P
$A_0$, $A_1$, $A_2$, $A_4$ ; LM 324, $A_3$; LM 339# 1
$e_1' = +1.13V$, $R_T^{\cdot}$ ; THERMISTOR ( EUP 4.7K)
$R_1 = 6.2K$, $R_3 = 9.4K$, $R_4 = 7.12K$]

FIG. 4.2 INSTRUMENTATION SCHEME (DETAIL OF ONE CHANNEL)

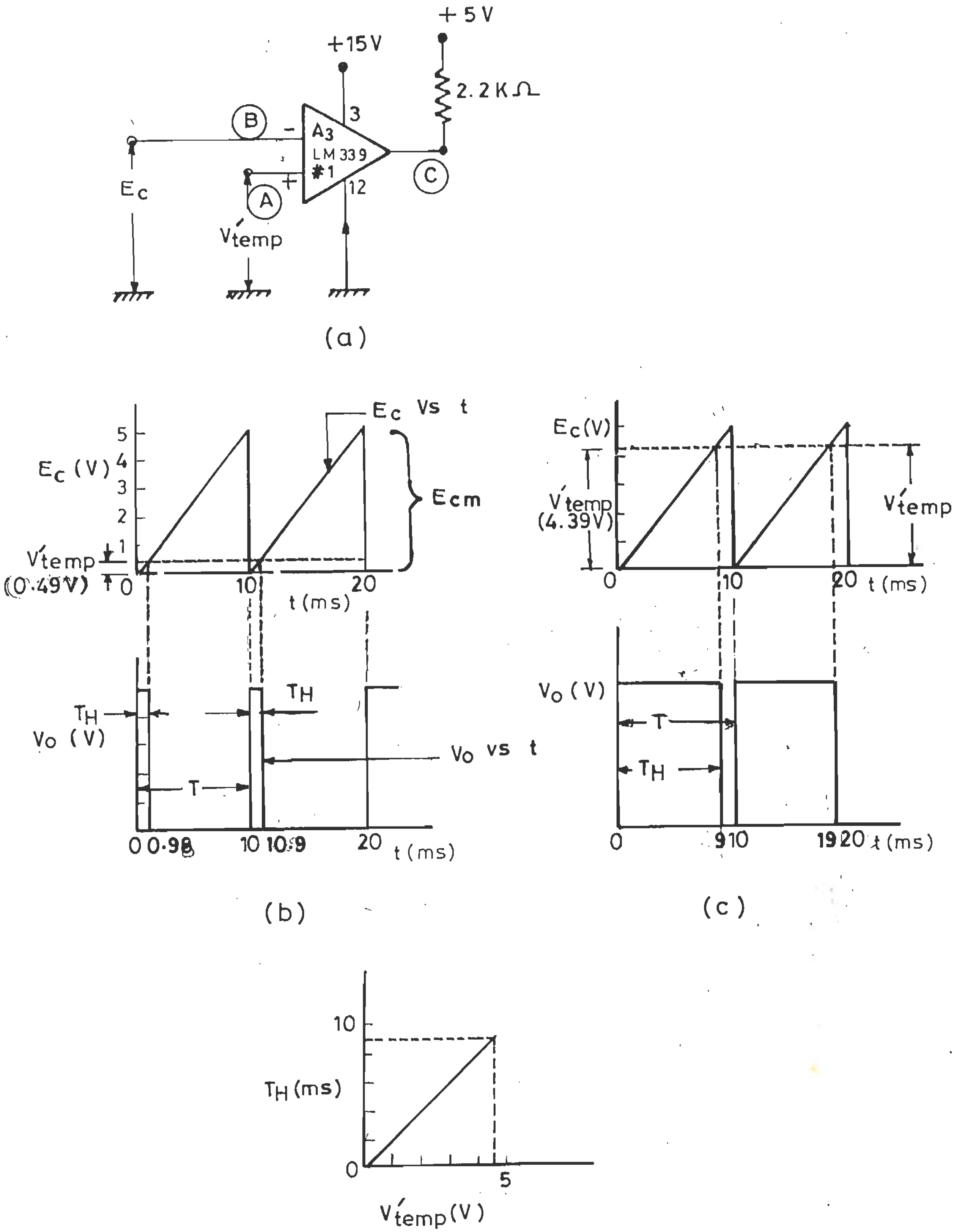



FIG. 4.3 (a) NON-INVERTING PULSE-WIDTH MODULATOR CIRCUIT
(b) INPUT AND OUTPUT WAVE FORMS FOR $V'_{temp}$= 0.49V
(c) INPUT AND OUTPUT WAVE FORMS FOR $V'_{temp}$= 4.36V
(d) OUTPUT HIGH TIME VS $V'_{temp}$

in $V'_{temp}$ to 4.38 volt (at 368 K), as shown in fig. 4.3 (b,c) respectively. Saw tooth wave is termed as carrier wave whereas temperature dependent voltage is message or information bearing signal for the telemetry scheme. The information bearing signal is impressed upon the carrier wave for modulation. The rate of change of $V'_{temp}$ is less than that of Ec to fulfil the sampling criteria. The width of output pulse is modulated by $V'_{temp}$. The constant period of output wave is set by Ec. Thus, Ec carries the information contained in $V'_{temp}$. $V_o$ is then said to be a pulse-width modulated wave. Operation of the non-inverting pulse-width modulator circuit is summarized by its input-output equation as

$$\text{Output } T_H = V'_{temp} \frac{T}{Ecm} \qquad \ldots (4.1)$$

where, T is the period and Ecm is maximum peak voltage of saw-tooth carrier. The width of the pulse at four specific temperatures is as

(i) For 298 K;

Output $T_H$ = 0.0 msec ( $V'_{temp}$ = 0.0 volt at reference temperature 298 K)

(ii) For 303 K;

$$\text{Output } T_H = \frac{0.49}{5} \times 10 \times 10^{-3} = 0.98 \text{ msec}$$

(iii) For 323 K;

$$\text{Output } T_H = \frac{2.5}{5} \times 10 \times 10^{-3} = 5 \text{ msec.}$$

(iv) For 368 K;

$$\text{Output } T_H = \frac{4.38}{5} \times 10 \times 10^{-3} = 8.76 \text{ msec}$$

$$= 9 \text{ msec (approximately)}$$

The slope of $T_H$ versus $V'_{temp}$ is linearly positive as shown in Fig. 4.3(d).

These (positive rectangular) pulses, in turn, forms one input to 8-channel multiplexer (IC-74153). But to avoid loading of the input pulses, these are passed through two gates of hex-inverter (IC-7404) and its output is fed to multiplexer. The selection of the desired data input is controlled by SELECT (or ADDRESS) inputs $S_0$, $S_1$, $S_2$ of multiplexer. Programmable peripheral interface (IC-8255 : 2) is used for interfacing multiplexer with microprocessor [32]. The latter sends control signal sequentially through bits $PA_0$, $PA_1$, $PA_2$ and depending upon the digital code applied at the select inputs, one out of 8 data source is selected and transferred to single output channel. Eight inputs to multiplexer are internally grouped as first four and next four inputs to provide their respective outputs at $Y_0$ and $Y_1$. In turn, the output from multiplexer is fed to one gate of 2-input - OR gate (IC - 7432) and finally this output is received at bit $PC_4$. To carry out desired counting of the pulse width and for interfacing, microprocessor (INTEL-8085) kit provides necessary facilities. Programmable timer/counter (IC-8253) has been used as counter for counting the number of pulses contained in pulse-width at a particular instance. The microprocessor checks the presence of input signal and converts the pulse-width to digital number by sending control signal through bit $PC_2$ at the gate of timer # 2. Clock signal to the timer is supplied via its clock input after dividing clock frequency (1.5 MHz) of the processor by 60 using external hardware circuitry around the ICs 7490 and 7492. The counter is used in down mode and its final value is complemented to get the actual count. The counting is started at the rising edge and stopped at the falling edge of the input pulse. The count values are obtained and are added for five successive pulses for each channel and then mean value is obtained to eliminate the chances of errors which could have resulted in considering only one pulse for counting. The mean count value is used to determine the address of locations in look-up table containing the vlaues of temperature. For more than one segment linearization, this process is repeated and separate look-up tables are used for every segment. Fig.4.4 shows the flow-chart of software developed for the complete process in the scheme. Prior to use the entries in the look-up tables are filled.

## 4.6 RESULTS AND DISCUSSIONS

The linear output of single thermistor active bridge amplifier is compared, after proper amplification, with a ramp wave of 5 volt and 100 Hz and the positive output pulses are produced varying from 0.0 msec to 9.0 msec (approx.) for a temperature range of 298 to 368 K. The width of each pulse

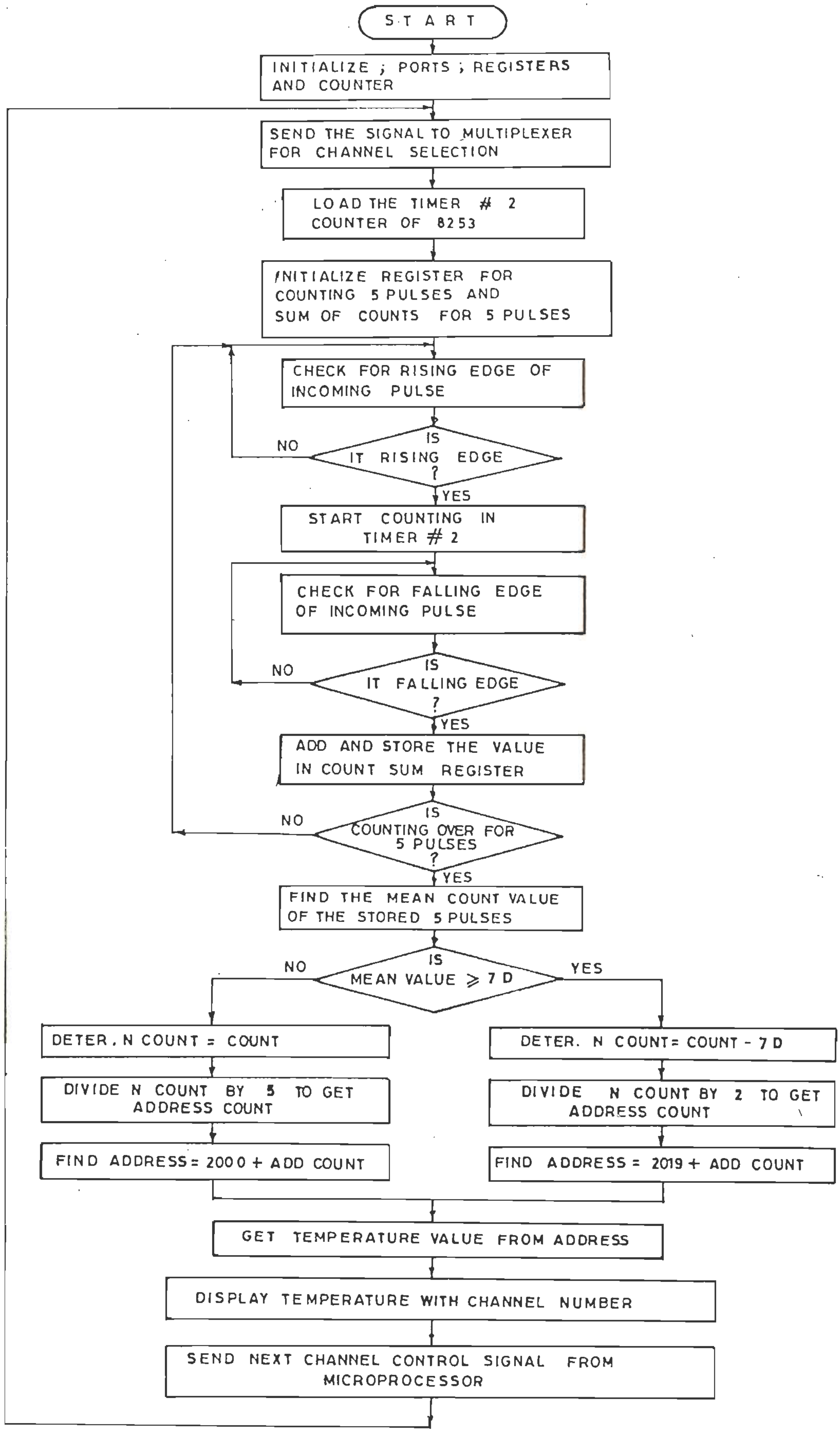



FIG. 4.4 FLOW CHART OF SOFT WARE DETAILS

(proportional to the temperature) is converted accordingly by microprocessor into mean counts for each temperature and results are shown in Table-I. First segment is from $(2000)_H$ to $(2019)_H$ and the second from $(2019)_H$ to $(2046)_H$.

Let the starting address for first look-up table is $(2000)_H$ and for the second $(2019)_H$. Then the address generated from 298 to 323 K and next from 323 to 368 K are as given below:

(i) From 298 to 323 K;

$$\text{Address} = 2000 + (\text{Count Value}) \div K_1$$

(ii) From 323 to 368 K;

$$\text{Address} = 2019 + (\text{Count Value} - 7D) \div K_2$$

where $K_1$ and $K_2$ are 5.0 and 2.0, respectively for adjusting the range of the scale. $V_{temp}$ is amplified by making use of non-inverting amplifier because this value is quite low to be compared in comparator.

## 4.7 CONCLUSIONS

The scheme has been developed for scanning and continuous monitoring of temperature in the range of 298 to 368 K from eight different locations in a sequential order. PWM technique has been used for transmitting the temperature data. Thermistor has been used as temperature sensor and non-linearity of response of the sensor has been corrected upto 80 per cent of full span. The scheme can be used for the monitoring and measurement of other physical parameters after appropriate modifications in the process of implementation of the scheme.

TABLE - I

| S. No. | Actual Temperature (K) | Count Value (Hex) | Address Generated | Temperature from Look-up Table (K) | |
|---|---|---|---|---|---|
| 1 | 298 | 00 | 2000 | 298 | Look-up Table for first segment |
| 2 | 303 | 19 | 2005 | 303 | |
| 3 | 308 | 32 | 200A | 308 | |
| 4 | 313 | 4B | 200F | 313 | |
| 5 | 318 | 64 | 2014 | 318 | |
| 6 | 323 | 7D | 2019 | 323 | |
| 7 | 328 | 87 | 201F | 328 | Look-up Table for first segment |
| 8 | 333 | 91 | 2023 | 333 | |
| 9 | 338 | 9B | 2028 | 338 | |
| 10 | 343 | A5 | 202D | 343 | |
| 11 | 348 | AF | 2032 | 348 | |
| 12 | 353 | B9 | 2037 | 353 | |
| 13 | 358 | C3 | 203C | 358 | |
| 14 | 363 | CD | 2041 | 363 | |
| 15 | 368 | D7 | 2046 | 368 | |

# CHAPTER - V

# DIGITAL TECHNIQUES FOR LINEARIZATION OF TRANSDUCER CHARACTERISTICS

## 5.1 INTRODUCTION

In general, transducers have nonlinear relationships between physical measurand and its electrical output through out the range of measurement. In fact, these relationships are either of exponentially rising or decaying nature, just like the charging and discharging of capacitors [78]. In some cases there may be deviations even from these characteristics. It has been observed that linear transducers often turn out to be less sensitive, more costly, and difficult to develop. But from application point of view, the transducer response is more useful when it is linear. In some of the cases, a transducer is used to operate over the linear portion of its transfer curve and the need for linearization is avoided. This also results in a compromise with sensitivity and linearity.

The linearity is achieved by various types of hardware and software techniques. In the hardware approach, it is achieved with the aid of electronic circuits which may differ from transducer to transducer and at time may not be applicable even to same trnansducer in its entire range of measurement. With the advent of low cost LSI technology, these days almost all data acquisition systemsand data loggers are based around microprocessors and microcomputers. In the software approach, the programming and computational capabilities of microcomputer are utilized for obtaining linear relationship between input and output characteristics of transducers. These techniques are of generalized nature and a particular software approach can be used to almost all types of response characteristics by little modification at the stage of its implementation.

## 5.2 EXISTING TECHNIQUES

Number of methods have been developed for linearization of undesired but predictable nonlinear transducer response characteristics. These are digital techniques and are used to obtain corrected, scaled and useful data in discrete form. There are some digital linearizing techniques which use the error curve (difference between nonlinear and linear responses) for response linearization [57,68]. The correction is made either by using integrated logic circuits or by straight line polygon. Epochwise, in one technique, Kollataj, et al, [57] implemented this technique by using a logic-circuit card which carries out the correction for requisite linearization. With the help of a programme, the correction factor is applied for each type of characteristics. Mayer, et al, [68] developed a technique which is implemented using software. In this approach the error curve is approximated by straight line polygon. By using these techniques, researchers obtained linear input-output characteristics for some transducers (thermocouples, RTDs, and flow measuring sensors). These techniques are difficult to programme and require transfer function of transducer to be known explicitly before linearization. Some digital techniques use read-only-memory (ROM) and the computational algorithm is implemented on microcomputer-based instrumentation system [14,58]. In one of the approaches, linear interpolation is carried out by making use of three calibration points. In this, each transducer requires its own look-up table with provisions to account for scaling errors, zero offset, and manufacturing tolerances of the sensor [10]. There is one other technique which is analytic in nature and provides correction for gain-nonlinearity and disturbing variables [66].

None of the existing techniques provides a general and convenient method of calibration and linearization for the transducer characteristics falling into exponentially decaying and rising categories. In the present work, a general digital technique for transduer response linearization is developed which can be used for the both type of characteristics.

## 5.3 LINEARIZING TECHNIQUE

This section presents a new digital technique for transducer response linearization using scale factor polynomial. The method is of generalized nature and is applicable to response characteristics of various types. The technique is applicable to both single segment and piece-wise-multi-segment linearization. In this approach, the non-linear response curve of the transducer is used to generate requisite data for linearization. First of all a straight line is fitted between two points L and H which represent low and high values of measurand as shown in Figs.5.1(a) and 5.2 (a). It can also be fitted between any desired points on the curve. The equidistant sampled points corresponding to different values of measurand on the curves are shifted on to fitted straight line. The respective ordinate values are calculated. Then the scale factors ($K_1, K_2, K_3, \ldots, K_n$) corresponding to abcissa ($x_1$, $x_2$, $x_3$, ..., $x_n$) are calculated to shift non-linear values to linear value. The set of values for scale factor function can be represented as a function of x, that is K = f(x). Now a scale factor polynomial function is obtained by making use of the scale factor points as

$$K = p_0 + p_1x + p_2x^2 + p_3x^3 + \ldots + p_nx^n \qquad \ldots (5.1)$$

where $p_0, p_1, p_2, \ldots, p_n$ are coefficients of polynomial. Least sqaures method [24] has been used for fitting the best possible curve. The limit of acceptable error is defined for the curve fitting. The polynomial is the equation of reference curve for gain-linearization. Later on, for a particular value of measurand, the scale factor is calculated using the polynomial and the value is obtained on linear response curve by multiplying the actual value with the scale factor.

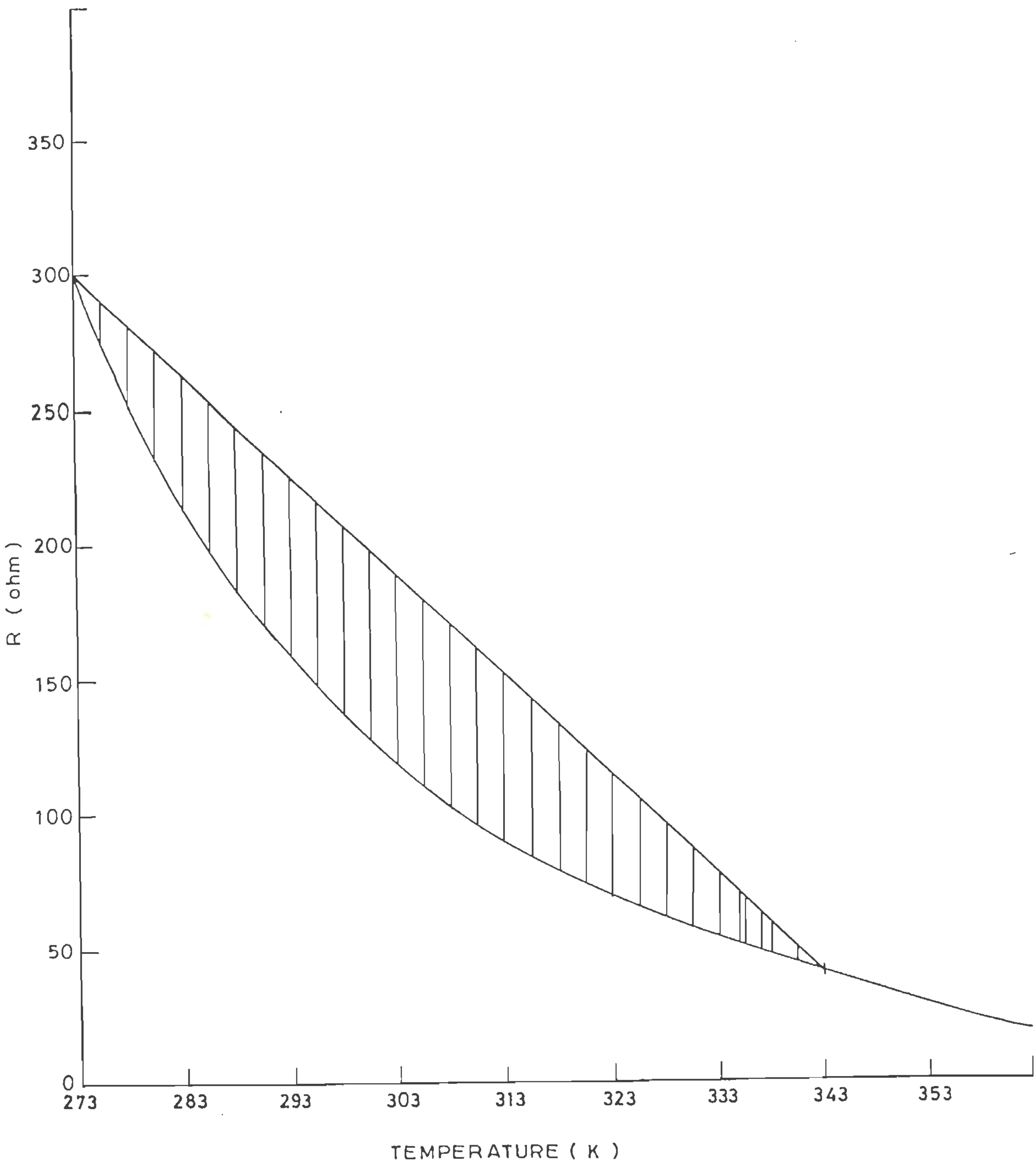



FIG. 5.1 (a) SINGLE SEGMENT LINEARIZATION FOR NTC THERMISTOR OUTPUT RESPONSE

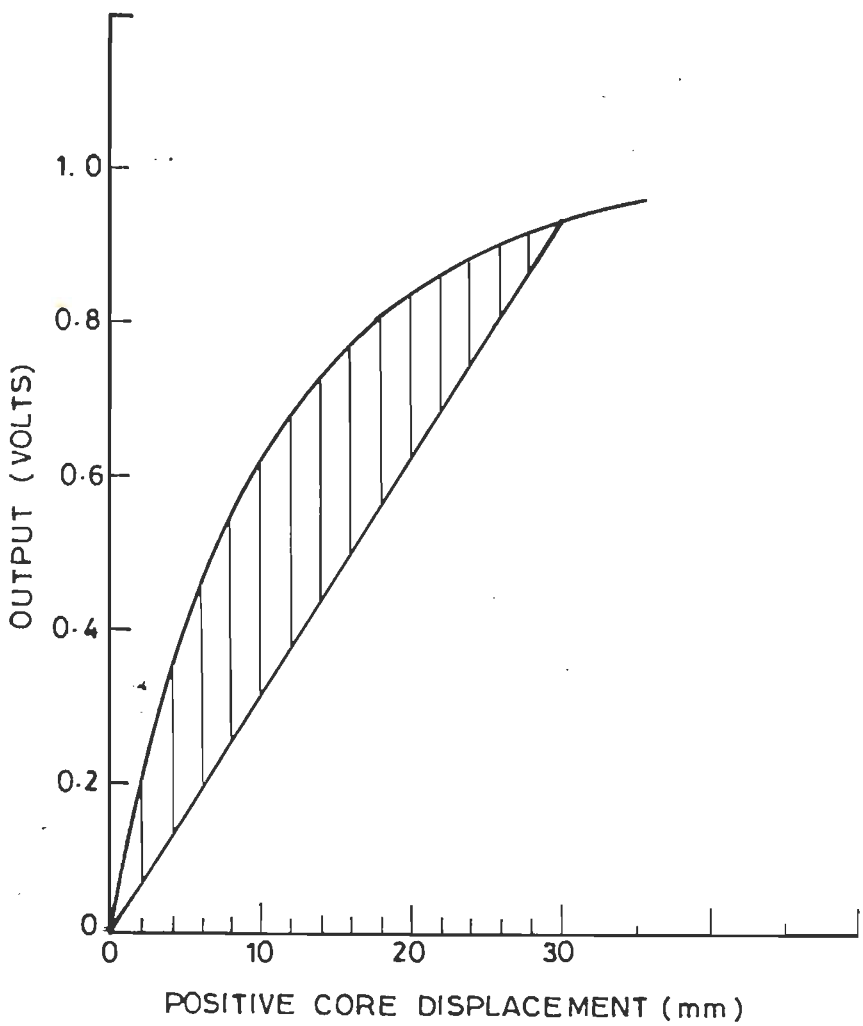



FIG. 5.2 (a) SINGLE SEGMENT LINEARIZATION FOR LVDT OUTPUT RESPONSE

The coefficients and order of best-fit polynomial are stored in memory.

Steps to determine the Scale-factor polynomial:

(i) Obtain the non-linear response curve of the transducer.

(ii) Fit a straight line curve between maximum and minimum values on the response curve.

(iii) Select equidistant points, $x_1, x_2, x_3, \ldots, x_n$ on the x-axis as shown in Figs.5.1(a) and 5.2(a)

(iv) Note the corresponding points (values) on non-linear curve $(y_1, y_2, y_3, \ldots, y_n)$ and on linear curve $(Z_1, Z_2, Z_3, \ldots, Z_n)$.

(v) Calculate values of the scale factors (at each point)

$$K_1 = \frac{Z_1}{y_1},\ K_2 = \frac{Z_2}{y_2},\ K_3 = \frac{Z_3}{y_3},\ \ldots\ ,\ K_n = \frac{Z_n}{y_n} \qquad \ldots (5.2)$$

(vi) Using method of least squares, determine the best fit (curve) polynomial for the scale factor given by equation (5.1).

(vii) Store the value of coefficients $p_0, p_1, p_2, p_3, \ldots, p_n$ for further calculations.

(viii) Now for a particular value of measurand x, the value of K is determined by the polynomial. The actual value is multiplied by K to get the linear value.

For highly nonlinear response characteristic, (at the extreme ends), as in case of NTC thermistor, it is not possible to stick to the criteria of equidistant sampling points. In such case the sampling points (probably at the ends) are selected at different intervals (other than equidistant) to increase the number of sampling points for obtaining a desired order of polynomials.

Application of piece-wise linearization approach on the response curves is shown in Figs. 5.1(b) and 5.2(b). In this case, number of segments are fitted in entire range of response curve and each individual segment is treated as a single segment. The number of segments recommended usually are between 5 and 7 [68].

## 5.4 EXPERIMENTAL RESULTS AND CONCLUSIONS

The efficacy of the proposed scale-factor polynomial technique has been investigated off-line on a main frame computer DEC-2050 for the linearization of response characteristic of NTC thermistor (a case of exponentially decaying response) and LVDT transducer characteristic (a case of exponentially rising response). The results of linearization are obtained for both by employing single segment and piece-wise linearization techniques. The results are given in Table-I and II. It is evidently clear from the Tables that the errors are much larger in case of single segment linearization compared to piece-wise linearization. In piece-wise

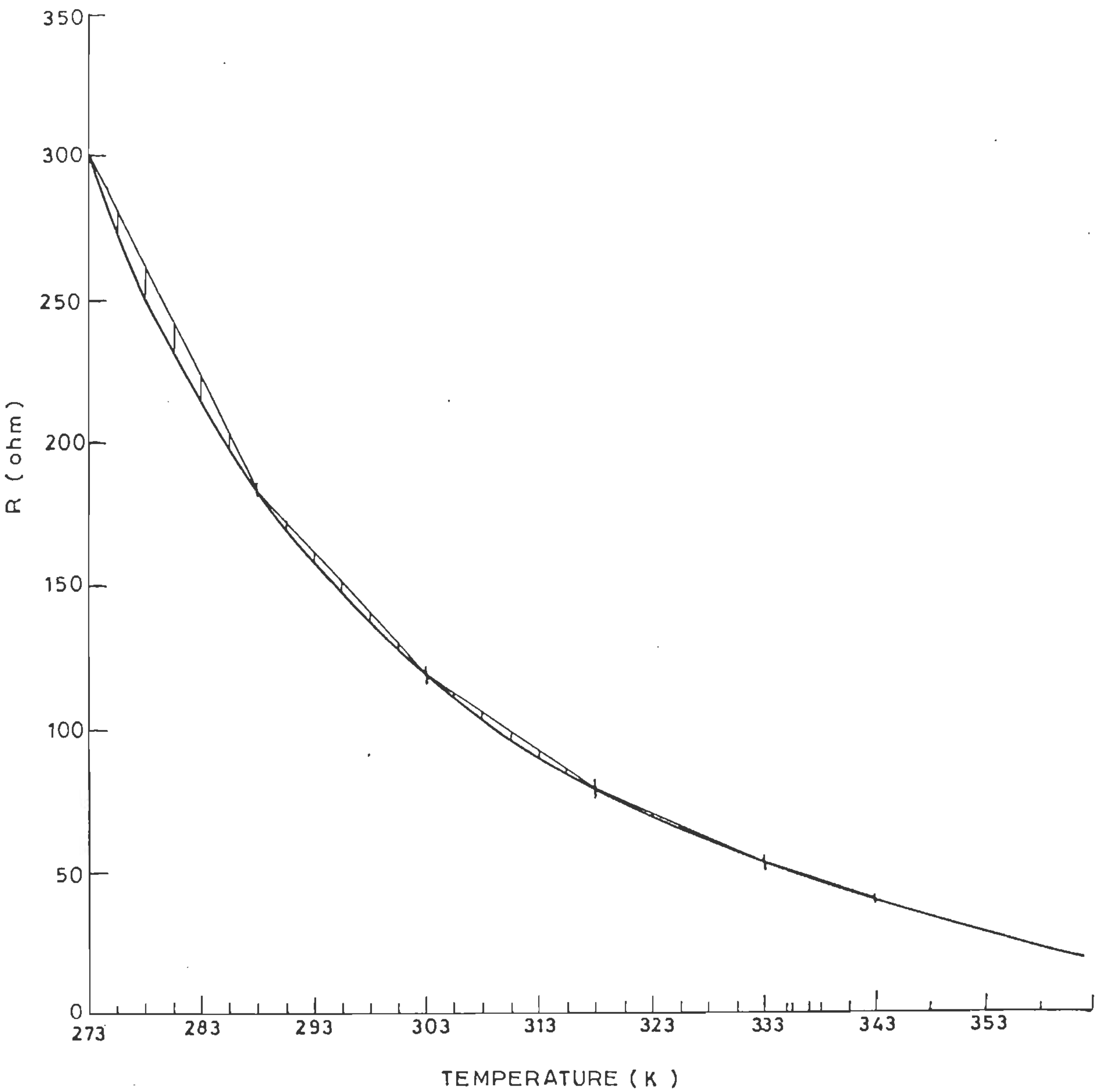



FIG. 5·1 (b) PIECE-WISE SEGMENT LINEARIZATION FOR NTC THERMISTER OUTPUT RESPONSE

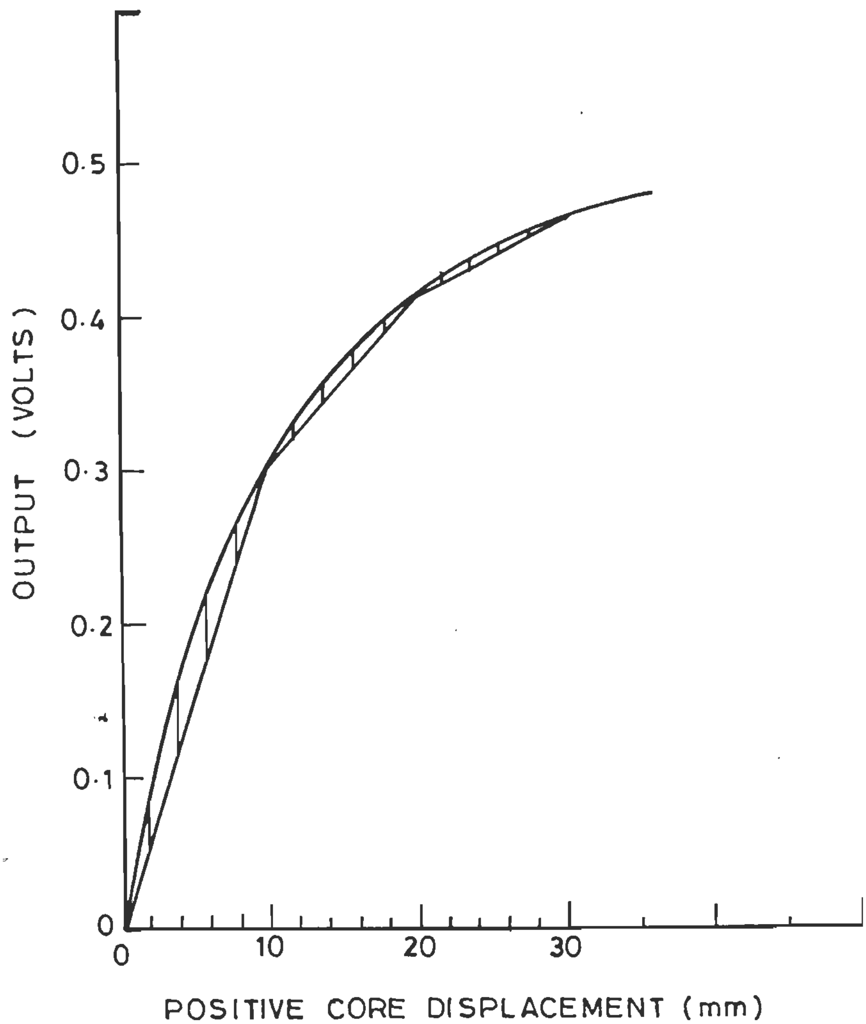



FIG. 5. 2 (b) PIECE-WISE SEGMENT LINEARIZATION FOR LVDT OUTPUT RESPONSE

TABLE - I

(For NTC Thermistor)

| T (K) | Order of Polynomial | | | |
|---|---|---|---|---|
| | Single segment | | Piece-wise | |
| | 7th | 5th | 4th | 3rd |
| 275.5 | 275.666 | 275.237 | 275.500 | 275.498 |
| 278.0 | 277.795 | 277.997 | 278.000 | 278.006 |
| 280.5 | 280.052 | 280.394 | 280.500 | 280.491 |
| 283.0 | 283.381 | 283.623 | 283.000 | 283.005 |
| 285.5 | 286.274 | 286.346 | 285.500 | 285.500 |
| 288.0 | 286.49 | 286.806 | | |
| 290.5 | 291.104 | 290.927 | 290.508 | 290.546 |
| 293.0 | 292.849 | 292.642 | 292.984 | 292.825 |
| 295.5 | 295.792 | 295.618 | 295.502 | 295.742 |
| 298.0 | 297.536 | 297.429 | 298.008 | 297.847 |
| 300.5 | 300.553 | 300.531 | 300.496 | 300.535 |
| 303.0 | 302.912 | 302.867 | | 2nd |
| 305.5 | 305.466 | 305.577 | 305.514 | 305.474 |
| 308.0 | 308.169 | 308.304 | 307.961 | 308.034 |
| 310.5 | 310.719 | 310.847 | 310.534 | 310.538 |
| 313.0 | 313.033 | 313.176 | 312.990 | 312.923 |
| 315.5 | 315.681 | 315.721 | 315.5 | 315.530 |
| 318.0 | 317.580 | 317.561 | | 2nd |
| 320.5 | 320.559 | 320.489 | 319.873 | 320.554 |
| 323.0 | 322.787 | 322.685 | 324.029 | 322.832 |
| 325.5 | 325.576 | 325.472 | 325.753 | 325.68 |
| 328.0 | 328.026 | 327.949 | 326.890 | 327.921 |
| 330.5 | 330.804 | 330.778 | 330.954 | 330.510 |
| 333.0 | 333.000 | 332.811 | | 3rd |
| 335.0 | 335.006 | 335.071 | 335.602 | 335.001 |
| 335.5 | 335.478 | 335.547 | 335.473 | 335.476 |
| 337.0 | 336.890 | 337.058 | 337.078 | 337.078 |
| 338.0 | 337.497 | 337.987 | 337.938 | 337.608 |
| 340.5 | 340.520 | 340.41 | 340.507 | 340.509 |

TABLE - II
FOR LVDT Transducer

| d mm | Order of Polynomial | | | |
|---|---|---|---|---|
| | Single Segment | | Piece-wise | |
| | 7th | 5th | 3rd | 2nd |
| 2 | 2.026 | 2.077 | 2.000 | 2.016 |
| 4 | 3.898 | 3.759 | 4.000 | 3.921 |
| 6 | 5.939 | 5.884 | 6.000 | 6.110 |
| 8 | 8.498 | 8.626 | 8.000 | 7.956 |
| 10 | 9.750 | 9.923 | | |
| 12 | 11.503 | 11.573 | 12.001 | 11.977 |
| 14 | 14.141 | 14.052 | 13.997 | 14.076 |
| 16 | 16.251 | 16.073 | 16.003 | 15.919 |
| 18 | 18.064 | 17.942 | 18.000 | 18.028 |
| 20 | 19.916 | 19.948 | | |
| 22 | 21.928 | 22.080 | 22.030 | 22.111 |
| 24 | 23.940 | 24.036 | 23.912 | 23.658 |
| 26 | 26.118 | 26.016 | 26.087 | 27.305 |
| 28 | 27.963 | 27.962 | 28.924 | 28.833 |

linearization, the order of polynomial is reduced and also differ from segment to segment. Thus piece-wise linearization produces better and encouraging results. It has been found that a trade-off has to be made in the order of polynomial, number of segments, and the degree of linearity. The linearity is obtained within ±0.1 per cent of full scale value using piece-wise linearization. In this method, the number of data points should be more than the order of polynomial. The technique is simple, general and convenient and can be implemented on any micro computer based data acquisition system. The results are satisfactory and in conformity with the results obtained by other existing digital techniques.

# CHAPTER - VI

## CONCLUSIONS AND FUTURE SCOPE OF WORK

### 6.1 AUTHOR'S CONTRIBUTIONS AND CONCLUSIONS

This chapter deals with the contributions made by the author to improve the performance of three types of inductive transducers and one type of thermistor (NTC). The important conclusions drawn during the course of investigations of this work are summarized below:

(i) The author has developed three self-compensated 'smart' indcutive transducers, namely linear variable differential transformer (LVDT) transducer, inductive ratio transducer (IRT), and inductive differential transducer (IDT) for the measurement of linear displacement in different types of environments including hostile conditions. The response characteristics of compensated transducers are linear in reasonable span and their overall performances have improved considerably compared to the uncompensated type of inductive transducers.

In first case by making use of dual sets of secondary windings a self-compensated LVDT transducer has been developed which has immunity to changes in input parameters and to variations of ambient temperature in and around the transducer assembly. This technique makes differential transformer transducer 'smart' enough to provide self-compensation without employing any additional component or reference LVDT. The dual sets winding arrangement has caused only marginal increase in the dimension of secondaries. The circuitry of the transducer does not put any additional constraints on the processing and conditioning of signals.

(ii) The other two self-compensated transducers are inductive ratio and inductive differential types, which have been developed by making use of two identical (matched) coils for precise measurement of linear displacement. Both IRT and IDT have been tested for the effect of variations of input excitation parameters and changes in ambient temperature. The response of IRT is not influenced by these variations. The response of IDT is not influenced by the variation of ambient temperature but there is a necessity of maintaining constant input excitation voltage and frequency for this transducer.

These three inductive transducers are quite stable in operation showing consistency in results with hysteresis free response in favourable as well as hostile environments. In addition to measurement of linear displacement, these transducers can also be used for measurement of other parameters like pressure, force, flow, etc. These are suitable for both indoor and outdoor applications due to high stability in performances.

(iii) In addition to improvement of performances of three inductive transducers, the work has been carried out for gain-linearization of NTC thermistors. Two analog schemes have been developed for response linearization of the thermistor. These schemes have been successfully tested for a temperature range from 298 to 368 K. Mathematical analysis has been carried out for both these schemes. The circuits have been built by using easily available and off-the-shelf electronic components at a moderate cost. In one of the schemes, thermistor has been placed in feed-back path of an active bridge amplifier and optimal condition for gain-linearity has been obtained. The condition is independent of the thermistor specifications. The input-output relationship of this developed bridge is linear over 80 per cent of full span. The second gain-linearization scheme is devised by employing log amplifier in conjunction with (monolithic) analog divider unit. Converted variable (temperature) is inverted by the divider unit and linearity of response is achieved. Transfer characteristic of this circuit obeys straight line equation.

(iv) By making use of linearized output of single thermistor-active bridge amplifier, a multichannel continuous temperature scanning and telemetry scheme using PWM technique around microprocessor-based system has been successfully developed and tested. A fast and efficient algorithm has been developed to select one out of eight channels of multiplexer sequentially and then to convert the width of the pulse (proportional to input temperature) into equivalent count. This scheme can be used for scanning and continuous monitoring of other industrial parameters with appropriate modifications in the software.

(v) The author has developed a general scale-factor digital technique for gain-linearization of transducers exhibiting response characteristics either of exponentially decaying or exponentially rising category. The validity of the technique has been tested off-line on main frame computer for response linearization of NTC thermistor falling into the former category and for LVDT transducer falling into the latter category. The technique has been applied to single segment (stroke) as well to piece-wise (multi segments) linearization. The order of polynomial is reduced in the piece-wise linearization approach and linearity (for both the transducers) is achieved within ±0.1 per cent of full span.

## 6.2 SUGGESTIONS FOR FUTURE WORK

The following suggestions are made for carrying out future work in this field:

(i) The performance of the inductive transducers can be studied under variations of other environmental parameters like humidity or pressure and by using core of different (materials) compositions and dimensions. The efforts can be made to make these transducers self-compensating type under these variations.

(ii) The work on similar lines can be carried out for development of smart and compensated capacitive, resistive, and other transducers.

(iii) In the scheme developed for gain-linearization of thermistor using log amplifier, the performance of the amplifier can be further improved in light of its low gain and temperature sensitivity due to $\frac{kTo}{q}$ factor. These limitations can be overcome by using modified log amplifier circuit [107] for increasing the gain and also providing temperature compensation. This improved log amplifier can be used for precise and linear temperature measurement.

(iv) In our laboratory, the facility was available to vary the temperature from 298 to 373 K to carry out the study for influence of temperature variation on the performance of transducers. Further the performance investigations can be carried out below 273 K and above 373 K on these transducers.

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# LIST OF PUBLICATIONS

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6. Microprocessor - Based Efficient Multipoint Temperature Scanning Telemetry System Using PWM Technique, 13th National System Conference (NSC-89), Dec. 13-15, 1989, I.I.T.,Kharagpur (W.B.).