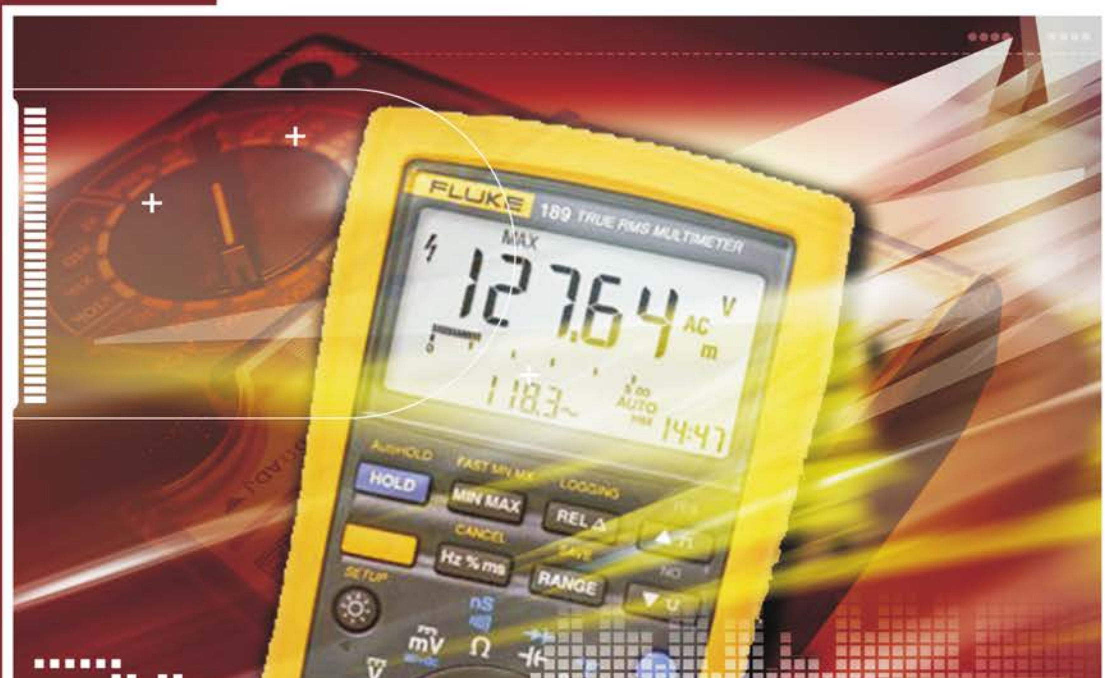
Electronic Measurements and Instrumentation



K. Lal Kishore

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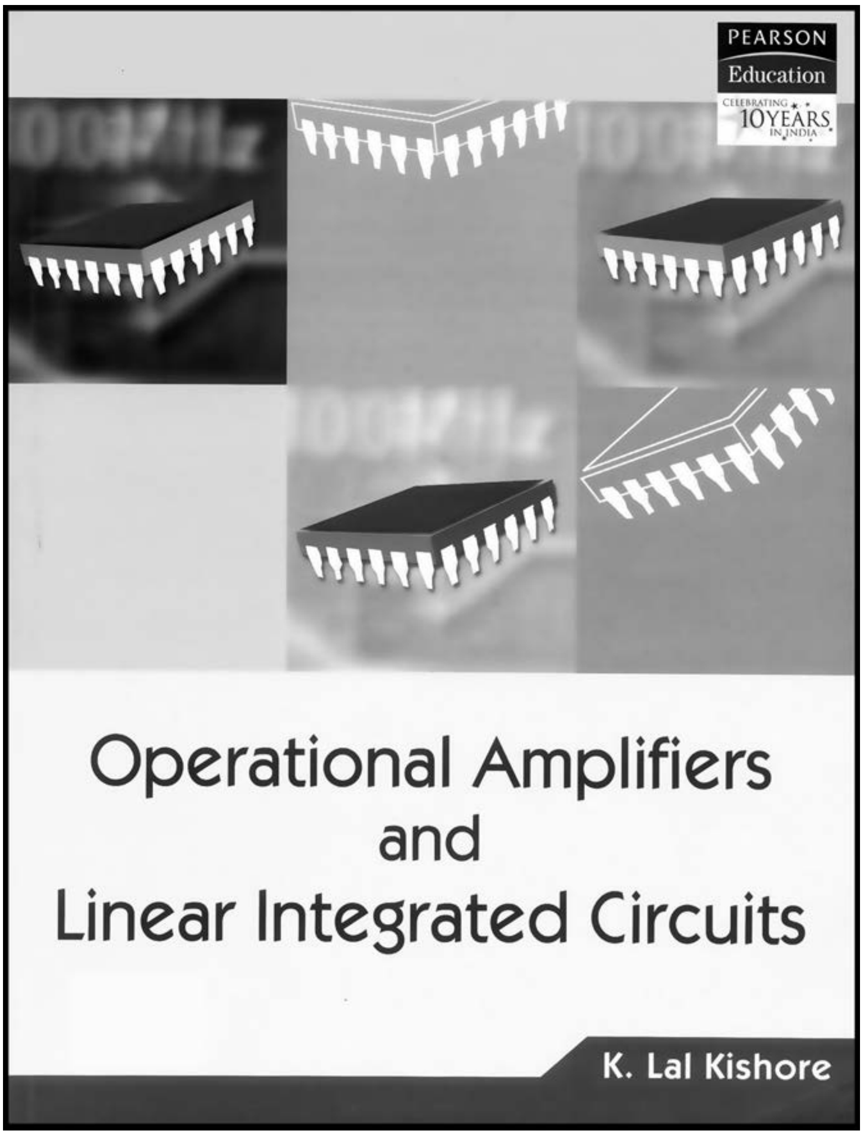
Operational Amplifiers and Linear Integrated Circuits is a textbook that caters to a complete course on the subject, scroes with a detailed and well-written treatment of designing circuits using op-amps, and their practical applications in the industry.



Electronic Measurements and Instrumentation

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Dedicated

to

The Almighty, who is omnipresent and omnipotent

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Foreword

It gives me immense pleasure to write a foreword for Electronic Measurements and Instrumentation. The author has extensive experience in teaching this subject to students of B.Tech and Electronics and Communication Engineering and has based this book on his insights into the training needs of those who have opted for this field of study.

Technical education is expanding rapidly and every year several new engineering colleges are coming up, particularly in the state of Andhra Pradesh. With an increased number of students choosing engineering education there is an urgent need for good textbooks in this discipline. As there is a dearth of well-written books which meet the requirement of subject coverage from the perspective of Indian students, this book by Dr Lal Kishore is an important contribution in this area. Written in a lucid style with a rich repository of pedagogical features, this book will be useful for undergraduate students of electronics and communication engineering, instrumentation engineering, and allied branches. This book will also be useful to AMIETE and AMIE (Electronics) students. The author has incorporated several illuminating features in this book, and these include:

1. Chapter outlines
2. Summaries at the end of each chapter
3. Points to remember covering important terms
4. Objective-type questions
5. Review questions
6. Solved and unsolved problems for each chapter

I am sure that this textbook will be beneficial to students and teachers as well.

Prof. D. N. Reddy
Vice Chancellor
Jawaharlal Nehru Technological
University Hyderabad, Hyderabad

Preface

The role of science is to discover the laws of nature, while the role of engineering and technology is to invent methods to apply these laws to serve human needs. The development of science began at a rapid pace in the 19th century, and it continues unabated to this day. Measurement plays an important role in the context of the progressive metamorphosis of science and technology. The subject of measurement is crucial to advancements in the field of electronics. It is interesting to note that time, as a parameter first attempted to be measured by primitive races, is the most accurate that can be studied by contemporary technology.

This textbook has been written against a backdrop of the imperative need to closely examine the science of measurement vis-à-vis its emergence as a discipline indispensable to the study of electronics and related fields. Though a number of textbooks have been written on the topic, a need was felt to address the typical difficulties faced by the student in learning the subject and to clearly explain the subtle points and principles involved. Written in a user-friendly manner, the book engages both the student and the teacher with its lucid style and focussed pedagogical features that include summary, points to remember, objective-type questions, review questions, and unsolved problems. The subject has been discussed with the undergraduate student in mind, and all topics generally covered at the undergraduate level are examined in this book. The reader is acquainted with specifications, parameters, and the typical values obtained during usage of the different types of instruments. The subject can be understood well when related numerical problems are solved. To this end, a number of problems are worked out in all chapters.

The textbook will be useful to students appearing for competitive examinations as well, since it includes the topics covered in a number of universities. It would also be beneficial to students of AMIETE, AMIE, M.Sc. (Electronics), and all programmes that include electronic measurements and instrumentation as a subject of study.

Suggestions and feedback to improve this book are welcome.

Acknowledgements

I am thankful to Dr. E. N. Ganesh, Assistant Professor, Department of Electrical Communication Engineering (ECE), BSA Crescent Engineering College, for his help in writing this textbook. I am also grateful to E. V. L. N. Rangacharyulu, Associate Professor, Department of ECE, Sri Indu College of Engineering and Technology, Hyderabad, for his inputs. A number

of textbooks, journals, instrument manuals, and data manuals have been referred to at the time of writing this book. Thanks are due to the authors of all these textbooks and research articles. I am obliged to M/s Tektronix Instruments, M/s Keithley Instruments, and M/s Peridot Technologies, Hyderabad, for providing technical information on various instruments.

K. Lal Kishore

About the Author

Professor K. Lal Kishore is a distinguished academician with over 30 years of teaching experience. He has several awards to his credit including one from Defence Engineering College, Ethiopia, for distinguished service and another from International Compendium for distinction in academics. A post-graduate and a Ph.D. degree-holder from the Indian Institute of Science Bangalore, Professor Kishore began his career in teaching at Jawaharlal Nehru Technological University (JNTU) Hyderabad in 1977. Appointed as professor of electrical communication engineering (ECE) in 1990, he has subsequently held many academic and administrative positions at JNTU Hyderabad including that of Chairman BOS; Head, Department of ECE; Principal; Academic; and Planning Director; UGC Academic Staff College; and Registrar. Currently he is Rector at JNTU Hyderabad.

With more than 76 research papers in national and international journals to his credit, Professor Kishore has presented several papers at national and international conferences and guided a number of research scholars. He was presented the Best Teacher Award for the year 2004 by the government of Andhra Pradesh. He has also received the S. V. C. Aiya Memorial Award and the Bapu Seetharam Memorial Award from the Institution of Electronics and Telecommunication Engineers (IETE).

A dedicated researcher in the field of electronics and telecommunications engineering, Professor Kishore is a member of the Institute of Electrical and Electronics Engineering (IEEE), Fellow of the Institute of Electronics and Telecommunication Engineers (IETE), and a lifetime member of the Indian Society of Technical Education (ISTE) and the International Society of Hybrid Microelectronics (ISHM).

1

Measurements and Instruments

Introduction • Terminology • Performance characteristics • Significant figures • Dynamic characteristics • Types of errors • Statistical analysis • Measurement standards • Suspension galvanometer • D'Arsonval movement • Direct current meter • D'Arsonval meter movement used in DC voltmeters • DC voltmeter • Ohmmeter • Multimeter • Alternating current-indicating instruments • Rectifier-type instruments • Meter protection • Extension of range • Frequency compensation • Electronic voltmeter (for DC) • Electronic voltmeter (for AC) • DC meter with amplifier • Chopper-stabilised amplifier • AC voltmeter using rectifiers • True *RMS*-responding voltmeter • Balanced bridge voltmeter (VTVM) • Transistor voltmeter (TVM) • Electronic multimeter • AC current measurement • Differential amplifier • Alternating current instruments (AC meters) • Electrodynamometer movement • Thermocouple meter • Digital voltmeter • Ramp-type DVM • Staircase ramp-type DVM • Dual slope integrating-type DVM • Successive-approximation conversion (SAC) • Continuous balance-type DVM • Automatic polarity indication for DVM • Autoranging for DVM • $3\frac{3}{4}$ Digit display • Picoammeter • Low-current ammeter applications • High-resistance measurements • Summary

1.1 INTRODUCTION

Although the measurement of simple physical parameters like time dates back to ancient times, measurement as a precise technology is only a few hundred years old. Perhaps, *time* was the first parameter measured by man. In olden days, time was measured during the day time by the position of sun in the sky. To count days or years, stones were placed at a particular position. Years were counted in the same way, by placing stones whenever a particular season arrived. The age of a person was also counted similarly. On the death of a person, the stones accumulated on his/her account were buried with his/her body. Though it is a crude technique, it was a kind of *measurement*. This method is still being followed as a ritual by some primitive races in Africa. Ironically, the same parameter '*time*' is that which can be measured most precisely compared to other parameters. Cesium–Hydrogen atomic clocks have an accuracy of 1 sec in 30,000 years. When Galileo found the simple pendulum, clocks were built based on this principle. No other physical quantity can yet be measured with such an accuracy as *time*.

The first major contribution to scientific instrumentation is from the science of optics. The telescope, the microscope, and the spectroscope were the first categories of scientific instruments. The spectroscope revealed new elements on Earth. The microscope showed that the cellular structure of living matter and microorganisms is the effective cause of diseases.

The role of *science* is to discover the laws of nature and how they operate in complex systems. The role of *engineering* is to apply the discoveries of science to human needs. Scientists make discoveries that increase our understanding of the world. Engineers make inventions intended to increase our productivity or ability to survive. Instrumentation is a branch of engineering that serves not only science but also all branches of engineering and medicine. The precise measurement of dimensions like temperature, pressure, power, voltage, current, impedance, etc., are as important to engineering as to science.

Accurate measurement is needed for economic design. A bridge several times stronger than needed to carry its heaviest possible load serves no one better and costs more. For millions of people watching TV, the dramatic moment of the Apollo-11 mission occurred when Neil Armstrong first set foot on the moon. But for many of the engineers who designed the lunar module, the dramatic moment occurred 2 hr earlier, when the lunar landing module set its feet on the moon. At that moment only 10 sec worth of fuel remained. Such a critical design was required since every gram of fuel saved could be used to increase the pay load of the lunar escape module.

The science of instrumentation plays an important role in technology. But electronics engineering plays a vital role in the field of instrumentation. The reason is that most physical quantities can be converted by transducers into electrical signals and once an electrical signal is available, the signal can be amplified, filtered, multiplexed, sampled, and measured. Nowadays, it is also possible to convert the signal into a digital form, interface it with a microprocessor (μP), or a microcontroller (μC), or a personal computer (PC) and can be monitored, controlled, and displayed effectively. The rapid strides made in semiconductor device technology and integrated circuits (IC) fabrication had its bearing on instrumentation also. The old instruments and systems using vacuum tubes were replaced by semiconductor devices and ICs. Now μP, μC, and PC-based measuring instruments and systems are the order of the day. Making use of the concept of composite materials, Thick-film Technology, Thin-film Technology, Hybrid Microcircuits, and new types of sensors that are *smaller in size and lesser in weight* are being developed. By embedding a processor, memory, and digital display, *intelligent sensors* are being developed.

Thus, electronic measurements and instrumentation is a rapidly changing subject and contributes significantly to science and technology.

In this chapter, the construction and principle of the working of electronic instruments are explained. The measurement of voltage, current, and resistance is given with figures and circuits. In addition, extension of range of the instruments and frequency compensation techniques have been discussed.

Due to mechanical movement, electromechanical type of instruments for the measurement of voltage, current, and resistance have inherent disadvantages. In this chapter, electronic circuits used for the measurement of these parameters are described. The student is expected to understand the working of circuits and the advantages of these instruments over electromechanical types. Average reading, peak reading, true rms reading voltmeters, principle of operation, and working of these meters as well as electronic AC voltmeters, balanced bridge voltmeter circuits, operation, and working are to be learnt, from this chapter.

1.2 TERMINOLOGY

This subject has broadly two divisions, one dealing with measuring instruments called electronic measurements and the other dealing with transducer sensors, monitoring, and controlling physical parameters called Instrumentation. In general, the manner in which instruments are put to use is classified as follows:

1. Monitoring of processes and operations.
2. Control of processes and operations.
3. Experimental engineering analysis.

This is only a general way of classification and there could be other applications as well.

Monitoring of processes and operations: Some measuring instruments have only a monitoring function; for example, thermometers and barometers. They simply indicate the condition of the environment and cannot control any function.

Control of processes and operations: Some instruments are put to use for the control of process operations; for example, home-heating system using a thermostat control. A temperature-measuring instrument senses the room temperature, thus providing information necessary for proper functioning of the control system. Missiles guided by the heat blast that is ejected from planes are another example.

Experimental engineering analysis: In research and development work, measuring instruments are used in carrying out experiments such as measurement of voltage, current, etc. Here the application of instruments is specifically intended for experimental purpose only.

1.2.1 Advantages of Instrumentation Systems

A few of the advantages of instrumentation systems include:

1. Remote measurement.
2. Accurate measurement.
3. Measurement in adverse conditions: nuclear reactors, space applications, etc.
4. Convenience: recording of data, printout, etc.
5. Reduction in size.

1.2.2 Block Schematics of Measuring Systems

Measurand: The quantity to be measured is called measurand.

Transducer: A device that converts a physical quantity into an electrical quantity or vice-versa.

Signal conditioner: Amplification, Filtering, Modulation, Demodulation, A/D conversion, etc.

Display/Record: The quantity is recorded using X–Y or strip-chart recorders or displayed on monitors, etc.

Any general measurement system as shown in Fig. 1.1 will have a transducer that converts the physical quantity into an electrical form, so that the signal may be modified suitably according to the requirements. Signal conditioning devices will carry out this task, and the parameter to be measured is displayed or recorded.

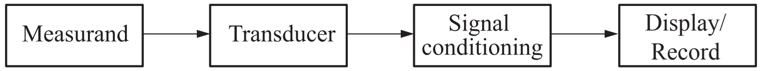


Figure 1.1 Block schematic of a general-purpose measuring system

1.2.3 Other Systems

1.2.3.1 Block schematic of telemetry system. Telemetry stands for measurement of a parameter from a distance. The parameter is measured in one place and recorded at a remote point (Fig. 1.2).

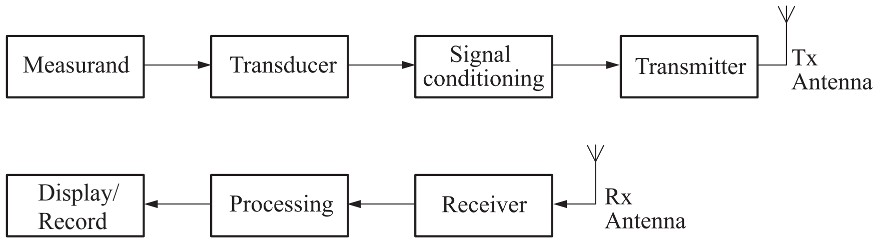


Figure 1.2 Block schematic of a telemetry system

1.2.3.2 Block schematic of control instrumentation. Here a feedback loop exists between the input and the output to control the parameter to be measured as shown in Fig. 1.3.

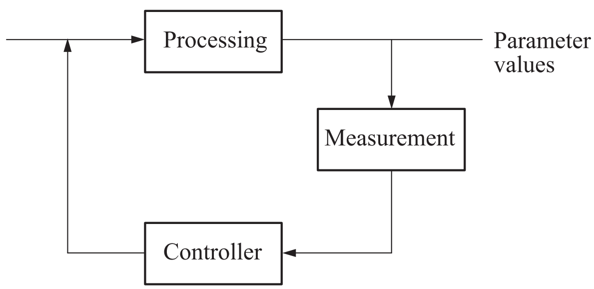


Figure 1.3 Control instrumentation system

1.2.3.3 Data process system. Here digital control techniques are employed using μP or μC on a PC as shown in Fig. 1.4. The analog signal is converted to a digital signal by ADC and is processed by microprocessor (μP) or microcontroller (μC) or PC.

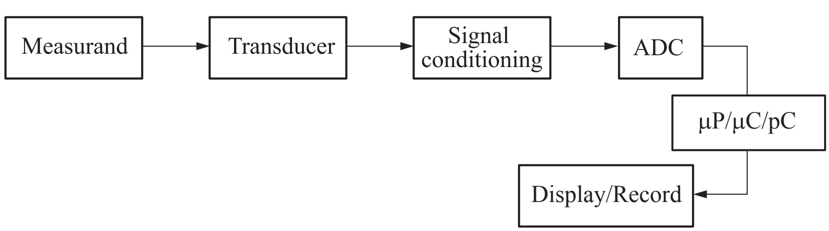


Figure 1.4 Data process system

1.2.4 Objectives of Measurement

1. To establish the validity of design.
2. To predict the limit of capacity.
3. To provide information needed to supplement further analytical work.

The type of instrument to be used depends upon the type of data. The data are classif ed as follows:

1. Steady-state data: If the data vary in the range of 0–5 Hz.
2. Transient data: If the parameter variation is at a much higher rate >5 Hz.
3. Dynamic data: The parameter variation is periodic.

Just as electronic engineering has different sub-branches like electronic devices, circuits, microwaves, radar engineering, etc., the subject of Instrumentation and Electronic Measurement have also vastly developed. Electronic instrumentation, mechanical instrumentation, bio-medical instrumentation meteorology are different sub-branches in this area.

The types of instruments used in electronic measurements are broadly classified as analog and digital. The instrumentation systems are also classified in the same manner. A comparison between them is given in Table 1.1.

1.2.5 Comparison between Analog and Digital Instruments

Table 1.1 Comparison of analog and digital instruments

	Analog	Digital
1.	Low precision	High precision
2.	Continuous stepless deflection	Numerical readout
3.	More flexible	Limited flexibility, different instruments for various ranges and applications
4.	Frequency response is large, from DC or low-frequency range to high-frequency range	Frequency response can also be very high
5.	Parallax error in readout is possible	No such error due to digital display
6.	Not convenient for readout	Convenience in readout, viz. in low light, from a distance, etc.
7.	Errors due to friction, spring tension can arise.	No such errors
8.	No direct PC or μP or μC compatibility	Compatibility with microprocessors (μP), microcontrollers (μC), and personal computers (PC) is possible

1.2.6 Factors for the Selection of Analog and Digital Equipments

Based on the requirements and applications, analog or digital instruments are selected considering the following factors:

1. Accuracy and precision of measurement (these parameters are introduced, subsequently, in the latter part of the notes).
2. Simplicity.
3. Minimum reading error.
4. Cost.
5. Resolution.
6. Display and readout.
7. Types of data to be measured.
8. Reliability.
9. Environment.
10. Availability.

Owing to the advantages associated with digital instruments, analog instruments are becoming obsolate. With μP and PC-based instrumentation systems becoming popular, digital instruments are being preferred.

1.3 PERFORMANCE CHARACTERISTICS

Measurement involves using an instrument as a physical means of determining the value of a quantity or a variable.

An instrument may be defined as a device for determining the value or magnitude of a quantity or a variable.

Different instruments are compared and analysed by the performance characteristic parameters. The performance characteristics are divided into

1. Static characteristics.
2. Dynamic characteristics.

Static characteristics, as the name implies, indicate the response of the instrument for slowly varying data or time-invariant data. Dynamic characteristics denote the behaviour of the instrument for time-varying quantities. The instrument design, testing, and evaluation are performed based on these parameters.

1.3.1 Definitions

1. Accuracy: Accuracy is the closeness with which an instrument reading approaches the true value of the variable being measured. Accuracy, in other words, indicates the maximum error, which will not be exceeded, as assured by the manufacturer of the instruments. If the accuracy of a 100 V voltmeter is ±1%, the maximum error for any reading will not exceed ±1 V. The term accuracy can be explained as '*conforming to truth*'.

2. Precision: The meaning may be given as '*sharply* or *clearly defined*'. It is the measure of order or degree to which a particular parameter is measured. This term is also distinctly different from the term '*reproducibility*'. The definition for the term reproducibility is given later in the section. A voltage reading expressed as 75.2347 V is a precise value. Therefore, precision indicates the degree or level or number of decimal places to which a particular quantity can be measured. But precision does not guarantee accuracy. The difference between the two terms must be understood clearly. For example, $\pi = 3.14$ is a correct or true value. It can be mentioned as an accurate value. But $\pi = 3.1428574$ is a precise as well as an accurate value. However, if the π value is given as 3.2428574, it is still a precise value because the value is expressed to more number of decimal places, but it is not an accurate value. Thus, precision does not guarantee accuracy. Digital instruments are more precise than analog instruments. A 3½ digit digital voltmeter (DVM) is more precise than a 2½ digit DVM.

3. Expected value: It is the value of the parameter being measured or computed. To use the right kind of instrument and to select the correct range for measurement, one should know the correct value of the parameter being measured. For example, if voltage is being measured, whether the value expected is in μV, mV or tens of volts, hundreds of volts, etc; then the proper range of the instrument is to be chosen. Only then will the measurement be accurate.

4. Sensitivity: Sensitivity of an instrument indicates the capacity of the instrument to respond truly to the change in the output, corresponding to the change in the input. For a voltmeter, sensitivity is referred to as $\Delta e_0 | \Delta e_i$, the ratio of the change in the output to the change in the input. If the output voltage changes by few millivolts, the output should also change by the same amount in the ideal case.

The ratio of change in the output signal to the change in the input is called sensitivity. For a voltmeter, it is the ratio $\Delta V_o / \Delta V_i$. If V_i changes by 0.1 V, the output reading should also change by 0.1 V. For a given meter, it may change by 0.08 V or less. This change in V_o for a change in V_i is expressed as sensitivity.

5. Resolution: Resolution is the smallest change in the measured value to which the instrument can respond. It is the smallest change the instrument can measure. For example, a 100 V voltmeter may not be able to measure 100 mV. Only when the minimum input is 0.5 V, the needle may deflect or the reading changes from 0. Any input or change in input less than 0.5 V may have no effect on the instrument. Therefore, the resolution for that particular instrument is 0.5 V. When V_i changes by 1 V and the output reading changes by 0.8 V only, the ratio is expressed as sensitivity. The difference between the two terms, sensitivity and resolution, must also be understood clearly.

6. Error: The difference between the measured value and the true value is called error. Error is expressed specific to a particular measurement. Accuracy indicates the maximum error for a given instrument. The error in a particular reading will be less than or equal to the value given by accuracy.

7. Repeatability: This is defined as the variation of scale reading when the input is randomly applied (with time gaps).

8. Reproducibility: This is the scale reading over a given period of time when the input is constantly applied. For a given voltmeter, if 10 V is applied as input, and the input is continuously connected to the instrument, the output reading of the voltmeter must be 10 V only. If it fluctuates, and reading changes, reproducibility of the instrument is poor. This parameter indicates the steady-state response of the instrument. If the input is intermittently applied, as long as the input is 10 V, the meter reading must be the same. If the reading is different each time, the repeatability of the instrument is poor.

9. Drift: There are three types of drifts.

1. *Zero drift* or *calibration drift*: If the whole calibration shifts by the same amount, because initially zero adjustment is not made, it is called zero drift.
2. *Span drift*: If the drift is not constant, but increases gradually with the deflection of the pointer, it is called span drift.
3. *Zone drift*: If the drift occurs only in a particular zone of the instrument, it is called zone drift.

10. Dead zone: The maximum value of the input to which the instrument does not respond due to hysteresis of the instrument is called the dead zone.

11. Threshold: It is the minimum value to which the instrument responds when the input is gradually increased from zero value.

12. Limiting error or guarantee error: The limit of deviation from the specified value is known as limiting error or guarantee error. Circuit components such as capacitors, resistors, etc., are guaranteed with a certain percentage of their rated value. If a resistor value is given as 100 Ω ± 1 %, the manufacturer guarantees that the resistance value falls between the limits 99 and 101 Ω. The error is guaranteed to be no greater than the limits set.

1.4 SIGNIFICANT FIGURES

An indication of the precision of the measurement is obtained from the number of significant figures in which it is expressed. The more the significant figures, the greater the precision. In the number 5.26, there are three significant figures. If the number is 5.267 there are four significant figures. A zero after the decimal has more significance if it is followed by a number. Observe the following numbers and the significant figures as shown in Table 1.2.

Table 1.2 Country significant figures

Number	Significant Figures
52	2
520	3
5.14	3
5.014	4
5.125	4
5.1250	5
50.02	4

If the resistance value is specified as 68 Ω, its value is closer to 68 than 67 or 69. If the value is specified as 68.0 Ω, its value is closer to 68 than 68.1 or 67.9 Ω. Therefore, a *zero after the decimal point is significant*.

When two or more measurements with different degrees of accuracy are added, the result is only as accurate as the least accurate measurement. For example, a bridge constructed with 10 pillars of which 9 are equally strong but with one weak pillar is just as strong as the weakest pillar.

13. Lag: Speed of response of the instrument is referred in terms of lag. If the input is changing rapidly, output must also change exactly at the same rate, in an ideal case. The time delay in the output to change with the input is termed as lag.

14. Dynamic error: Error in the reading is the difference between the true value and the measured value. If the error value is constant, it is referred to as static error. If the error in the measurement is not constant, but is changing, it is called dynamic error.

1.5 DYNAMIC CHARACTERISTICS

The dynamic characteristics of the instrument are important, if the instrument is to be used for varying or dynamic inputs. In practice, the parameters may vary at different rates. It is not possible to study or design instruments for these specific variations. Hence, certain standards have been prescribed for determining the dynamic behaviour of the system.

The inputs specified are as follows:

1. Step input.
2. Ramp.
3. Sinusoidal input.

The dynamic characteristics of an instrument are:

15. Fidelity: It is the quality of indication by the instrument with regard to the changes in input. This is indicated as dynamic error. Dynamic error is the difference between the true value of the process variable and the meter indication.

16. Speed of response: It is the rapidity with which the instrument responds to the changes in input. The delay in the response is known as lag.

The dynamic characteristics of the instruments are represented by differential equations. The order of the equation represents the order of the instrument. But, in general, the instruments are classified as

1. Zero order.
2. First order.
3. Second order.

This classification applies to electromechanical-type instruments with a moving element.

Zero-order instrument: E.g. Linear potentiometer. It is represented by the equation

$$a_0 y = bx$$

where a_0 and b are constants

$$x = \text{input}$$
$$y = \text{output}$$

The output follows the input with a proportionality constant $k = b/a_0$.

First-order instrument: E.g., Mercury thermometer. It is represented by the differential equation

$$a_1 \left(\frac{dy}{dt}\right) - a_0 y = b_x$$

Depending upon the type of input, bx is to be modified.

For a step input $x = x_o$

For a ramp input $x = x_r t$

For a sine input $x = x_s \sin \omega t$

Consider a step input

$$\therefore \quad \left(\frac{a_1}{a_0} s + 1\right) y = \left(\frac{b}{a_0}\right) x_0$$

This can be written in the form

$$(s\tau - 1)\, y = K x_0$$

where

$$K = \frac{b}{a_0}; \quad \tau = \frac{a_1}{a_0} \tag{1.1}$$

The solution to equation (1.1) with initial conditions, $y = 0$ at $\tau = 0$, is

$$y = K(x_0)\,(1 - e^{-t\tau}) \tag{1.2}$$

The measurement lag of the instrument is obtained from the graph by calculating the time taken by the output to reach 90% of the final steady-state value. It is also called the *measurement lag*. Though the input changes in a step function form, the instrument takes a definite amount of time.

The instruments are designed using these mathematical equations to meet the given specifications. The corresponding figure is shown in Fig. 1.5.

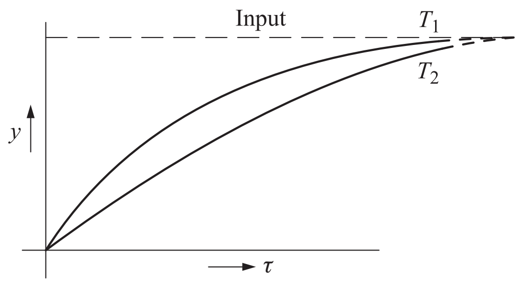


Figure 1.5 Response of the instrument for step input

Second-order instrument: E.g., Spring balance (Fig. 1.6). These are represented by the equation

$$a_2 \frac{d^2 y}{dx^2} + a_1 \frac{dy}{dx} + a_0 y + bx = 0 \tag{1.3}$$

where

x = input; y = output

$$f = M\frac{d^2 y}{dx^2} + B\frac{dy}{dx} + K_s y$$

where

M = mass
B = damping coefficient
K_s = spring constant

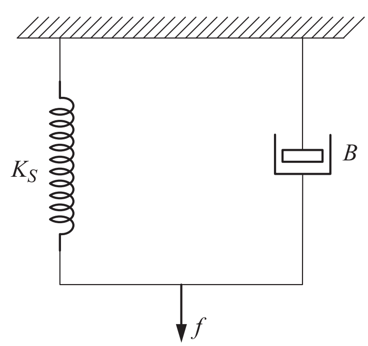


Figure 1.6 Spring balance equivalent circuit

1.6 TYPES OF ERRORS

Different types of errors that occur in measurements have been classified into three categories. These include

1. Gross errors.
2. Systematic errors.
3. Random errors.

This will enable users to avoid these errors and thus prove and increase the correctness in measurement.

1.6.1 Gross Errors

These are basically human errors caused by the operator or person using the instrument. The instrument may be good and may not give any error but still the measurement may go wrong due to the operator. The different types of gross errors are:

(i) Taking wrong readings.
(ii) Reading with parallax error.
(iii) Incorrect adjustments of zero and full-scale adjustments.
(iv) Improper applications of instruments: Using a 0–100 V voltmeter to measure 0.1 V, etc.
(v) Wrong computation: When power is to be determined, '*V*' and '*I*' are measured. If the computation goes wrong, even though '*V*' and '*I*' have been measured correctly, the measurement of power will be wrong. Thus, wrong computation can result in error.

1.6.2 Systematic Errors

These are divided into two categories:

(i) Instrumental errors: Due to shortcomings of the instruments.
(ii) Environmental errors: Due to external conditions affecting the instrument.

1.6.2.1 Instrumental errors. Even if human errors are avoided or proper care is taken to see that such errors do not occur, errors can still occur in measurements due to the instrument. The possible reasons can be as follows:

(a) Friction in bearings of various moving components can cause incorrect readings.
(b) Irregular spring tension in analog meters.
(c) Calibration errors due to aging.
(d) Zero setting not adjusted properly.
(e) Full-scale setting not adjusted properly.
(f) Faulty display circuit in digital instruments.

These errors can be avoided by

- Applying correction factors.
- Selecting suitable instruments.
- Calibrating the instruments.

1.6.2.2 Environmental errors. Ambient parameters such as temperature, pressure, humidity, magnetic and electrostatic fields, dust, and other such external parameters can affect the performance of the instrument. Improper housing of the instrument also can give wrong readings. Such errors can be avoided by air-conditioning, magnetic shielding, cleaning the instruments, and housing the instruments properly depending on the application and type of the instrument.

1.6.3 Random Errors

Though gross errors and systematic errors can be avoided by taking proper care, some other errors can also occur in measurements. No specific reason can be assigned and precaution could be taken to avoid these errors.

Such errors are categorised as random errors. Noise that is impracticable can cause random errors in measurements. To avoid these errors, frequency of measurement is to be increased, i.e., the same parameter is to be measured often. Error in such measurements can be estimated by statistical analysis.

1.7 STATISTICAL ANALYSIS

This method is employed to estimate the value or error when unpredictable errors or random errors are dominant. When the reason for specific error cannot be identified and the deviation from the true value is to be estimated, the statistical analysis method is to be employed. This will give the deviation from the true value, and the correctness of the readings taken. The statistical analysis is employed by taking a large number of readings of a particular parameter, and calculations are made in the following ways:

Arithmetic mean $\overline{x}$*:* A large number of readings are taken and the average value is computed. The average reading is the most probable value.

$$\overline{x} = \frac{x_1 + x_2 + x_3 + \cdots + x_n}{n} \tag{1.4}$$

$\overline{x}$ = arithmetic mean

n = number of readings taken

Deviation: It is the departure of a given reading from the arithmetic mean of the group of readings.

d_n = deviation of the reading x_n from the average $\overline{x}$

$$d_1 = x_1 - \overline{x}; \qquad d_2 = x_2 - \overline{x}; \qquad d_3 = x_n - \overline{x} \tag{1.5}$$

The deviation can be positive or negative.

Average deviation D: It is an indication of precision of the instruments used in making measurements. Highly accurate instruments with good reproducibility will have a low average deviation:

$$D = \frac{|d_1| + |d_2| + |d_3| + \cdots |d_n|}{n} = \frac{\sum |d|}{n} \tag{1.6}$$

Standard deviation σ*:*

$$\sigma = \sqrt{\frac{d_1^2 + d_2^2 + d_3^2 + \cdots d_n^2}{n}} = \sqrt{\frac{\sum d_n^2}{n}} \tag{1.7}$$

Variance σ^2*:* It is the mean square deviation, and σ^2 is calculated. *Probable error* = ± 0.6745 σ. Thus, the probable error in a given set of readings can be calculated.

1.7.1 Probability of Errors and Gaussian Curve

A method of presenting test results is in the form of a histogram or block diagram. The number of readings of the same quantity is taken and the measured value readings obtained are as given in Table 1.3.

Table 1.3 Set of measurements to estimate random errors

Measured Value (Hz)	Number of Readings
49.6	5
49.7	8
49.8	12
49.9	19
50.0	20
50.1	19
50.2	12
50.3	10
50.4	4

Now a graph is plotted between the number of readings and the value of frequency (Fig. 1.7). This graph is called a HISTOGRAM. The distribution and the shape of the curve indicate that the average value is the most probable value. The deviation on the positive and negative sides of the average value is uniform. Therefore, the middle or the average value is the most appropriate one.

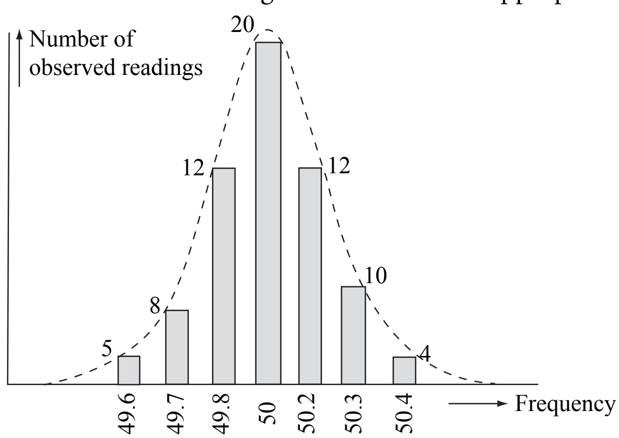


Figure 1.7 Histogram showing distribution of parameter readings

The Gaussian law or normal law: The Gaussian law of error is the basis for the study of random effects. When random errors are predominant, the probable error in a particular measurement can be estimated. The equation for the Gaussian law is

$$y = \frac{h}{\sqrt{\pi}} e^{-h^2\omega^2} \tag{1.8}$$

where

$$h = \text{constant} = \frac{1}{\sigma\sqrt{2}}$$
$$\sigma = \text{standard deviation}$$
$$\omega = \text{magnitude of deviation from the mean}$$
$$y = \text{probability of occurrence of deviation } \omega$$

For each reading, the probability of occurrence of deviation is calculated. A graph is plotted between y and ω as shown in Fig.1.8.

The deviation is calculated in terms of σ on both sides of the average value, and a graph is plotted. If the average value is the true value, the probability of occurrence of zero deviation is maximum. Therefore, corresponding to the average value, we will have a peak. If the shape of the graph is as shown in Fig. 1.9, we can confidently say that the average value is the most correct value.

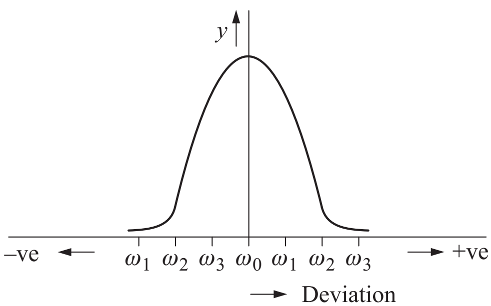


Figure 1.8 Probability of occurrence of deviations ω

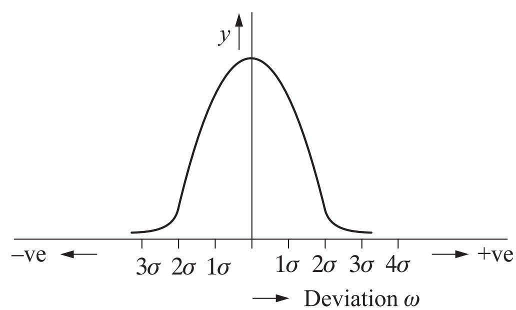


Figure 1.9 versus σ curve

Root Sum Squares (RSS) formula: In the statistical analysis method, to associate the probable error, the number of readings of the same parameters is taken, and graphs are plotted. A simple mathematical formula to estimate the probable error, by taking only one set of readings, is the Root Sum Squares (RSS) formula. If a parameter N is a function of other parameters $u_1, u_2, \ldots, u_n$, N is computed by measuring $u_1, u_2, \ldots, u_n$, and each measurement of $u_1, u_2, \ldots, u_n$ has a certain degree of error or uncertainty; the overall error, in the computed value of N, E_{RSS} is given by

$$N = f(u_1, u_2, \ldots, u_n)$$

$$E_{RSS} = \sqrt{\left(\Delta U_1 \frac{\partial f}{\partial u_1}\right)^2 + \left(\Delta U_2 \frac{\partial f}{\partial u_2}\right)^2 + \cdots + \left(\Delta U_n \frac{\partial f}{\partial u_n}\right)} \tag{1.9}$$

This formula is derived based on numerical mathematical analysis. The usage of the formula is illustrated in Example 1.1.

Example 1.1

A current of 10 A with a probable error of ± 0.1 A passes through a resistor of 100 Ω, with a probable error of $\pm 2\Omega$. Determine the power dissipated and the probable error.

Solution

$$P = I^2 R$$

$$u_1 = I; \quad u_2 = R; \quad N = P$$

Applying E_{RSS}

$$\Delta P = \text{variation in } P \text{ due to probable error}$$

$$\Delta P = \sqrt{\left(\Delta R \frac{\partial P}{\partial R}\right)^2 + \left(\Delta I \frac{\partial P}{\partial I}\right)^2} \quad (1.10)$$

$$\frac{\partial P}{\partial R} = I^2; \quad \frac{\partial P}{\partial I} = 2IR$$

Therefore,

$$\Delta P = \sqrt{(\Delta R I)^2 + (\Delta I \, 2IR)^2}$$

$$I = 10 \text{ A}; \qquad \Delta I = 0.1 \text{ A}$$

$$R = 100\ \Omega; \qquad \Delta R = 2\ \Omega$$

$$\Delta P = \sqrt{(2 \times 100)^2 + (0.1 \times 20 \times 100)^2}$$

$$\Delta P = \sqrt{8 \times 10^4} = 283 \text{ W}$$

$$P = I^2 R = 10^2 \times 100 = 10 \text{ kW} \quad (1.11)$$

Hence, in the computed value of power = 10 kW, the probable error is ±283 W. Thus, by making only one set of measurements, the probable error can be estimated.

1.8 MEASUREMENT STANDARDS

To establish uniformity in measurement and relative comparison and understanding, measurement standards have been established by international convention. The fundamental unit or mass in the International System (SI) is the kilogram, which is defined as the mass of a cubic decimeter of water at its temperature of maximum density of 4°C. The mass of the international prototype kilogram consisting of a platinum–iridium alloy cylinder is preserved at the International Bureau of Weights and Measures at Sèvres near Paris. Similar standards have been developed for other units including standards for the fundamental units as well as for some of the derived mechanical and electrical units.

For convenience and local use by industries, laboratories, and research organisations, the standards of measurement are classified as:

1. International standards.
2. Primary standards.
3. Secondary standards.
4. Working standards.

The international standards are derived by international agreement. International standards are periodically evaluated and checked by absolute measurements in terms of fundamental units.

Primary or basic standards are maintained by national standards laboratories in different parts of the world. The National Bureau of Standards in Washington is responsible for the maintenance of standards in North America. In India, the National Physical Laboratory (NPL) maintains primary standards. The function of the primary standards is the verification and calibration of secondary standards.

Secondary standards are reference standards used for industrial applications. These are maintained by the concerned industry and are checked locally against other reference standards. Secondary standards are generally sent to the national standards laboratories periodically for calibration and returned to the industry with a certification from the NPL.

Working standards are maintained in a laboratory. These are the principal tools of measurement. A manufacturer of precision resistances uses a standard resistor in the quality control department of his plant to check his testing equipment.

The standard for 1 m is the distance in which light propagates in vacuum in $\left(\dfrac{1}{29{,}97{,}92{,}458}\right)$ sec. The International Committee of Weights and Measures has defined seconds in terms of frequency of the cesium transition assigning a value of 9, 192, 631, and 770 Hz to the hyperfine transition of cesium atom unperturbed by external fields. IEEE has evolved standard procedures, nomenclature, definitions, etc., which are the standard test methods for testing and evaluating various electronic systems and components. IEEE 488 standard is one of the most important standards in digital interfacing for programmable instrumentation.

Example 1.2

A voltmeter has 100 scale divisions and can measure up to 100 V. Each division can be read to $\frac{1}{2}$ division. Determine the resolution of the voltmeter in volts.

Solution

Resolution is the smallest change in input that can be measured. The meter can be read to ½ division.

The resolution is $\frac{1}{2}$ division and its value in volts is

$$100\text{ div} = 100\text{ V}$$
$$1\text{ div} = 1\text{ V}$$
$$\tfrac{1}{2}\text{ div} = 0.5\text{ V}$$

Therefore, the resolution of the instrument is 0.5 V.

Example 1.3

A $3\frac{1}{2}$ digit DVM can measure 19.99 V. Determine the resolution in volts. (A $\frac{1}{2}$ digit in digital instruments corresponds to 1. A full digit can be display decimal numbers from 0 to 9.)

Solution

The maximum number of counts that can be made with $9.3\frac{1}{2}$ digit DVM is 1999. The smallest change in input that can be measured is 1 count.

1 count in volts corresponds to resolution

$$\begin{aligned} 1999\text{ counts} &= 19.99\text{ V} \\ 1\text{ count} &= ? \\ 19.99/1999 &= 0.01\text{ V or } 10\text{ mV} \\ \therefore\ \text{Resolution} &= 10\text{ mV} \end{aligned}$$

Example 1.4

A voltmeter having a sensitivity of 10 kΩ/V reads 75 V in its 100 V scale when connected across an unknown resistor when the current through the resistor is 1.5 mA. Calculate the percentage error due to loading effect.

Solution

Consider Fig. 1.10. Owing to the finite resistance of the voltmeter, it draws some current from the source. Ideally, the voltmeter must have infinite resistance and should not draw any current. Therefore, the net resistance measured V/I is the parallel combination of R_x and R_m and not R_x alone. This is called *loading of the source* by the meter and results in certain error in measurement.

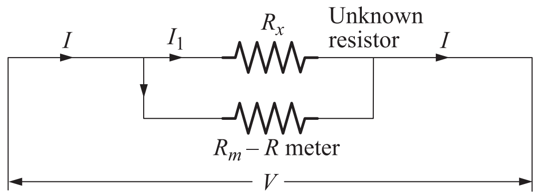


Figure 1.10 For Example1.4

$$R_{apparent} = \frac{75\text{ V}}{1.5\text{ mA}} = 50\text{ k}\Omega$$

$$R_{meter} = 100\text{ V} \times \frac{10\text{ k}\Omega}{\text{V}} = 1\text{ M}\Omega = 1000\text{ k}\Omega$$

R_x parallel with

$$R_m = \frac{R_x R_m}{R_x + R_m} = 50\text{ k}\Omega; \qquad R_m = 1\text{ m}\Omega; \qquad R_x = ?$$

$$R_x = 52.63\text{ k}\Omega$$

Percentage error due to loading effect: The true value of the unknown resistor R_x is 52.63 kΩ. The measured value is 50 kΩ.

$$\%\text{ error} = \frac{52.63 - 50}{52.63} \times 100$$

$$= 4.99$$

1.9 SUSPENSION GALVANOMETER

This is used for DC measurements and is one of the earliest measuring instrument mechanisms. A coil of fine wire is suspended in a magnetic field produced by a permanent magnet. When electric current is passed through the coil, the coil will rotate in the magnetic field. The fine filament suspension of the coil serves to carry the current. The elasticity of the filament sets up a moderate torque in opposition to the rotation of the coil.

The deflecting torque is the electromagnetic (EM) torque. The balancing torque is the mechanical counter torque of the suspension. The coil will deflect until these two torques become equal. The pointer attached to the coil indicates the corresponding position. The coil deflection is the measure of the current carried by it. An optical effect is also used by attaching a mirror to the coil, which deflects a beam of light.

The principle governing the suspension galvanometer applies to the *permanent magnet moving-coil* (PMMC) mechanism shown in Fig. 1.11. When current flows in the coil, an EM torque is developed and causes the coil to rotate. This EM torque is balanced by the mechanical torque or control springs attached to the movable coil. The expression for torque is

$$T = BAIN \tag{1.12}$$

where

T = torque produced (Newton-meters)

B = flux density in air gap $\left(\dfrac{\text{Wb}}{\text{m}^2} \text{ or Tesla(T)}\right)$

A = effective coil area (m^2)

A = Lw; L = Length of the coil; w = width of the coil

I = current in the movable coil

N = turns of wire on the coil

From equation (1.12), T is directly proportional to $B, A, I,$ and N. The flux density and the coil area are fixed for a given instrument. Typical values are: the coil area A will be in the range 0.5–2.5 cm^2, B ranges from 1500 to 5000 gauss or 0.15–0.5 Tesla. For a PMMC panel, instrument of 1 mA range and free-scale deflection of 100°, $A = 1.75\ \text{cm}^2$, $B = 0.2$ T, $N = 84$ turns, $T = 2.92 \times 10^{-6}$ Nm, coil resistance = 88 Ω, and power dissipation = μW.

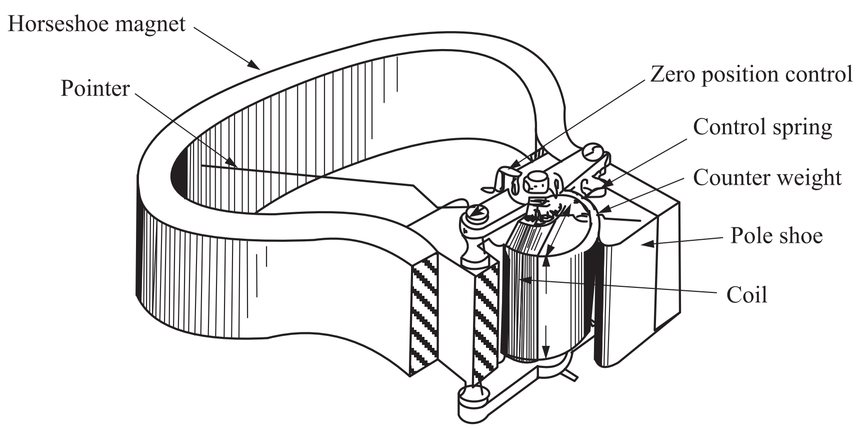


Figure 1.11 PMMC movement

1.10 D'ARSONVAL MOVEMENT

The PMMC movement was invented by D'Arsonval. Hence, it is also called D'Arsonval movement. The power consumption in this system is very low. It requires low current for free-scale deflection (FSD). This system has a permanent magnet in the horseshoe form with soft pole pieces attached to it. Between the pole pieces is a cylinder of soft iron. It provides uniform magnetic field in the air gap between the pole pieces and the cylinder. The coil is wound on a light metal frame and is mounted so that it can rotate freely in the air gap. The pointer attached to the coil moves over a graduated scale, which indicates the angular deflection of the coil and therefore the current through the coil. Thus, the *deflecting torque* T_d is produced by the EM effect. The *controlling torque* T_c is provided by *two phosphor–bronze conductive* springs. Spring tension must be uniform for maintaining the accuracy of the instrument. The moving system is balanced by three weights. The V-jewel is commonly used in instrument bearings. The top of the pivot may have a radius from 0.01 to 0.02 mm. The basic PMMC instrument is a *linear reading* DC instrument, because T_d, the torque, is directly proportional to the coil current.

$$T_d \propto I$$

The power requirements of the D'Arsonval movement will be in the range 25–200 μW. The accuracy of the instrument is of the order of 2–5% of *full-scale reading* (FSR). The PMMC instrument is unsuitable for AC measurements, unless the current is rectified before the application to the coil. Figure 1.12 shows the details of a moving coil for PMMC movement, Fig. 1.13 shows details of instrument bearings, and Fig. 1.14 shows the construction of the core magnet.

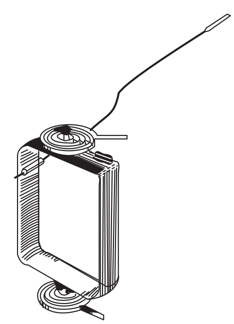


Figure 1.12 PMMC movement showing the control springs and the indicator

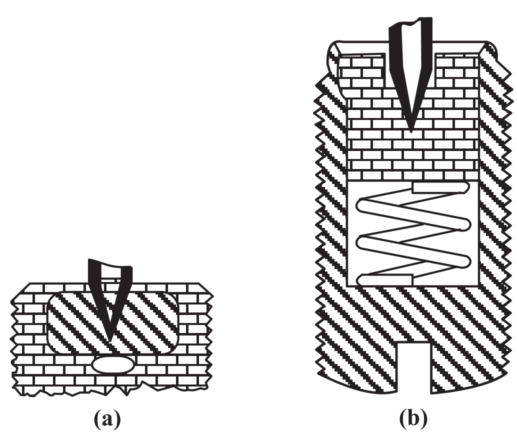


Figure 1.13 (a) V-jewel bearing and (b) spring-back jewel bearing

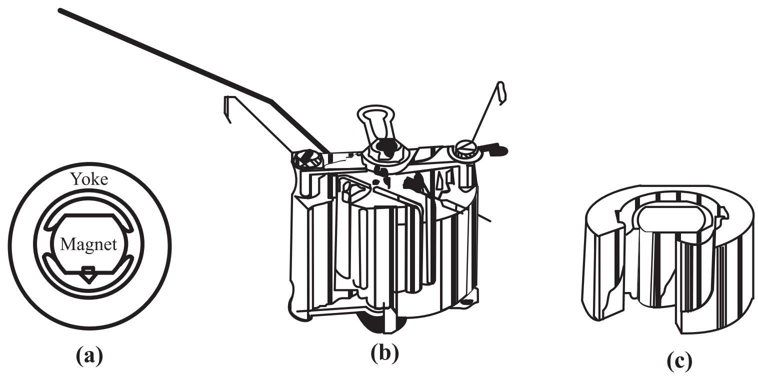


Figure 1.14 Construction of the core-magnet moving-coil mechanisms

1.10.1 Taut-Band Suspension

This is based on the suspension-type galvanometer mechanism. This is used in laboratories where high sensitivity is required and the deflecting torque is very low. When the torque is low, even the low friction of pivots and jewels are to be eliminated. Figure 1.15 shows construction details of a taut-band suspension mechanism.

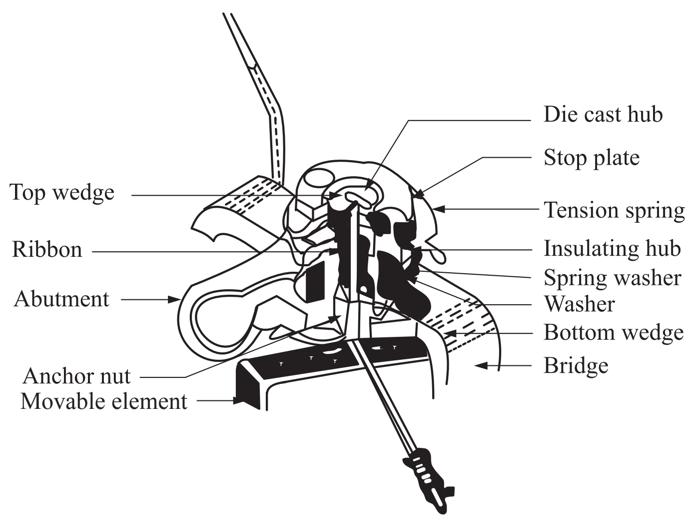


Figure 1.15 Taut-band suspension eliminates the friction of conventional pivot and jewel suspensions

The movable coil is suspended by means of two torsion ribbons. The ribbons are placed under sufficient tension to eliminate any sag. Taut-band instruments are relatively insensitive to shock and temperature. They are capable of withstanding greater overloads. The advantage in these instruments is that friction due to jewel–pivot suspension is eliminated.

1.10.2 Temperature Compensation

PMMC movement can be made insensitive to temperature effects by providing temperature compensation, and the appropriate use of series and shunt resistors of copper and *manganin*. As the temperature increases, the magnetic field strength and the spring tension decrease. The coil resistance increases with increase in temperature. The uncompensated PMMC mechanism tends to read low by approximately 0.2% per °C rise in temperature. The resistors that provide temperature compensation are called *swamping resistors*. These are connected in series with the moving coil, as shown in Fig. 1.16.

The swamping resistor is made of manganin and copper. Manganin has practically a zero-temperature coefficient. Manganin is mixed with copper in the ratio of 20/1 to 30/1. As the temperature increases, the total resistance of the coil and swamping resistor increase slightly. This will counteract the effect of temperature on springs and magnet. Therefore, the overall temperature effect is zero. A copper shunt resistor is also connected as shown in Fig. 1.17.

The resistance of the copper shunt resistor increases more than the series combination of coil and manganin resistor. Therefore, a large fraction of the total current passes through the coil circuit. By correct proportioning of the copper and manganin parts in the circuit, complete cancellation of temperature effects may be accomplished. The disadvantage is that full-scale deflection sensitivity is reduced.

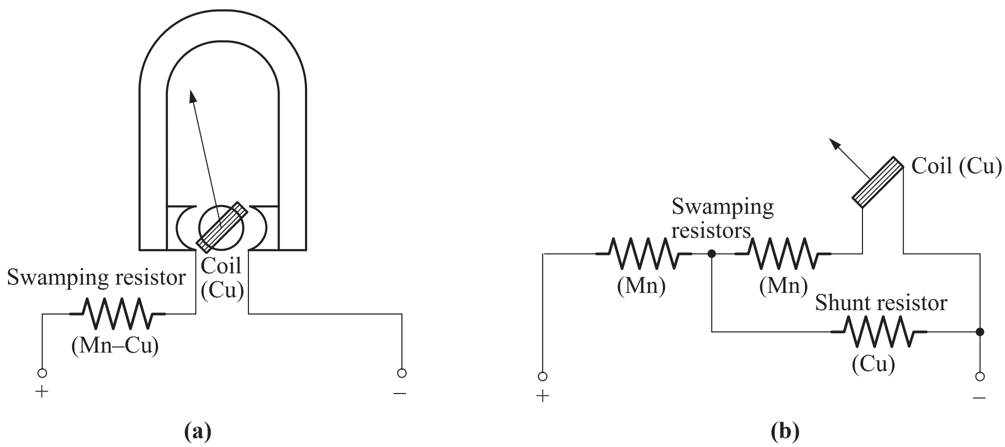


Figure 1.16 Placement of swamping resistors for temperature compensation of a meter movement. (a) Simple compensation circuit and (b) improved compensation using series and shunt resisters

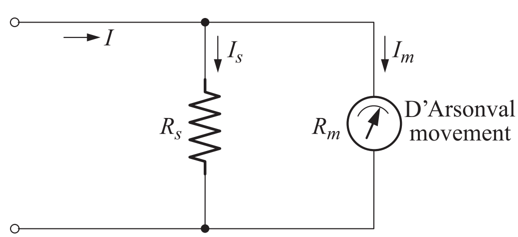


Figure 1.17 Basic DC ammeter circuit

1.10.3 Shunt Resistor

The coil of a PMMC movement can take very small current. If the instrument is to be used as ammeter to measure large currents, the excess current must be bypassed through a shunt.

The value of the shunt resistance to be connected can be calculated as shown below:

R_m = internal resistance of the coil
R_s = resistance of the shunt
I_m = full-scale deflection current of the movement
I_s = shunt current
I = full-scale current of the ammeter including the shunt

The voltage drops across the shunt and movement must be the same, as they are in parallel.

$$V_{shunt} = V_{movement}$$

$$I_{sh} R_{sh} = I_m R_m ; \quad R_s = \frac{I_m R_m}{I_{sh}} \tag{1.13}$$

$$I_{sh} = I - I_m \tag{1.14}$$

Therefore,

$$R_{sh} = \frac{I_m R_m}{I - I_m} \tag{1.15}$$

The *shunts are made from alloys of manganin or constantine*. Evenly spaced sheets of resistive material are welded into a large block of heavy copper on each end of the sheet.

1.10.4 Ayrton Shunt

The current range of the DC ammeter can be further extended by a number of shunts selected by a range switch. The Universal Shunt or the Ayrton Shunt is shown in Fig. 1.18. By this the overall meter resistance will be higher. Direct current ammeters are available over wide ranges from 20 μA to 50 A

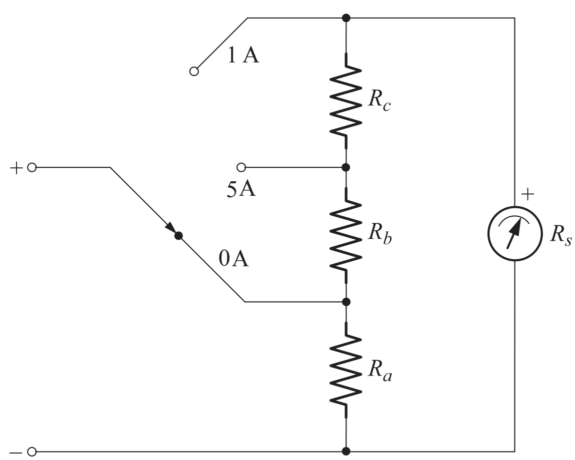


Figure 1.18 Universal or Ayrton shunt

full scale. With the external shunt, the range can be extended to 500 A. Ammeters should be connected in series with the voltage source. They should not be connected in parallel because due to the low resistance, heavy current may flow and destroy the delicate movement.

Advantages

1. Wide range of measurement.
2. Meter (ammeter) with a deflecting pointer always has a resistance in shunt.

1.11 DIRECT CURRENT METERS

The first DC measurement was done by Hans Oesterd in 1820, when he discovered a relationship between current and magnetism. In 1881, Jacques D'Arsonval developed and patented a PMMC meter.

In D'Arsonval meter, two types of suspension mechanisms are employed:

1. Jewel and pivot: Full-scale current: 50 μA.
2. Taut-band mechanism: Full-scale current: 2 μA.

The more sensitive and more expensive Alnico alloy of aluminium, nickel, and cobalt for magnetic materials and spiral phosphor bronze springs are used in D'Arsonval's meter. It also has a horseshoe-shaped permanent magnet with two soft iron poles (see Fig. 1.19).

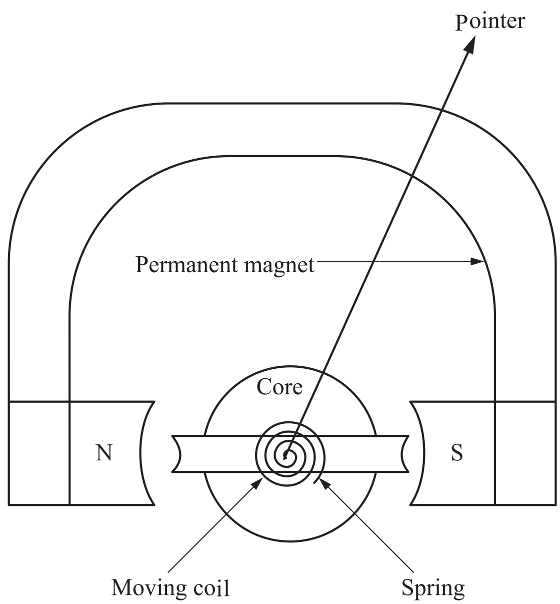


Figure 1.19 PMMC (permanent magnet moving coil) movement

Between the N and S poles, a cylindrical-shaped soft iron core is used about which a coil of fine wire is wound. This fine wire is wound on a longest metal frame and is mounted as a jewel setting, so that it can rotate freely.

The DC current to be measured passes through the windings of the moving coil and inducing forces make it to behave like an electromagnet, making the coil to rotate.

The pointer deflects when current passes through the coil in the proper direction. Therefore, all the DC meters have polarity in direction.

$$T_d = T_c$$

The pointer comes to rest when the deflecting torque T_d is equal to the controlling torque T_c. The direction of the current through the coil determines the poles of the induced electromagnet (Fig. 1.20).

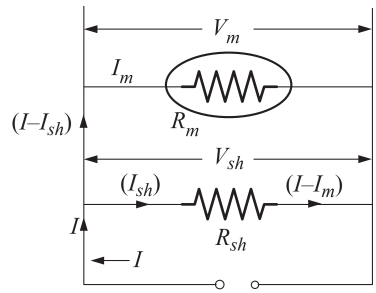


Figure 1.20 Current division of I into I_m and I_{sh}

For a DC ammeter:

$$V_m = (I_m R_m)$$
$$V_{sh} = V_m$$
$$I_{sh} = (I - I_m)$$
$$R_{sh} = \frac{V_{sh}}{I_{sh}} = \frac{I_m R_m}{I_{sh}} = \frac{I_m}{I_{sh}} R_m = \frac{I_{sh}}{I - I_{sh}} \times R_m$$

Example 1.5

Given 1 mA ammeter with $R_m = 100\ \Omega$ to measure 10 A, find R_{sh}.

Solution

$$I_{sh} = I - I_m = 10 \text{ mA} – 1 \text{ mA} = 9 \text{ mA}$$
$$V_m = I_{sh} R_{sh} = 1 \text{ mA} \times 100\ \Omega = 0.1 \text{ V}$$
$$R_{sh} = \frac{V_{sh}}{I_{sh}} = \frac{0.1 \text{ V}}{9 \text{ mA}} = 11.11\ \Omega$$
$$I = \text{Practical range of ammeter}$$
$$I = n I_m$$
$$n = \text{Multiplication factor} = \frac{I}{I_m}$$
$$R_{sh} = \frac{R_m I_m}{n I_{sh} - I_m}$$
$$R_{sh} = \frac{R_m}{n - 1}$$

Power dissipation of the shunt is the main criteria.

Example 1.6

Given ammeter range = 100 μA, R_m = 800 Ω, and range to be extended to 100 mA and 10A, find R_{sh} (Figs. 1.21, 1.22).

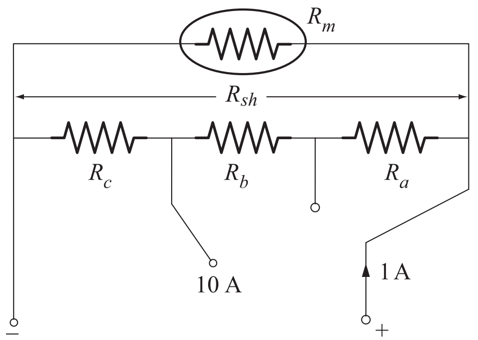


Figure 1.22 For Example 1.6

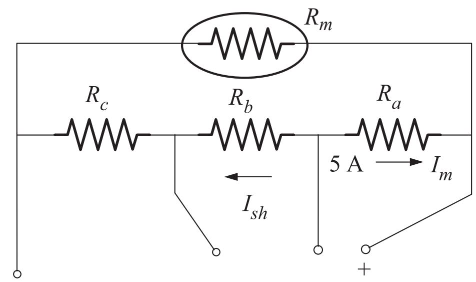


Figure 1.22 For Example 1.6

Solution

$$\frac{I}{I_m} = n = \frac{100\text{ mA}}{100\ \mu\text{A}} = 1000$$

$$\frac{R_m}{n-1} = R_{sh} = \frac{800}{999} = 0.80\ \Omega$$

1–A range:

$$R_a + R_b + R_c = R_{sh}$$

$$R_{sh} = \frac{R_m}{n-1}$$

$$n = \frac{I}{I_m}$$

$$I_m = \text{Meter current}$$

$$I = \text{Total range}$$

$(R_b + R_c)$ is in parallel with $R_m + R_a$. Because I_m current flows through $(R_m + R_a)$

$$V_{Rb + Rc} = V_{Ra + Rm}$$

or

$$(R_b + R_c)\,(I - I_m) = I_m\,(R_a + R_m)$$

or

$$I\,(R_b + R_c) - I_m\,(R_b + R_c) = I_m\,[R_{sh} - (R_b + R_c) + R_m]$$

$$\therefore\ R_a = [R_{sh} - (R_b + R_c)]$$

$$\therefore I(R_b + R_c) - I_m(R_b + R_c) = I_m R_{sh} - I_m (R_b + R_c) + I_m R_m$$

or

$$R_b + R_c = \frac{I_m (R_{sh} + R_m)}{I}$$

R_{sh} is known. Therefore, it is

$$R_a + R_b + R_e = \frac{R_m}{n-1}$$

$$\therefore R_a = R_{sh} - (R_b + R_c)$$

In the maximum current range of 10 A, I_m will flow through R_b, R_a, and R_m, and I_{sh} through R_c.

$$R_c = \frac{I_m (R_{sh} + R_m)}{I_{sh}}$$

$$\therefore I_{sh} R_c = I_m (R_a + R_b + R_m)$$

or I_{sh}

$$\begin{aligned}
(I - I_m) R_c &= I_m (R_a + R_b + R_m) \\
\therefore I R_c &= I_m (R_a + R_b + R_m) + I_m R_c \\
&= I_m (R_a + R_b + R_c + R_m) \\
\therefore R_b &= R_{sh} - R_a - R_c \\
IR_c &= I_m (R_{sh} + R_m) \\
\therefore R_c &= \frac{I_m (R_{sh} + R_m)}{I} = \frac{100 \times 10^{-6} (0.8 + 800)}{10 \text{ A}} \\
&= 0.08\ \Omega
\end{aligned}$$

Example 1.7

For the meter circuit shown in Fig. 1.23, determine the values of R_a, R_b, and R_c. $I_m = 100\ \mu$A; $R_m = 1$ kΩ.

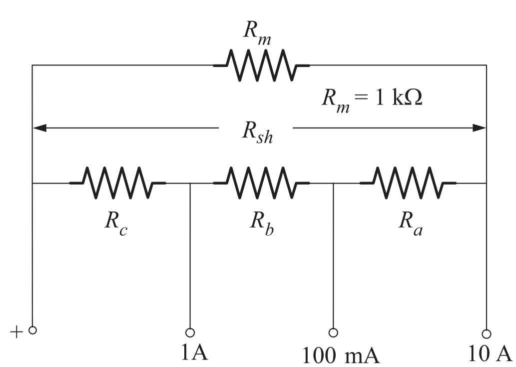


Figure 1.23 For Example 1.7

Solution

$$I_m = 100\ \mu A$$

$$R_{sh} = \frac{R_m}{n-1}$$

$$\therefore\ n = \frac{I}{I_n} = \frac{1000}{10} = 100\ \frac{1\times10^{-3}}{10\times10^{-6}} = 100$$

$$\therefore R_{sh} = \frac{1\,k\Omega}{100-1} = \frac{1\,k\Omega}{99} = 10.1\ \Omega$$

$$R_b + R_c = \frac{I_m\left(R_{sh}+R_m\right)}{I} = \frac{100\,\mu A(10.1\,\Omega + 1\,k\Omega)}{100\,mA} = 0.10\ \Omega$$

R_c is the resistance in the 1 A range. It comes in parallel with R_{sh}.

$$\therefore\ IR_c = I_m\,(R_{sh} + R_m)$$

or

$$R_c = \frac{I_m\,(R_{sh}+R_m)}{I}$$

$$= \frac{100\,\mu A(10.1\,\Omega + 1\,k\Omega)}{1\,mA} = 1.01\ \Omega$$

$$R_b = ?\ (R_b + R_c) = 1.01\ \Omega$$

$$R_c = 0.101\ \Omega$$

$$\therefore R_b = 1.01 - 0.101 = 0.909\ \Omega$$

$$R_a = R_{sh} - (R_b + R_c)$$

$$= 10.1\ \Omega - (0.909\ \Omega + 0.101\ \Omega) = 9.09\ \Omega$$

1.12 D'ARSONVAL METER MOVEMENT USED IN DC VOLTMETERS

By controlling resistance R_s in series with the meter, D'Arsonval meter can be used for voltage measurements. The purpose of R_s is to limit the current through the meter and extend the range to measure voltage. The circuit is as shown in Fig. 1.24.

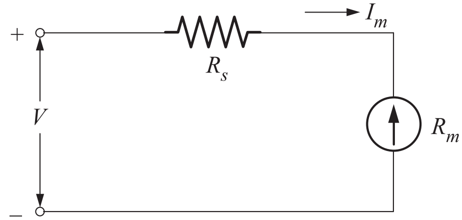


Figure 1.24 Circuit for voltage measurements

Sensitivity S is defined as $\frac{1}{I_{fs}}$, where I_{fs} = Full-scale deflection current.

$$\text{Sensitivity} = \frac{1}{\text{Ampere}} = \frac{1}{\text{Volt/Ohm}} = \frac{\text{Ohms}}{\text{Volts}}$$

$$R_s = ([S \times \text{Range}] - R_m)$$
$$= ((\text{Sensitivity} \times \text{Range}) - \text{Meter resistance})$$

Example 1.8

Calculate the value of the multiplier resistance on the 50 V- range of a DC voltmeter that uses a 500 μA meter movement with an internal resistance of 1 kΩ (Fig. 1.25).

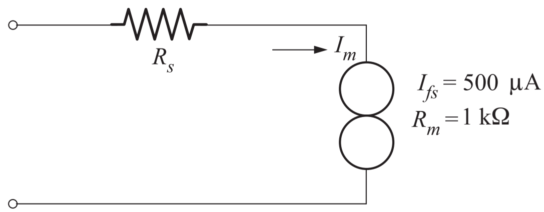


Figure 1.25 For Example 1.8

Solution

$$(R_s + R_m) = \text{Total Resistance} = (S \times \text{range})$$
$$R_m = (S \times \text{range}) - R_m$$
$$S = \frac{1}{I_{fs}} = \frac{1}{500\ \mu\text{A}} = 2\ \text{k}\Omega/\text{V}$$
$$R_s = S \times \text{Range} - R_m$$
$$= \frac{2\ \text{k}\Omega}{\text{V}} \times 50\ \text{V} - 1\ \text{k}\Omega = 100\ \text{k}\Omega - 1\ \text{k}\Omega = 99\ \text{k}\Omega$$

Example 1.9

Calculate the value of multiplier resistances for the multiple range DC voltmeter. The circuit is shown in Fig. 1.26.

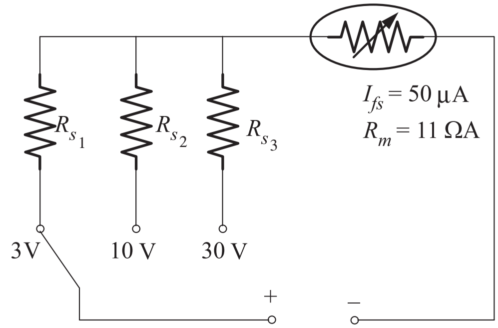


Figure 1.26 For Example 1.9

Solution

$$S = \frac{1}{I_{fs}} = \frac{1}{50\,\mu\text{A}} = 20\,\frac{\text{k}\Omega}{\text{V}}$$

The values of multiplier resistances in the different ranges are

$$3-\text{V range}: \quad R_{s1} = S \times \text{Range} - R_m$$
$$= \frac{20\ \text{k}\Omega}{\text{V}} \times 3\ \text{V} - 1\ \text{k}\Omega$$
$$= 50\ \text{k}\Omega$$

$$10-\text{V range}: \quad R_{s2} = S \times \text{Range} - R_m$$
$$= \frac{20\ \text{k}\Omega}{\text{V}} \times 10\ \text{V} - 1\ \text{k}\Omega$$
$$= 199\ \text{k}\Omega$$

$$30-\text{V range}: \quad R_{s3} = S \times \text{Range} - R_m$$
$$= \frac{20\ \text{k}\Omega}{\text{V}} \times 30\ \text{V} - 1\ \text{k}\Omega$$
$$= 599\ \text{k}\Omega$$

1.12.1 Ammeter Loading Effect

An ideal ammeter should have Rm = 0. It should not reduce the current flowing through the circuit due to internal resistance. This is referred to as loading effect due to ammeters (see Fig. 1.27).

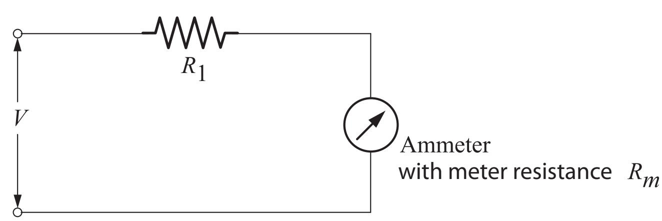


Figure 1.27 Ammeter loading effect

The expected current in the circuit in an ideal case is $I_e = \dfrac{V}{R_1}$

But due to finite resistance of the ammeter R_m

$$I_m = \left(\frac{V}{R_1 + R_m}\right)$$

$$\therefore \frac{I_m}{I_e} = \frac{R_1}{R_1 + R_m}$$

Example 1.10

A current meter that has an internal resistance of 78 Ω is used to measure the current through resistance R_c. Determine the % errors of the reading due to ammeter loading.

Solution

Consider the circuit shown in Figs. 1.28 and 1.29.

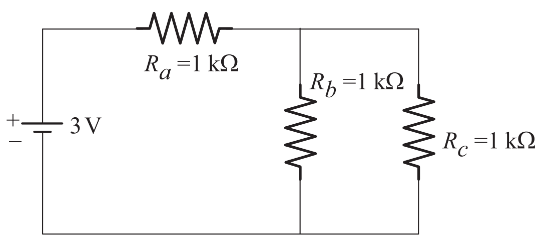


Figure 1.28 For Example 1.10

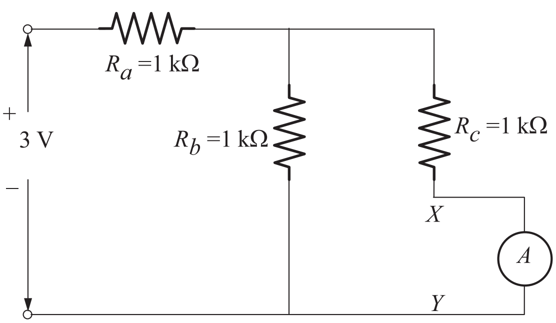


Figure 1.29 For Example 1.10

Thevenin's resistance:

$$R_1 = R_c + \frac{R_a R_b}{R_a + R_b}$$

$$R_1 = 1\text{ k}\Omega + 0.5\text{ k}\Omega = 1.5\text{ k}\Omega$$

$$\frac{I_m}{I_e} = \frac{R_1}{R_1 + R_m} = \frac{1.5\text{ k}\Omega}{1.5\text{ k}\Omega + 78\ \Omega} = 0.95$$

$$\therefore \quad I_m = 0.95\, I_e$$

The reading shown by the meter is 95% of the expected value

$$\therefore \quad \text{Error} = 5$$

$$\text{Loading error} = \left[1 - \left(\frac{I_m}{I_e}\right)\right] 100\% = 5\%$$

1.13 DC VOLTMETERS

By adding a series resistance or a multiplier, the D'Arsonval movement can be converted into a dc voltmeter. The series resistance R_s or the multiplier limits the current through the meter, so as not to exceed the full-scale deflection current I_{FSD} (Fig. 1.30).

The value of the multiplier resistance required to extend the voltage range is calculated as follows:

I_m = deflection current of the movement

R_m = internal resistance of the movement

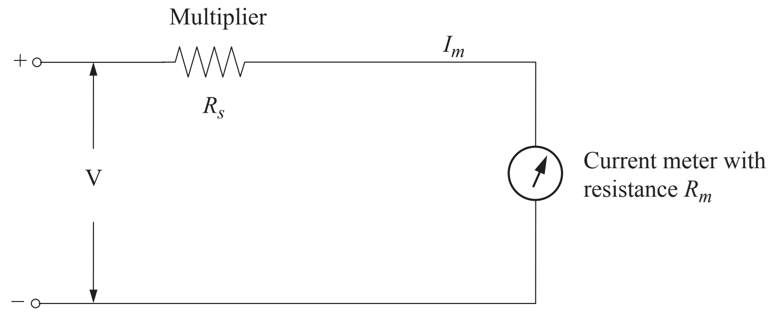


Figure 1.30 For DC voltmeter

$$R_s = \text{multiplier series resistance}$$

$$V = \text{full-range voltage of the instrument}$$

$$V = I_m (R_s + R_m) \tag{1.16}$$

$$\therefore \quad R_s = \frac{V - I_m R_m}{I_m} = \left(\frac{V}{I_m} - R_m\right) \tag{1.17}$$

Direct current voltmeters are available up to 500 V. The multiplier resistance is built into the meter. For higher voltage ranges, R_m is mounted separately.

1.13.1 Multirange Voltmeter

A voltmeter with different ranges can be obtained by connecting a number of multipliers. This is shown in Fig. 1.31.

R_1 is the multiplier resistance for the voltage range V_1
R_2 is the multiplier resistance for the voltage range V_2

and so on.

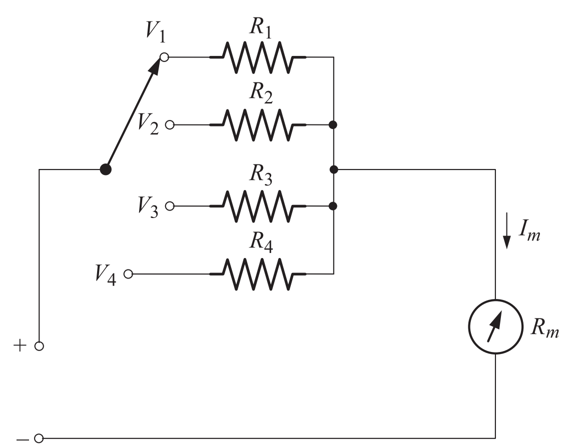


Figure 1.31 Multirange voltmeter

To determine the value of R_1, R_2, . . . , etc., the following equations are used:

$$I_m (R_m + R_1) = V_1 \tag{1.18}$$

$$I_m (R_m + R_2) = V_2 \tag{1.19}$$

$$\vdots \qquad \vdots$$

$$I_m (R_m + R_n) = V_n \tag{1.20}$$

The multiplier resistances can also be connected in series as shown in Fig. 1.32. The advantage with this system is that all multipliers except the first have standard prevention values and can be obtained commercially.

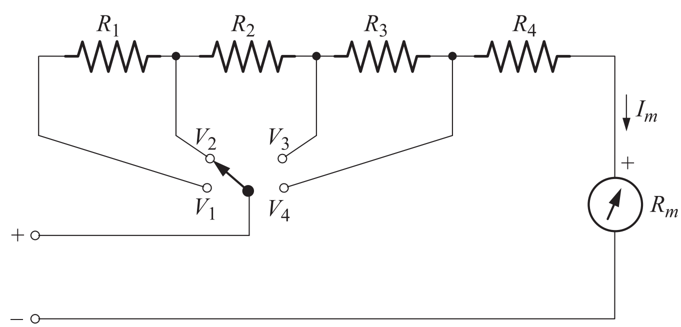


Figure 1.32 More practical arrangement of multiplier resistors in the multirange voltmeter

The resistance offered by the voltmeter for each range is expressed as the *sensitivity* of the voltmeter. It is expressed in Ω/V. It is also called the *ohm-per-volt rating of the voltmeter*. *An ideal voltmeter should have infinite input resistance*. When the voltmeter is connected across two points, it shunts the circuit or source. Therefore, the net resistance decreases. Due to this, the voltage measured will be less than the actual voltage. This is known as the *loading effect*.

1.14 OHMMETER

The basic D'Arsonval meter can be used to measure resistance. There are two types of ohmmeter:

1. Series type.
2. Shunt type.

1.14.1 Series-Type Ohmmeter

This meter consists of a D'Arsonval movement connected in series with a resistance and a battery to a pair of terminals where the unknown resistance R_x is connected. The current passing through the meter depends on the value of R_x. Therefore, the deflection of the pointer is proportional to R_x. The circuit diagram is shown in Fig. 1.33.

R_1 = current-limiting resistor
R_2 = zero-adjust resistor
E = internal battery

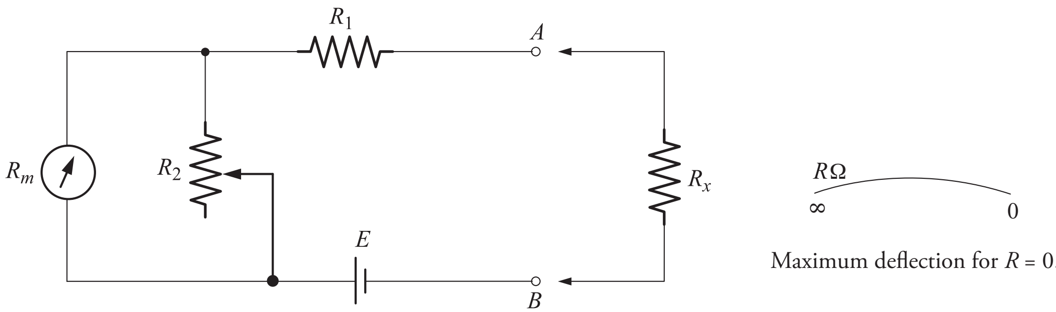


Figure 1.33 Series-type ohmmeter

R_m = internal resistance of D'Arsonval movement
R_x = unknown resistor

The disadvantage with the series-type ohmmeter is that it does not compensate for the decrease in battery voltage due to aging. As the source voltage decreases, the deflection or full-scale current decreases.

1.14.2 Shunt-Type Ohmmeter

The circuit diagram of a shunt-type ohmmeter is shown in Fig. 1.34.

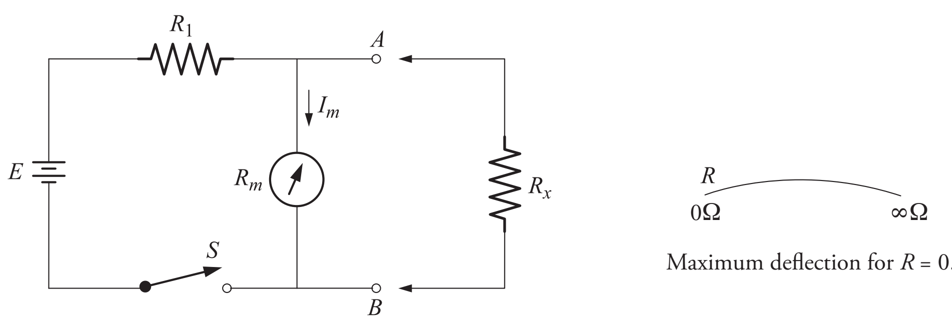


Figure 1.34 Shunt-type ohmmeter

The battery is in series with an adjustable resistance R_1 and the D'Arsonval movement. The unknown resistance is connected across terminals A and B in parallel with the meter. An on–off switch is to be provided to disconnect the battery from the circuit when not being used. If $R_x = 0$, the meter current is zero. If $R_x = \infty$, the current finds a path only through the meter. The meter can be made to read full scale by adjusting R. Thus, the meter deflection is proportional to the value of the unknown resistance R_x.

This meter is more suitable for measuring low values of resistances, upto 100 kΩ. This instrument is used in laboratories.

1.14.3 D'Arsonval Meter Movement Used in Ohmmeter

The basic D'Arsonval meter can also be used to measure resistance. The circuit is shown in Fig. 1.35.

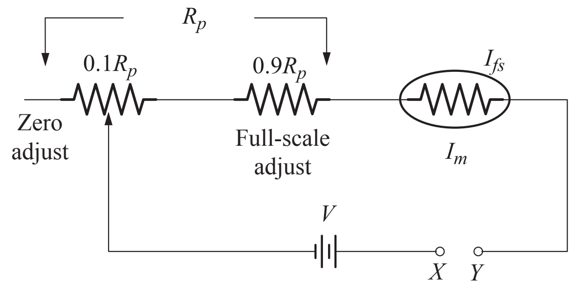


Figure 1.35 Equivalent circuit with 'zero adjust' and full-scale adjust

In Fig. 1.35, between the terminal X–Y, the unknown resistor R_x is connected. Therefore, the current flowing through the circuit is proportional to the resistance R_x.

If AB is shorted, a full-scale deflection I_{fs} will be obtained. If the needle deflection is less, it is adjusted by the potentioammeter a R_p. Usually this potential will have 10% of the total resistance R_p. This is the zero and ∞ adjustment in the ohmmeter.

Full-scale current

$$I_{fs} = \frac{V}{R_P + R_m}$$

$$\text{Current } I \text{ when resistor } R_x \text{ is connected} = \frac{V}{R_P + R_m + R_x}$$

$$\therefore \quad \frac{I}{I_{fs}} = \frac{R_P + R_m}{R_P + R_m + R_x}$$

This ratio $\frac{I}{I_{fs}}$ is represented as P and is equal to

$$\frac{R_P + R_m}{R_P + R_m + R_x} = \rho$$

Example 1.11

A 1 mA meter movement is to be used as an ohmmeter. R_m = 200 Ω. Battery supply = 5 V. Determine the resistor values for full-scale, 20%, 40%, and 50% of deflection.

Solution

I_{fs} = 1 mA; V = 5 V; R_m = 200 Ω; R_p = ?

$$I_{fs} = \frac{V}{R_m + R_P}$$

$$1 \text{ mA} = \frac{5V}{200\,\Omega + R_P}$$

$$\therefore \quad R_p = 4.8 \text{ k}\Omega$$

Value of R_x with 20% full-scale deflection,

$$P = \frac{I}{I_{fs}} = 20\% = 0.2 = \left(\frac{R_P + R_m}{R_m + R_P + R_x}\right) R_x = ?$$

$$\therefore R_x = \frac{R_p + R_m}{P} - (R_p + R_m)$$

$$= \frac{4.8\,\text{k}\Omega + 0.2\,\text{k}\Omega}{0.2} - (4.8\ \text{k}\Omega + 0.2\ \text{k}\Omega)$$

$$= (25 - 5)\ \text{k}\Omega = 20\ \text{k}\Omega$$

R_x for 40% FSD,

$$= \frac{R_p + R_m}{P} - (R_p + R_m) = \frac{5\ \text{k}\Omega}{0.4} - 5\ \text{k}\Omega$$

$$R_x = 12.5\ \text{k}\Omega - 5\ \text{k}\Omega = 7.5\ \text{k}\Omega$$

R_x for 50% full-scale deflection,

$$= \frac{R_p + R_m}{P} - (R_p + R_m)$$

$$= \frac{5\ \text{k}\Omega}{0.5} - 5\ \text{k}\Omega = 10\ \text{k}\Omega - 5\ \text{k}\Omega = 5\ \text{k}\Omega$$

∴ The multimeter scale is non-uniform.

For 50% of FSD, $R_x = (R_p + R_m) = R_e$ gives the total internal resistance of the multimeter.

1.14.4 Multiple Range Ohmmeters

(a) *R × 1 range, Equivalent circuit*: Equivalent circuit 20 Ω resistor is connected at $X - Y$ terminals in the $R \times 1$ range (Fig. 1.36).

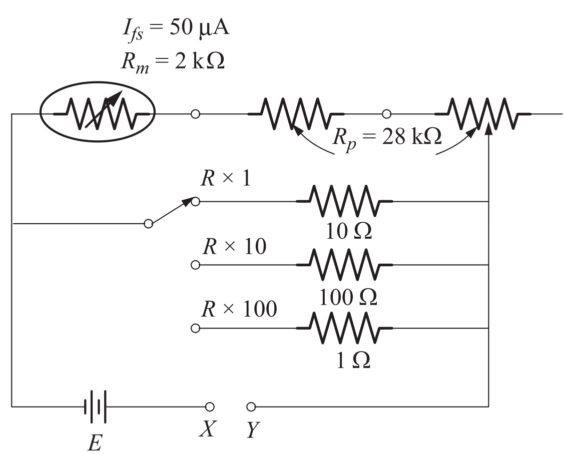


Figure 1.36 Equivalent circuit for ***R*** × *I* range

Voltage across 10 Ω resistor, V

$$= 1.5\text{ V} \times \left(\frac{10\ \Omega}{10\ \Omega + 20\ \Omega}\right) = 0.5\text{ V}$$

The current through the meter movement is (Fig. 1.37)

$$I_m = \frac{0.5\text{ V}}{30\text{ k}\Omega} = 16.6\ \mu\text{A}$$

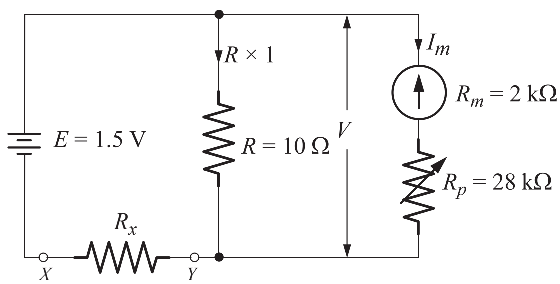


Figure 1.37 Equivalent circuit for $R \times 1$ range

(b) *R × 10 range*: When the ohmmeter is set on the $R \times 10$ range when 200 Ω is connected at $X - Y$ terminals (Fig. 1.38).

V across 100 Ω is

$$V = 1.5\text{ V} \times \frac{100\ \Omega}{100\ \Omega + 200\ \Omega}$$

$$= 1.5\text{ V} \times \frac{1}{3} = 0.5\text{ V}$$

Current through the ohmmeter (Fig. 1.38), when $R_x = 200\ \Omega$ is

$$I_m = \frac{0.5\text{ V}}{30\text{ k}\Omega} = 16.6\ \mu\text{A}$$

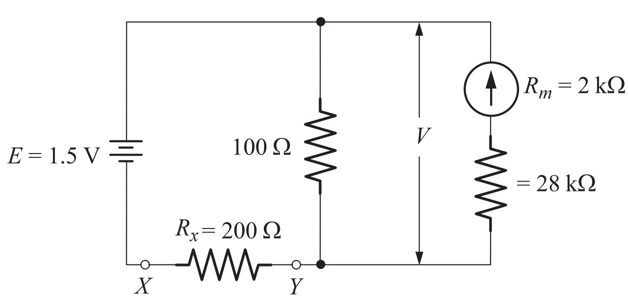


Figure 1.38 Equivalent circuit for R × 10 range

Multimeter
This is an instrument that measures *I*, *V*, and *R*. It usually uses D'Arsonval meter movement with a current range of 50 μA. Sensitivity = 20 kΩ/V.

Ranges for *I*, *V*, and *R*
Direct current: 0 – 50 mA, 0 – 1, 10, 100, 500 mA, 0 – 10 A.
DC volts : 0 – 250 mV, 0 – 2.5 V, 10, 50, 250, 1000, 5000 V.
AC volts : 0 – 2.5, 10, 50, 250, 1000, 5000 V.
Ohms: $R \times 1$, $R \times 100$, $R \times 10{,}000$.
Other specs: Input $|z|$, frequency range:10 Hz.

1.14.5 Electrolyte Capacitor Leakage Tests

If a capacitor is fully charged, from a DC source V_s, the charging current should stop. When $V_s = V_c$ and if the capacitor has a leakage, V_c tends to decrease and the charging current continues to flow. The current that flows, after V_c reaches a value close to V_s, is called the *leakage current*.

Usually for an electrolyte capacitor, the leakage current will be more. Therefore, using an ammeter, the leakage current of the capacitor can be detected. The circuit is shown in Fig. 1.39.

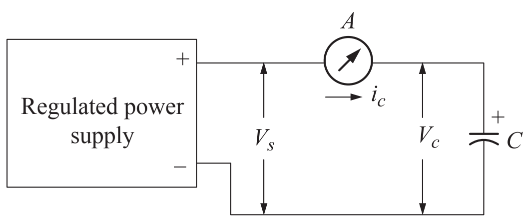


Figure 1.39 Measurement of leakage current of an electrolytic capacitor

Typical values of leakage current are:

Capacitor *C* rated for 300 V or more: 0.5 mA.

Capacitor *C* rated for 100 – 300 V: 0 – 2 mA.

Capacitor *C* rated < 100 V: 0.1mA.

1.14.6 For Non-Electrolyte Capacitors

An ideal capacitor should have infinite resistance in parallel with *C* (see Fig. 1.40).

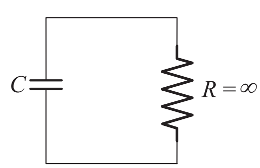


Figure 1.40 Equivalent circuit for an ideal capacitor

If the capacitor is not an ideal one, the supply voltage V_s will get divided across the meter resistance R_m of the voltmeter C and the finite resistance R of the capacitor (see Fig. 1.41).

$$\therefore\ R = R_m\left(\frac{V_s - V_c}{V_c}\right)$$

$$\therefore\ V_c = \frac{V_s R}{R + R_m}$$

$$\therefore\ V_c(R + R_m) = V_s R$$

or

$$R(V_c - V_s) = -V_c R_m$$

$$\therefore\ R = V_c R_m / (V_s - V_c)$$

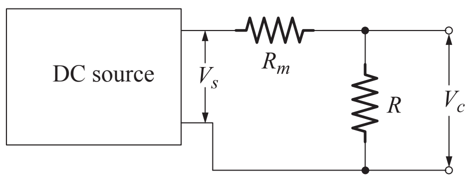


Figure 1.41 Equivalent circuit for a capacitor with finite shunt resistance '*R*'

1.15 MULTIMETER

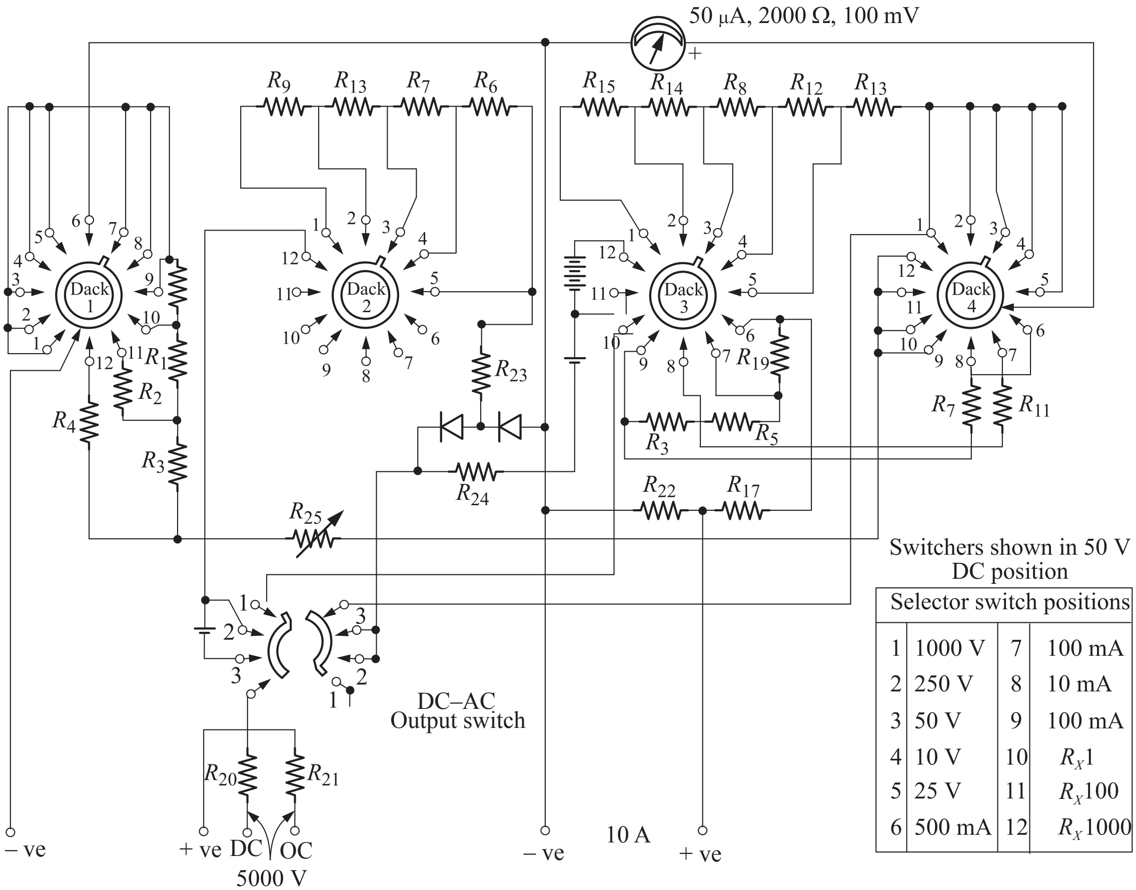
Combining the three measurements of *I*, *V*, and *R*, a single instrument can be constructed known as *volt-ohm-milliammeter* or *multimeter*. D'Arsonval movement can be used for all the three measurements. Hence, a single deflecting mechanism can be built to measure all the three parameters. The circuit diagrams of this type of instrument are shown in Fig.1.42.

The movement has a full-scale current of 50 μA.
Internal resistance = 2000 Ω; sensitivity = 20 kΩ/V.

Flux density produced = 1000–4000 gauss.

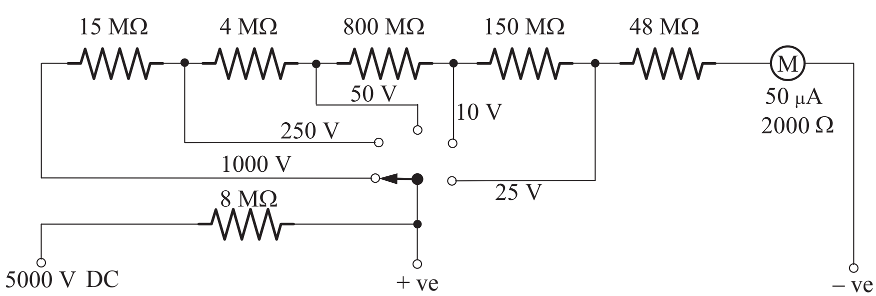
1.16 ALTERNATING CURRENT-INDICATING INSTRUMENTS

In D'Arsonval deflection system, the deflecting torque 'T_d' is directly proportional to current through the circuit 'I'; $T_d \propto I$. If the AC input is given for such an instrument, during the negative half cycle, a positive deflection is produced. During the negative half cycle, deflection in the opposite direction is produced. For a symmetrical AC input, the net deflection will be zero. If the frequency of the AC input is high, the instrument will not be able to respond at all. By rectifying the AC input and applying D'Arsonval meter, AC current and voltage can be measured. Another method is to use the heating effect of the AC to produce an indication of its magnitude.

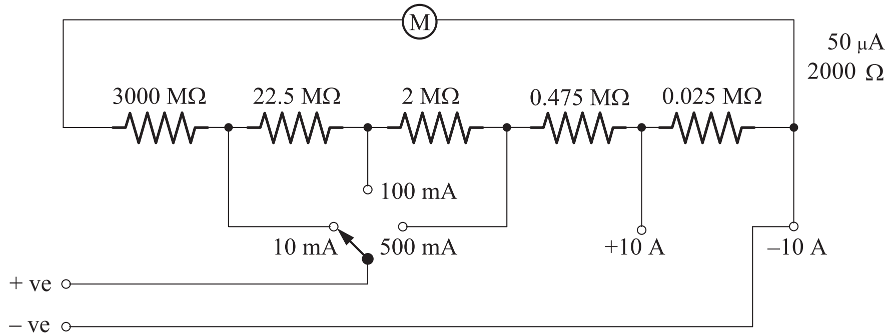


Selector switch positions			
1	1000 V	7	100 mA
2	250 V	8	10 mA
3	50 V	9	100 mA
4	10 V	10	R_X1
5	25 V	11	R_X100
6	500 mA	12	R_X1000

(a)



(b)



(c)

Figure 1.42 (a) Schematic diagram of a multimeter (b) DC voltmeter section of a multimeter and (c) DC ammeter section of the Simpson model 260 multimeter

1.16.1 Electrodynamometer

An electrodynamometer is used in the frequency range of 50 Hz and also up to the lower-order audio frequency range. Making modifications, an electrodynamometer can also be used as:

1. Watt meter.
2. VAR meter.
3. Power-factor meter.
4. Frequency meter.

Figure 1.43 shows the schematic diagram of an electrodynamometer movement.

In this type of a deflecting system, the permanent magnet used in D'Arsonval system to produce a magnetic field is replaced by a fixed value split into two equal halves. The current passing through these coils produces the magnetic field. The two coil halves are connected in series with the moving coil and are fed by current under measurement. The movable coil rotates. A pointer attached to the movable coil deflects and it is balanced by counter weights. Its rotation is controlled by springs. The complete assembly is surrounded by a laminated shield to protect the instrument from stray magnetic fields, which may affect its operation. *Damping is provided by aluminium air vanes*, moving in sector-shaped chambers. Using the same equation for the torque produced,

$$T = BAIN \tag{1.21}$$

where

T = Torque produced

B = Magnetic flux density

A = Cross-sectional area of the coil

I = Current through the coil

N = Number of turns in the coil

where the magnetic flux density B depends on I since it is an electromagnet (Fig. 1.43).

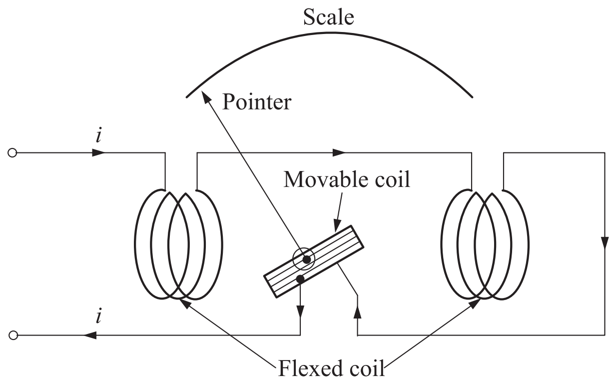


Figure 1.43 Electrodynamometer movement

Therefore,

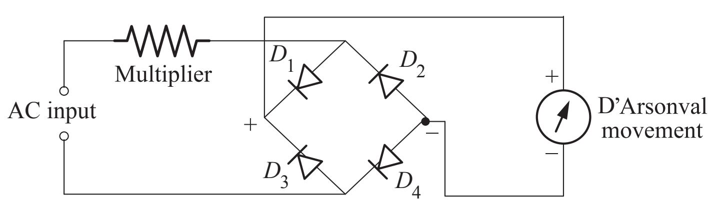
$$T = \text{IAIN} = I^2\text{AN}$$

Therefore, the torque produced $T \propto I^2$.

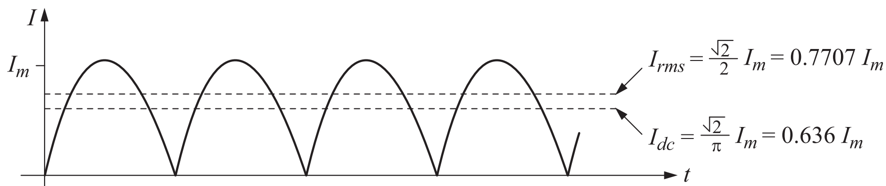
Hence, this instrument can be used for AC measurements in addition to DC measurements. It obeys the square law and the deflection scale is not linear. Even for the negative half cycle of AC as $T \propto I^2$, a positive torque is produced. The flux density is in the range of 60–100 gauss. However, in D'Arsonval movement, this will be 1000–4000 gauss. Due to the low value of 'B' in an electrodynamometer-type instrument, its sensitivity will be low. By connecting a series resistance, the electrodynamometer can be used as a voltmeter. The sensitivity of this voltmeter is in the range 10–30 kΩ/V compared to 20 kΩ/V of a D'Arsonval meter. If the frequency of the AC input increases, the reactance of the coils also increases. Therefore, the application of the instrument is limited to low-frequency AC inputs. At the 50 Hz frequency it is quite accurate and it is used as a secondary standard.

1.17 RECTIFIER-TYPE INSTRUMENTS

The problem of low sensitivity associated with the electrodynamometer movement for AC applications is overcome by rectifying the AC input and by giving to D'Arsonval movement, which has high sensitivity. A bridge rectifier circuit with germanium or silicon diodes is used. Copper oxide and selenium rectifiers have become obsolete because they have small Peak Inverse Voltage (PIV) ratings. The circuit and waveforms are shown in Fig. 1.44.



(a)



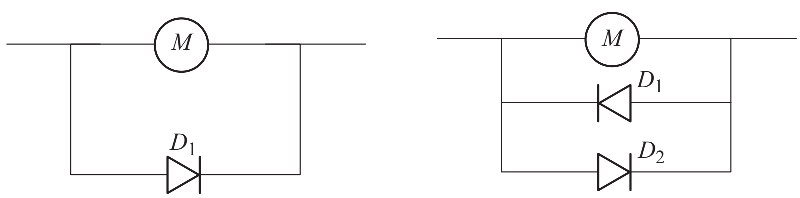
(b)

Figure 1.44 D'Arsonval meter for AC measurements

The bridge rectifier produces a pulsating unidirectional current through the meter movement over the complete cycle of the input voltage. Because of the inertia of the moving coil, the meter will indicate a steady deflection proportional to the average value of the current. AC voltages and currents are expressed in *rms* values. But the D'Arsonval deflection is proportional to the average value. Hence, the scale is to be calibrated in terms of the *rms* value of a sinusoidal waveform.

1.18 METER PROTECTION

The DC meters are usually protected by connecting a diode across the meter (Fig 1.45(a)). If a silicon diode is used, the voltage drop across the meter is maintained. At 0.6 V or just below the cut-in voltage V_γ of the diode, when the input exceeds the maximum value, the diode starts conducting as its cut-in voltage value is reached due to the excess input. Therefore, the current is diverted through the diode, protecting the meter. If only one diode is used it is called *single diode protection*. If two diodes are used for both half cycles or either direction of input, it is called *double diode protection* (Fig. 1.45(b)).



Single diode protection

(a)

Double diode protection

(b)

Figure 1.45 Meter protection

1.19 EXTENSION OF RANGE

The range of measuring instruments can be extended suitably by connecting a resistance in series or shunt as the case may be. If the range of an ammeter is to be extended, a 'shunt' resistance of a value lower than the meter resistance is connected in parallel with the meter as shown in Fig.1.46. The value of the resistance to be connected depends on the current to be measured.

Let

I = total current to be measured

I_s = current through the shunt

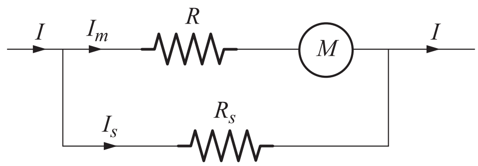


Figure 1.46 Extension of range of an ammeter

I_m = maximum current that can be permitted through the meter

R = resistance of the meter

R_s = shunt resistance

$$I = I_m + I_s \tag{1.22}$$

$$I_m = (I - I_s)\ ; \qquad I_s = (I - I_m) \tag{1.23}$$

$$I_s R_s = I_m R$$

$$\therefore R_s = \frac{I_m R}{I_s} = \frac{I_m R}{(I - I_m)} \tag{1.24}$$

$$\therefore R_s = \frac{R}{(I/I_m - 1)} \tag{1.25}$$

The ratio of the total current to the instrument current I / I_m is called *the multiplying power of the shunt*. This is usually expressed as N.

$$\boxed{\therefore R_s = \frac{R}{(N-1)}} \quad \text{or} \quad \boxed{N = 1 + \frac{R}{R_s}}$$

The shunts are available commercially. They consist of one or more thin strips of manganin, the ends of which are soldered to two heavy copper blocks. A special pair of leads are usually supplied with ammeters intended to be used. *Manganin has low temperature coefficient*. Therefore, it will not get heated up. The shunts will have good heat radiation ability.

The range of voltmeters can be extended in a similar way by connecting a high resistance in series with the meter, so that only the permeable maximum current passes through the meter (Fig. 1.47).

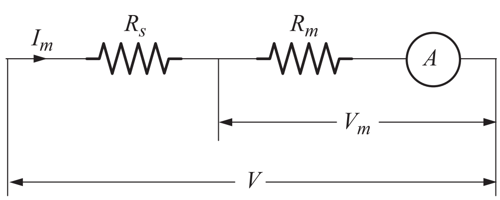


Figure 1.47 Extending the range of voltmeter

Let

J_m = maximum current that can pass through the meter

V_m = maximum voltage range of the meter

R_m = meter resistance

R_s = series resistance to be connected to extend the range of the meter

V = extended voltage range

$$\therefore V = I_m R_s - I_m R_m$$

$$I_m R_s = (V + I_m R_m)$$

$$\therefore R_s = \frac{(V + I_m R_m)}{I_m} \tag{1.26}$$

$$\boxed{R_s = \left(\frac{V}{I_m} + R_m\right)} \tag{1.27}$$

Thus, the value of R_s to be connected in series with the meter to extend its voltage range from V_m to V can be calculated.

1.20 FREQUENCY COMPENSATION

The inductance associated with the coil puts a limitation to the maximum frequency of the AC input that can be measured. The frequency range of a given AC instrument can be extended by using a capacitor (C) and a resistor (r) as compensating elements. Thus, a parallel combination of r and C is connected in series with the coil as shown in Fig. 1.48.

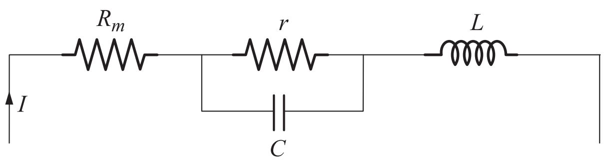


Figure 1.48 Frequency compensation

Let

R_m = meter resistance
L = inductance of the coil
r = compensating resistor
C = compensating capacitor

$$Z = R_m + j\omega L + \frac{r}{1 + j\omega cr} \tag{1.28}$$

$$Z = R_m + j\omega L + \frac{r(1 - j\omega cr)}{1 + \omega^2 c^2 r^2} \tag{1.29}$$

For frequency error elimination, the impedance Z must be a pure resistance.

$$\therefore \omega L - \frac{\omega cr^2}{1 + \omega^2 c^2 r^2} = 0 \tag{1.30}$$

$$\omega^2 c^2 r^2 = 1$$

$$\omega L - \omega cr^2 = 0 \tag{1.31}$$

$$\text{or } L = cr^2 \tag{1.32}$$

$$\boxed{c = \frac{L}{r^2}} \tag{1.33}$$

This is the value of C to be used for a given value of L and r to get frequency compensation.

Example 1.12

Calculate the resistance of a shunt required to make a milliammeter, which gives maximum deflection for a current of 15 mA and which has a resistance of 5 Ω, read up to 10 A (Fig. 1.49).

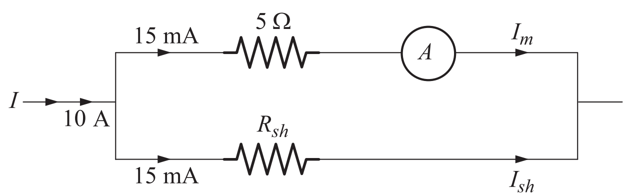


Figure 1.49 For Example 1.12

Solution

$$\begin{aligned}
I_{sg} &= (I - I_m)\\
&= (10\text{ A} - 0.015)\\
&= 9.985\text{ A}\\
9.985\, R_{sg} &= 15 \times 10^{-3} \times 5\ \Omega\\
\therefore\ R_{sg} &= \frac{15\times10^{-3}\times5}{9.985} = \frac{0.075}{9.985}\\
R_{sg} &= 0.007511\ \Omega
\end{aligned}$$

Example 1.13

The coil of a measuring instrument has a resistance of 1 Ω and the instrument reads upto 250 V when a resistance of 4.999 Ω is connected in series. Find the current range of the instrument when used as an ammeter with the coil connected across a shunt of resistance $\left(\dfrac{1}{499\,\Omega}\right)$ and the exact value of series resistance to be connected to measure 500 V DC.

Solution

$$I_{meter} = \frac{250}{4.999+1} = \frac{1}{20}\text{ A} = 0.05\text{ A}$$

The current through the meter should not exceed 0.05 A. $R_m = 1\ \Omega$. Hence, the voltage drop across the meter is $V = 0.05 \times 1 = 0.05$ V. I_{sg} when 1/499 Ω connected across the meter is

$$\frac{0.05}{(1/499)} = (0.05)\ 499 = 24.95\text{ A}$$

$$\text{Current range} = I_{sg} + I_m = 24.95 + 0.05 = 24\text{ A}$$

$$0.05\,(1 + R_s) = 500\text{ V}$$

$$\text{Series resistance } R_s = \frac{(500-0.05)}{0.05} = 999\ \Omega$$

Example 1.14

For the meter shown, determine the values of R_a, R_b, and R_c. Given I_m= 100 μA, R_m = 1 kΩ (Fig. 1.50).

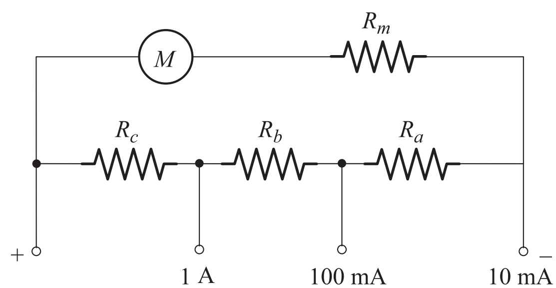


Figure 1.50 For Example 1.14

Solution

$$R_g = \frac{R_m}{N-1}$$

$$N = \frac{I}{I_m}$$

$$I = 10\text{ mA}$$

$$I_m = 100\ \mu\text{A}$$

$$\frac{I}{I_m} = N = \frac{10\times10^{-3}}{100\times10^{-6}}$$

$$\text{Shunt resistance, } R_{sh} = \frac{1\text{k}\Omega}{100-1} = \frac{1\text{k}\Omega}{99} = 10.1\ \Omega$$

In a 100 mA range,

$$R_b + R_c = \frac{I_m(R_{sh}+R_m)}{I}$$

$$= \frac{100\ \mu\text{A}\ (10.1\Omega+1\text{k}\Omega)}{100\text{ mA}}$$

$$= 1.01\ \Omega$$

R_c is the resistance in a 1 A range. It is in parallel with R_{sh}:

$$I\,R_c = I_m\,(R_{sh} + R_m)$$

$$R_c = I_m\,(R_{sh} + R_m)\,/\,I$$

$$R_c = 100\ \mu A\ (10.1\ \Omega + 1\ k\Omega)/1A = 0.101\ \Omega$$

$$R_b = ?$$

$$(R_b + R_c) = 1.01\ \Omega; \qquad R_c = 0.101\ \Omega$$

$$R_b = 10.1\ \Omega - R_c$$

$$R_b = 1.01\ \Omega - 0.101\ \Omega$$

$$\therefore\ R_b = 0.909\ \Omega$$

$$R_a = R_{sg} - (R_b + R_c)$$

$$\therefore\ R_a = 10.1 - (0.909 + 0.101) = 9.29\ \Omega$$

1.21 ELECTRONIC VOLTMETER (FOR DC)

Electronic voltmeters, in general, consist of the following:

1. A *potential divider network* to reduce the input in case it is high, to make it suitable, to give as input to the amplifier.
2. An *amplifier circuit* to enhance the signal so that the sensitivity and resolution of measurement improve.
3. A *rectifier and filter circuit* in case the meter for deflection is a DC meter.

Usually, the electronic circuits generate a current proportional to the quantity being measured. Many digital instruments have auxiliary provisions to make permanent records of measurement results using printers or magnetic tape recorders. mP-based instruments with PC compatibility are the new type of instruments developed so far. The general block schematic of an electronic DC voltmeter is shown in Fig. 1.51.

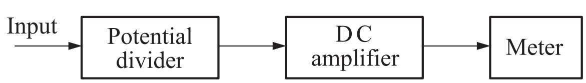


Figure 1.51 Block diagram of an electronic DC voltmeter

The potential divider network is nothing but a series of resistors to alternate the value of input in case it is large. If the signal magnitude is small and needs no attenuation, it is passed directly. The signal is amplified by the DC amplifier and then given to the meter. The meter is calibrated in terms of the parameter to be measured. The DC amplifier used can be

1. Direct-coupled amplifier.
2. Chopper-type DC amplifier.

Direct-coupled amplifiers are preferred because they are economical. A typical circuit for a FET input electronic DC voltmeter is shown in Fig. 1.52.

The DC input voltage is applied to a *Range attenuator*, which is a potential divider network. It is calibrated on the front panel control. The attenuator is necessary to provide input voltage levels, which the DC amplifier can take. The input stage of the amplifier consists of a JFET. It provides high input impedance and isolates the meter circuit from the input. Hence, loading of the input on a circuit under test can be prevented. The two BJTs form a direct-coupled DC amplifier, driving a meter movement. The transistor Q_2 is in a common-base configuration. It provides the voltage gain. Transistor Q_1 is in

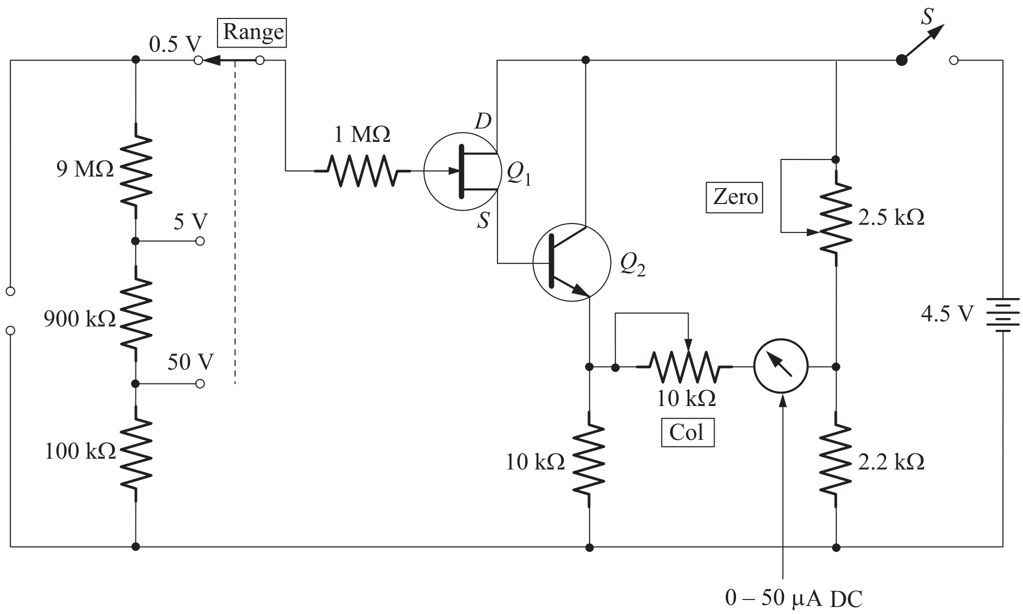


Figure 1.52 Circuit diagram for an electronic DC voltmeter

a common-collector (CC) configuration. Its voltage gain is less than one, but provides a large current gain. The input to the meter is the amplified version of actual input. The output from the collector of Q_2 is directly coupled to the base input of Q_1. Hence, it is named direct-coupled amplifier. The output current from the emitter of Q_1 since it is in the CC configuration is given to the meter. Zero adjustment of the meter can be done with the help of R_2. Full-scale adjustment can be done with R_s. The gain in the amplifier allows the instrument to be used for the measurement of even millivolts.

This circuit has the added advantage that accidental high voltages do not damage the instrument because amplifier saturation limits the maximum current through the meter.

A chopper-type DC amplifier is used to avoid drift problems. These are used in high-sensitive instruments. Zero adjustment is done using R_2 to make the transistor Q_2 to go to cut-off.

1.22 ELECTRONIC VOLTMETER (FOR AC)

Electronic AC voltmeters are similar to electronic DC voltmeters except that the signal is rectified before being amplified. The block schematic is shown in Fig. 1.53.

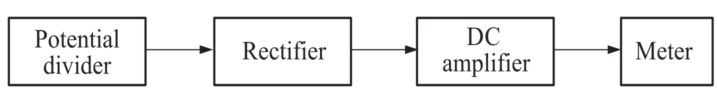


Figure 1.53 Block diagram for an electronic AC voltmeter

However, designing a DC amplifier is more difficult and expensive, because the input signal right from 0 Hz is to be amplified. The input AC signal is first amplified and then rectification and filtering are done and a DC meter is used for deflection. Therefore, sometimes an AC amplifier is found to be more convenient. The block schematic of an AC voltmeter is shown in Fig. 1.54.

Figure 1.54 Block diagram for an electronic AC voltmeter (another version)

AC voltmeters are subdivided into three categories:

1. Average reading voltmeter.
2. Peak responding voltmeter.
3. True RMS responding voltmeter.

The difference between average and peak responding voltmeters is only in filter circuits.

1.22.1 Average Reading Voltmeter

The simplified circuit diagram is shown in Fig. 1.55. This meter reads the average value of a positive half cycle or a negative half cycle of the AC input. It depends on the position of diode D. In the above figure, the diode conducts during the positive half cycle of the AC input. It provides half wave rectification of the input. The average value of the positive half cycle is developed across the resistor R. This is applied to the DC amplifier and meter. If the diode position is reversed, the meter indicates the voltage across the negative half.

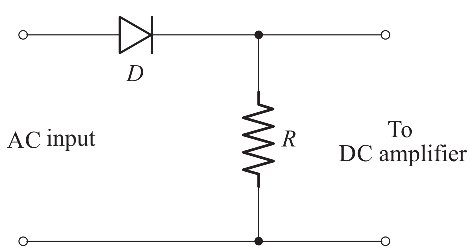


Figure 1.55 Circuit diagram for an average reading voltmeter

1.22.2 Peak Reading Voltmeter

A circuit diagram for a peak reading voltmeter is shown in Fig. 1.56.

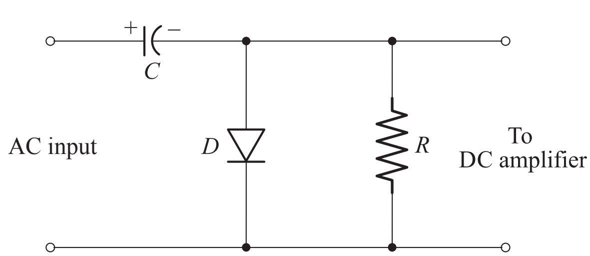


Figure 1.56 Circuit diagram for a peak reading voltmeter

In this circuit, since the diode D is connected as shown in Fig. 1.56, it gets forward biased during the positive half cycle. Therefore, the capacitor gets charged to the positive peak. The charge cannot leak off rapidly because of the one-way conduction of the diode. The voltage across the meter stays near the peak value of the input. The value of resistor R is greater than the forward resistance of the diode but less than the reverse resistance. Therefore, when the diode is forward biased, it gets charged through the diode. When the diode is reverse biased, the discharge path is through resistor R. If the position of the diode is reversed, the meter reads negative peak value of the input.

Peak reading voltmeters or Peak detectors are used in coaxial configurations to measure signals up to 40 GHz, by keeping the diode and the capacitor in a probe without applying to the amplifier or meter.

In the average reading circuits, the input is full-wave rectified (FWR), and the low-pass filtering characteristic of the meter movement is used to extract the average value.

The *rms* reading meter circuit approximates the required square law parabola with a few straight line segments in the fashion of a diode function generator. The voltage applied to the meter is only an average value. However, the scale is calibrated for *rms* value.

1.22.3 Peak-To-Peak Detector

In the peak detector instrument, only the positive peak or negative peak of the input will be measured. If the input waveform is not symmetrical, the positive and negative peak values will be different. An unsymmetrical waveform causes turnover error in meters. This is overcome in peak-to-peak reading voltmeters. The circuit diagram for this is shown in Fig. 1.57.

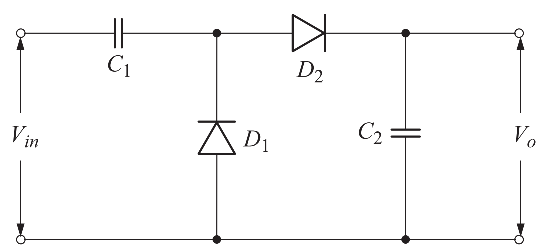


Figure 1.57 Circuit diagram for a peak-to-peak reading voltmeter

The detection efficiency of the peak-to-peak detector is twice that of a peak responding meter. During the negative half cycle of the AC input, diode D_1 becomes forward biased. C_1 charges up to approximately the negative peak voltage. When voltage V_1 goes positive, D_1 is reverse biased and D_2 becomes forward biased. The change on C_1 is gradually transferred to C_2 during the initial transient period. When the circuit is in steady-state operation, the output voltage is the sum of the voltage developed across C_1 during the negative portion of the input V_1 and positive peak of V_1, which is equal to the peak-to-peak input voltage. C_1 and C_2 must be large enough, so that the voltage does not change appreciably across C_2 during one period of the input voltage and the voltage across C_1 does not appreciably change in the process of recharging C_2.

Peak-to-peak detectors are used for non-sinusoidal waveforms or complex waveforms. These are used in communication systems for modulated waves.

1.23 DC METER WITH AMPLIFIER

Another typical circuit of a DC voltmeter employing an amplifier is shown in Fig. 1.58. This amplifier circuit has high input impedance. Therefore, the current drawn from the circuit under test will be less. This circuit has unity gain amplifier with a high input resistance. A source follower drives an emitter follower.

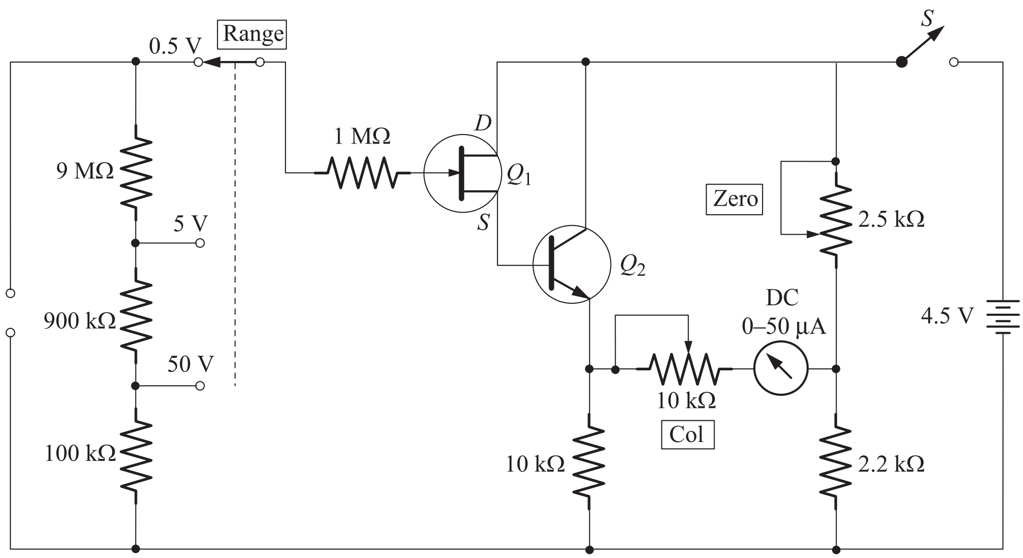


Figure 1.58 DC voltmeter circuit with FET input

The input impedance of the meter is 10 MΩ; the power required for 0.5 V deflection is 0.025 μW.

The power required for a meter without amplifier for the same 0.5 V deflection will be 25 μW. Therefore, there is an increase in sensitivity by 100 times.

Figure 1.59 shows the simplified circuit schematic of a meter that can measure voltage and current. The input voltage is amplified and applied to a meter. If the gain in the amplifier is 10, the sensitivity of the meter also increases 10 times.

For meter deflection, if a milliammeter is being used, the driving current required must be of the same order. If a DC current of nanoamperes is to be measured, the amplifier of the meter should have a gain of 10^6. For a DC application, direct coupling is to be employed. Capacitor coupling cannot be used since the input is DC when the gain in the operational amplifier (Op-amp) being used is so large, and the limitations of the Op-amp also come into the picture. The offset current, drift, etc., will also be of the same order. Therefore, using trim-potentiometers accessible on the panel of the meter, adjustments are to be provided to reduce errors due to drift, offset, etc. However, temperature and time-induced drifts would cause errors and these cannot be minimised easily. Instrumentation amplifiers with the required characterisation have to be used in such cases.

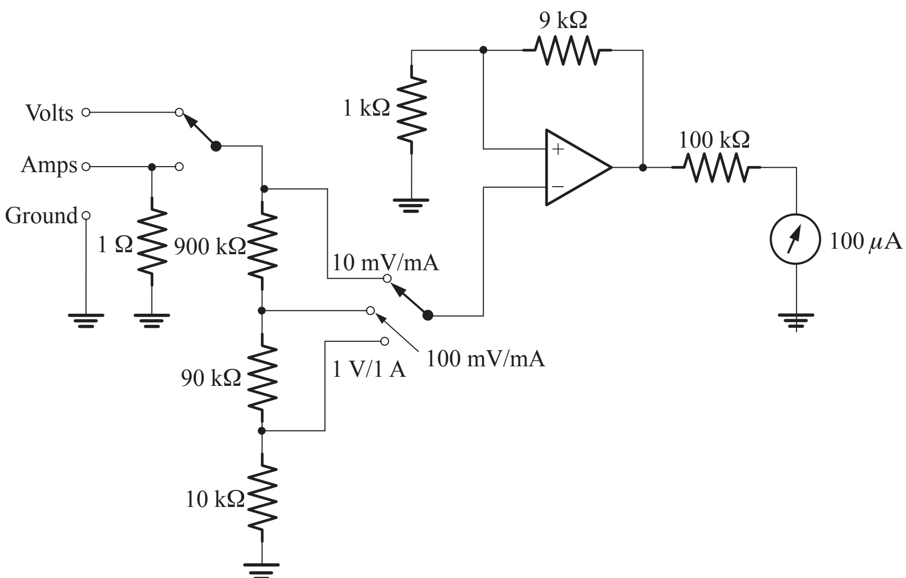


Figure 1.59 Amplifier voltage and current meter

1.24 CHOPPER-STABILISED AMPLIFIER

To measure very small DC voltages and currents, a high-gain DC amplifier is required. However, this amplifier will have drift and offset problems. If an AC amplifier can be used, these problems can be overcome. Therefore, if the DC is converted to AC, an AC amplifier can be used. This is the technique employed in meters with chopper-stabilised amplifiers. Figure 1.60 shows the circuit schematic.

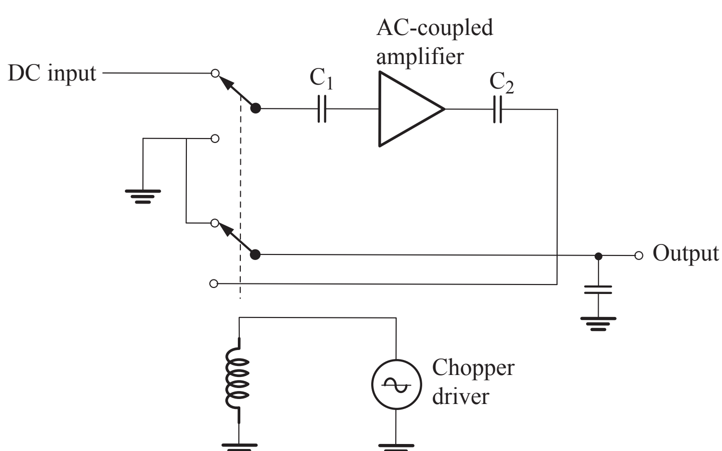


Figure 1.60 An AC-coupled amplifier can be used to amplify DC signals if the input and output are chopped using the circuit shown

The input signal is converted to AC by chopping, that is, the input is connected between the amplifier and the ground by an electronic switch or an electromechanical chopper, which is similar to a relay. The output of the chopper is an AC signal with a peak value equal to the input DC voltage. The resulting waveform will have a DC component of approximately one-half of the input DC voltage. The chopped voltage is fed to an AC amplifier.

The amplified signal is chopped in a similar manner as the input and in synchronism with the input chopper. The synchronised chopping restores the DC value of the input signal amplified by the AC gain of the amplifier. As the amplifier does not have a DC gain, the effect of the DC offset voltage and the current are eliminated.

The difficulties with the chopper are:

1. When two dissimilar metals are joined, depending on the temperature, small voltages can be generated. The chopper is to be specifically made to reduce these thermally generated voltages.
2. The electromechanical chopper being a mechanical device has a shorter life. Therefore, electronic choppers are preferred.
3. The chopper should not inject any current into the circuit being chopped. If a MOSFET is used, since it has high input impedance, the leakage current will be less.

A series-shunt chopper using two MOSFETs is shown in Fig. 1.61. The chopping signal is fed to the inverter, which drives the two chopper FETs, one on each half of the chopping cycle. The input impedance of the chopper- stabilised amplifier is very high for DC. The series chopper switches the input to the AC-coupled amplifier every half cycle. The series chopper switch is always opened before the shunt switch is closed. Therefore, there is no path to ground.

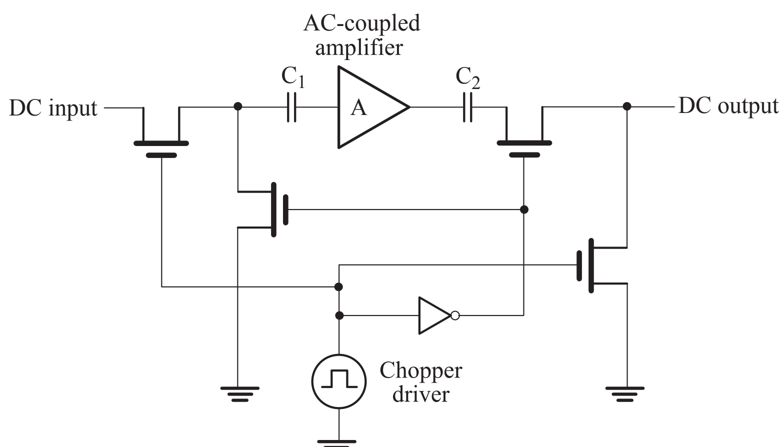


Figure 1.61 All electric chopper circuits use field effect transistors

1.25 AC VOLTMETER USING RECTIFIERS

Electronic AC voltmeters are similar to DC meters, except that AC is rectified before it is applied to the DC meter circuit. Here a good AC amplifier is to be designed. In some cases, rectification is done before amplification. In this case, a good DC amplifier with zero drift characteristics and unity voltage gain are required. This is shown in Fig. 1.62.

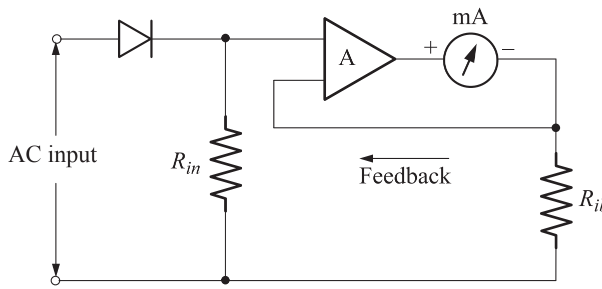
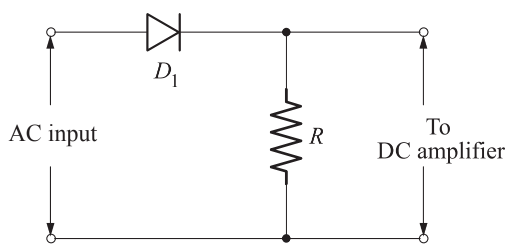


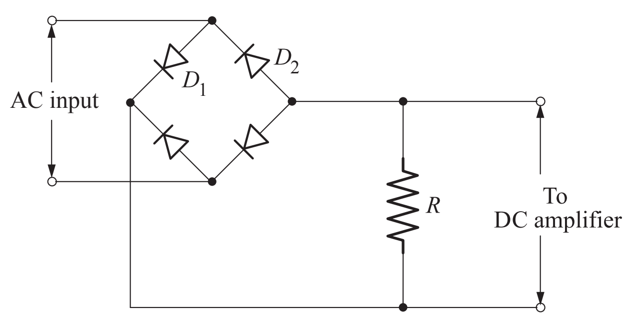
Figure 1.62 DC mode of operation-based AC voltmeter circuits

Figure 1.63 shows the circuit wherein the AC signal is rectified and amplified. The AC amplifier must have high open-loop gain and large amounts of negative feedback to overcome the non-linearity of the rectifier diodes. Different rectifier circuits used in AC meters are shown in Fig. 1.63.

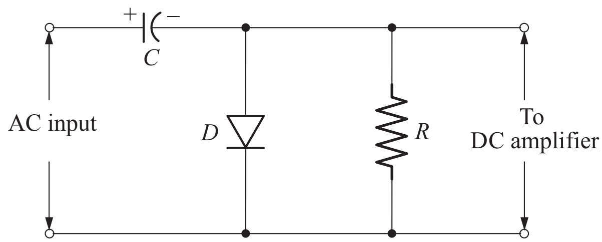
The general form of an AC input is a sine wave (Fig 1.64). Therefore, all the AC instruments are calibrated for the sine wave. For AC measurements, the *rms* value is measured. However, all the AC meters are usually average responding. Therefore, the pointer deflects proportional to the average value of the input but the scale is calibrated in terms of the *rms* value. The average reading is multiplied by the form factor and the scale is calibrated in terms of the *rms* values. If a non-sinusoidal input such as a square wave or a triangular wave is applied to a sine wave meter, there will be error in the measurement.



(a)



(b)



(c)

Figure 1.63 Rectifier circuits used in AC voltmeter: (a) series-connected diode providing half-wave rectification for an average reading meter (b) four diodes in a bridge circuit for full-wave rectification and application to an average reading meter and (c) shunt-connected diode used in a peak reading voltmeter

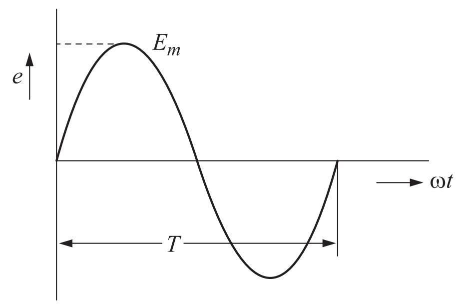


Figure 1.64 Sine wave AC input

Form factor

$$k = \frac{\text{RMS value}}{\text{Average value}}$$

$$\text{RMS value} = \sqrt{\frac{1}{T}\int_0^T e^2\,dt}$$

$$= \sqrt{\frac{1}{T}\int_0^T E_m^1 \sin^2 \omega t\,dt} \;;\; T = \frac{2\pi}{\omega}$$

$$= \sqrt{\left(\frac{\omega}{2\pi}\right)\int_0^{2\pi/\omega} E_m^2 \sin^2 \omega t\,dt}$$

$$= \frac{1}{T/2}\int_0^{T/2} c\,dt$$

$$\text{Average value} = \frac{2}{T}\int_0^{T/2} E_m \sin\omega t\,dt; \qquad T = \frac{2\pi}{\omega}$$

$$= \left(\frac{\omega}{\pi}\right)\int_{0}^{(\pi/\omega)} E_m \sin\omega t\, dt$$

$\therefore$ Form factor

$$k = \frac{\sqrt{\left(\frac{E_m^2}{4\pi}\right)[\omega t - \sin\omega t - \cos\omega t]_0^{2\pi/\omega}}}{\left(\frac{E_m}{\pi}\right)[-\cos\omega t]_0^{\pi/\omega}}$$

$$k = \frac{0.707\ E_m}{0.636\ E_m} = 1.11$$

Therefore, for an average responding voltmeter, the scale reading is to be multiplied by 1.11 to get the *rms* value. For a square wave, the form factor is 1.0, that is, the average and *rms* values are same.

If a square wave is applied to a meter calibrated to read the *rms* values of a sine wave, there will be an error in the measurement.

1.26 TRUE *RMS*-RESPONDING VOLTMETER

To measure the values of complex AC inputs, true *rms*-responding voltmeters are to be used. The heating power of a given input signal is proportional to the square of the *rms* value of the voltage. Sensing this power, this meter produces a deflection using a thermocouple. Thermocouple outputs are non-linear, in general. This difficulty is overcome by placing two thermocouples in the heating environment. The schematic of this meter is shown in Fig. 1.65.

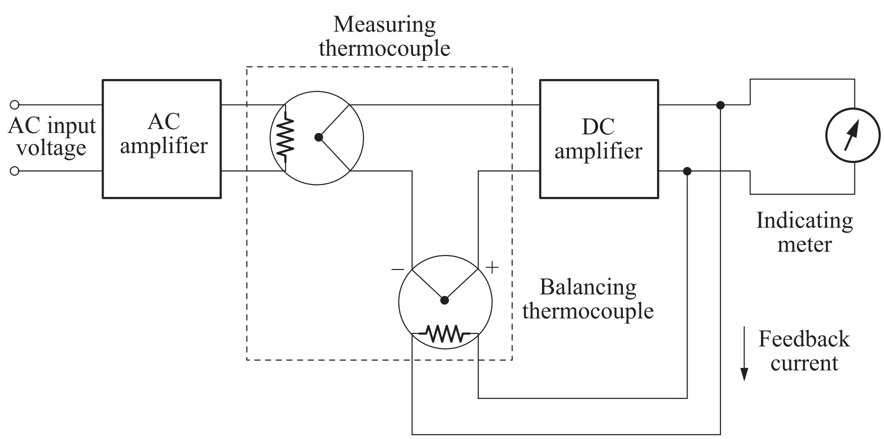


Figure 1.65 Block diagram of a true *rms*-reading voltmeter. The measuring and balancing thermocouples are located in the same thermal environment

The unknown AC input voltage is amplified and applied to the heating element of the measuring thermocouple. The thermocouple in the input side is the measuring thermocouple. The thermocouple in the feedback path is the balancing thermocouple. They are similar thermocouples and form a complementary pair. Therefore, the non-linearity due to the measuring thermocouple is cancelled by the similar non-linearity of the balancing thermocouple. The heater coil gets heated due to the AC input given. The measuring thermocouple produces a voltage, which upsets the balance of the bridge. The imbalance voltage is amplified by the DC amplifier and feedback to the heating element of the balancing thermocouple. Bridge balance is restored when the feedback current delivers sufficient heat to the balancing thermocouple, so that the voltage output of both thermocouples are the same. The DC current in the heating element of the feedback thermocouple is equal to the AC current in the input thermocouple. Therefore, this DC current is directly proportional to the effective or *rms* value of the input voltage. This is indicated on the meter movement in the output circuit of the DC amplifier. Thus, the true *rms* value of the AC input can be measured irrespective of the shape of the input.

$$\text{Crest factor} = \frac{\text{Peak value of input}}{\text{RMS value}}$$

Laboratory-type meters can measure inputs with a crest factor of 10/1.

Voltage range: 100 μV to 300 V.

Frequency range: 10 Hz to 10 MHz.

1.27 BALANCED BRIDGE VOLTMETER (VTVM)

The Vacuum Tube Voltmeter (VTVM) is used to measure both AC and DC voltages. Owing to the disadvantages associated with vacuum tubes, this type of meter is not in use now. The triodes used in the balanced bridge amplifier circuit are replaced by BJTs and JFETs, and such electronic meters are used for measurement.

VTVM consists of the following:

1. Balanced bridge DC amplifier together with a DC meter circuit.
2. Rectifier section to convert AC to DC.
3. Internal battery and circuit to measure resistance.
4. Switching arrangement to select the function of the meter and potential divider for range selection.

The circuit consists of two triode amplifiers, which form the two arms of the bridge. The other two arms of the bridge are two resistors with a zero adjust control. The DC meter indicator is connected to the anodes of the triodes. The input is applied to the grid of one triode through an attenuator. The grid of the other triode is grounded.

VTVM has provision to measure both AC and DC current and also the resistance. The advantages and disadvantages of VTVM are as follows:

1.27.1 Advantages

1. High input impedance of the order of 100 MΩ.
2. Large voltage range from μV to hundreds of V.
3. Large current range.
4. Provision to measure resistance.

1.27.2 Disadvantages

1. Size is bulky due to the usage of vacuum tubes.
2. Time lag for the amplifiers to respond when switched on, due to thermionic emission cathodes.
3. Short life due to thermal wear and tear of thermionic cathodes.
4. Needs large voltage and power to operate.

1.28 TRANSISTOR VOLTMETER (TVM)

VTVMs are not used due to their disadvantages, but the same circuit principle is employed, replacing the vacuum triodes with BJTs. Such a meter is called a Transistor Voltmeter (TVM). The basic bridge circuit configuration used is as shown in Fig. 1.66.

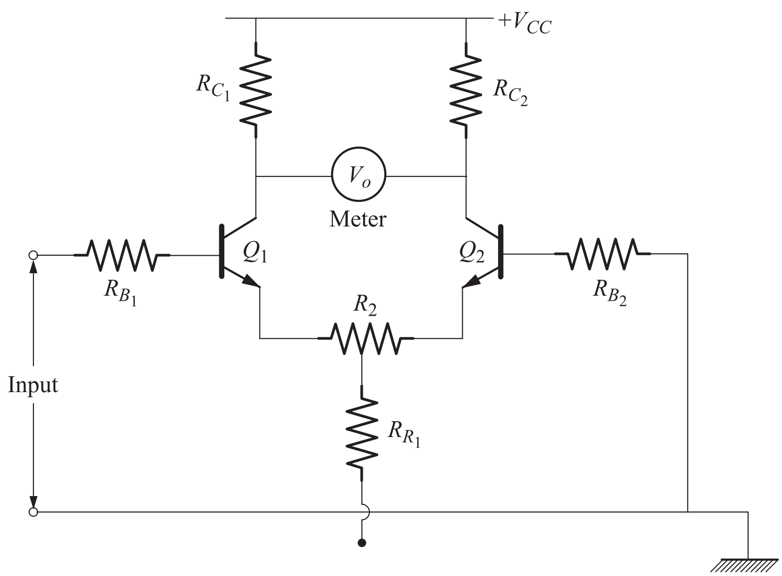


Figure 1.66 TVM bridge circuit

Q_1 and Q_2 form the lower arms of the bridge circuit. R_{C1} and R_{C2} form the upper two arms of the bridge. R_2 is the zero adjust resistor. Base B_2 of Q_2 is grounded through R_{B2}. Input is applied to R_{B1}. Zero adjustment can be done using R_2. The differential output ($V_{C1} - V_{C2}$) is proportional to the input. A potential divider circuit modifies the input to make it suitable to be applied at R_B. Current measurement can also be done.

1.29 ELECTRONIC MULTIMETER

This meter can measure DC and AC voltages and currents as well as resistance. Therefore, these meters are also called VOM meters; that is, Volts, Ohms, Milliampere-measuring instruments. Generally, a multimeter consists of the following:

1. A balanced bridge DC amplifier and an indicating meter.
2. An input attenuator or range switch to limit the magnitude of the input voltage to the desired value.
3. A rectifier section to convert an AC input to a proportional DC value.
4. An internal battery and additional circuitry to measure resistance.
5. A function switch to select various measurement functions of the instrument.

Figure 1.67 shows the circuit diagram of a balanced bridge DC amplifier using JFETs. The circuit operation is similar to the one shown in Fig. 1.66. Instead of BJTs, JFETs are being used here. The two JFETs form the upper arms of a bridge circuit. Resistors $_1$, R_2 together with zero adjust resistor R_3 form the lower bridge arms.

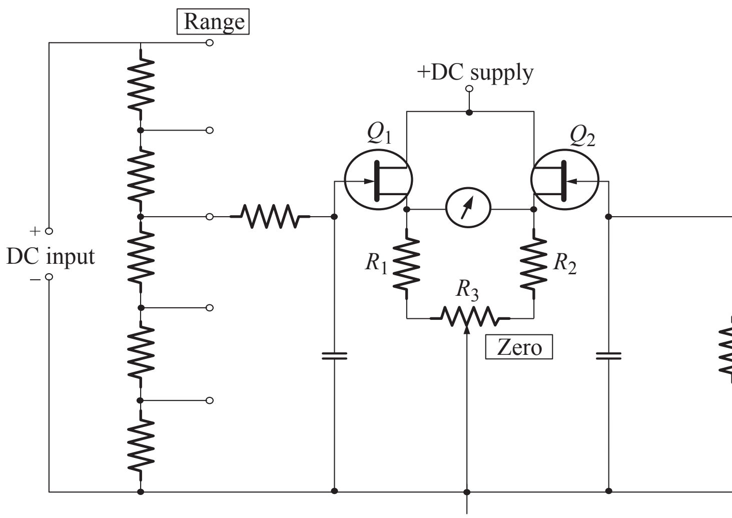


Figure 1.67 Balanced bridge DC amplifier with input attenuator and indicating meter

The meter movement is connected between the source terminals of the FET, representing the two opposite corners of the bridge. The maximum voltage that can be applied to the gate of Q_1 depends on the operating range of the JFET. It will usually be a few volts. The input is usually applied through a range switch. A typical circuit of a range switch is shown in Fig. 1.68.

1.29.1 Resistance Ranges

When the selector switch of the multimeter is put in the resistance range, the unknown resistor is connected in series with an internal battery. The voltage drop across the resistor is measured and it is proportional to the value of the unknown resistor. The meter is calibrated in terms of the resistance. The circuit diagram is shown in Fig. 1.69.

The unknown resistor R_x is connected to the two terminals and the range switch is kept in the ohms position. The 1.5 V internal battery supplies current through one of the range resistors and the unknown resistor to the ground. A voltage drop V_x across R_x is applied to the input of the bridge amplifier and causes a deflection on the meter. The meter is calibrated in terms of the resistance values. If the resistance value is more, the voltage drop will be more; therefore, the meter deflection is also more. Hence, in electronic multimeters, the resistance values increase from left to right similar to

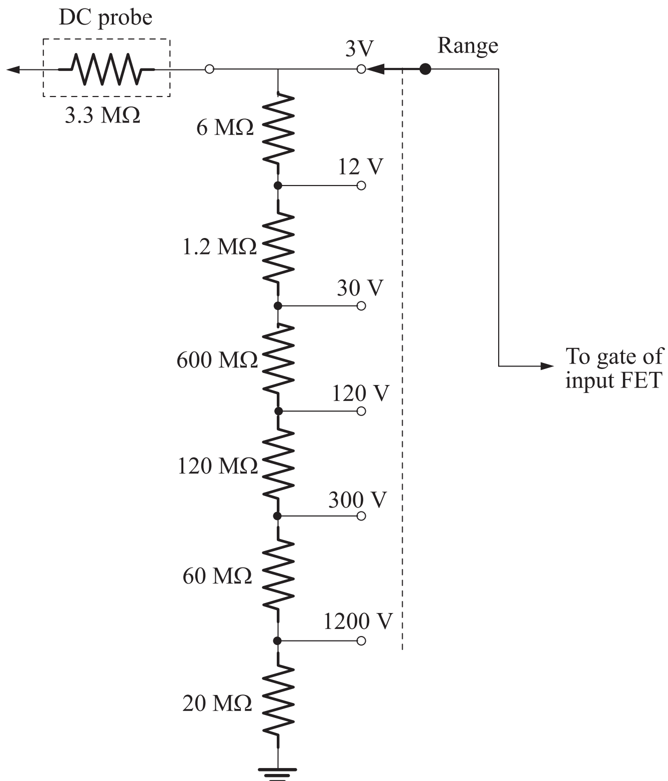


Figure 1.68 Typical input voltage attenuator for a VOM. The range switch on the front of the panel of the VOM allows selection of the desired voltage range

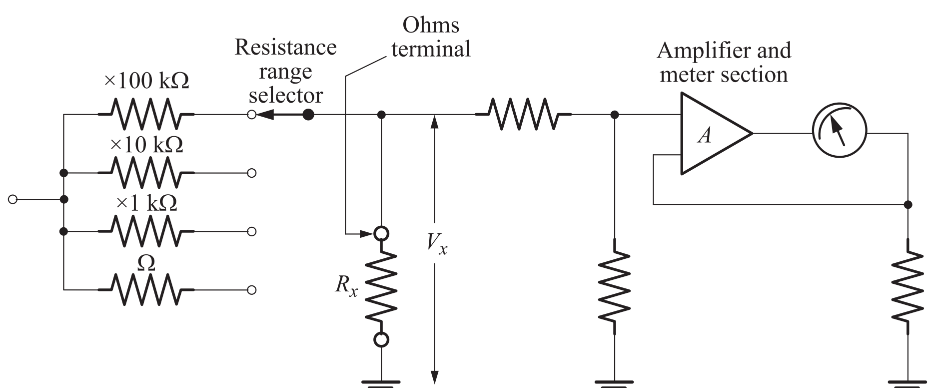


Figure 1.69 Resistance range selector circuit of a VOM

voltage and current scales. In conventional multimeters, a current meter is used for deflection. The current is inversely proportional to resistance. Therefore, the meter deflection is minimum when the value of the resistance is large. Hence, in such meters, the resistance values decrease from left to right on the scale. Figure 1.70 shows the simplified metering circuit of an electronic VOM.

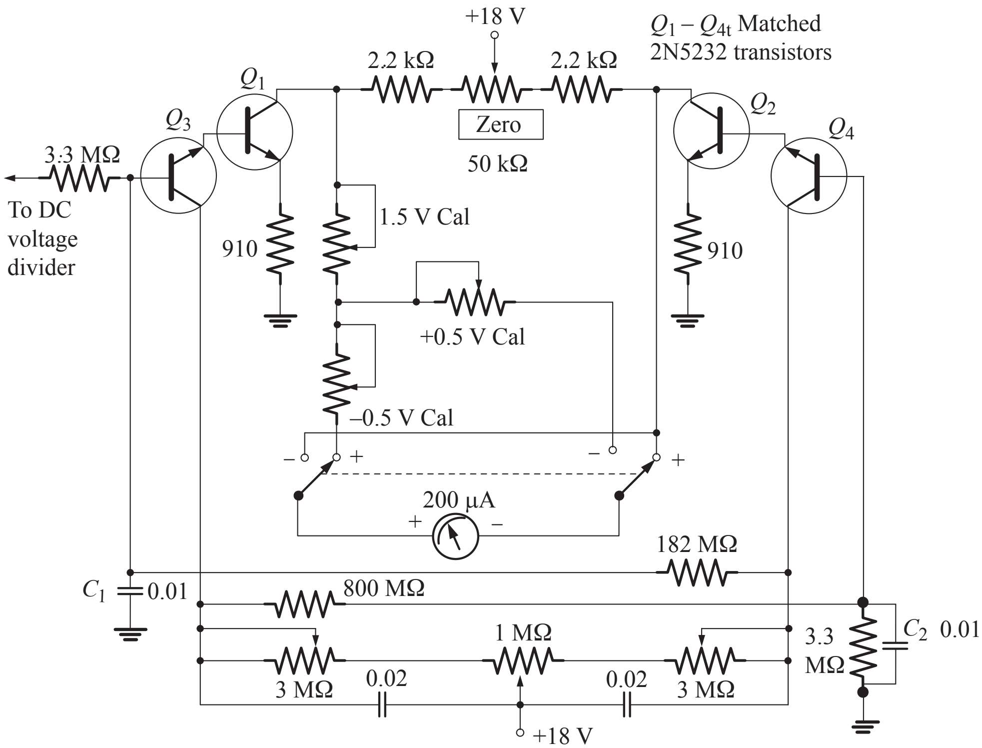


Figure 1.70 Typical metering circuit of a solid-state VOM

1.30 AC CURRENT MEASUREMENT

The current to be measured is passed through a series resistor of suitable value in the current scale. The voltage drop across the resistor can be measured by the voltmeter and the meter deflection is calibrated in terms of the current. In some cases, an AC current probe is also employed. The probe clips the wire under test, without disturbing the circuit. The curve through which current is passing and which is to be measured forms the one-turn primary of a transformer with a finite core and a many-turn secondary within the current probe body. The signal induced in the secondary winding is amplified, and the output voltage of the amplifier is applied to a suitable AC voltmeter for measurement. The amplifier is designed so that 1 mA in the wire being measured produces 1 mV at the amplifier output. The current is read on the voltmeter, which is calibrated in mA or amps.

1.30.1 Differential Voltmeter

It is one of the most accurate means of measuring an unknown voltage. Here the unknown voltage is compared to a known voltage. It is also called *potentiometric voltmeter,* since the principle of operation is similar to a potentiometer (see Fig. 1.71).

A *precision resistance-divider* network is used to divide down an accurately known reference voltage. The divider is adjusted until the output voltage equals the unknown voltage. The multimeter indication is proportional to the difference of the potentials between the reference source and the unknown voltage

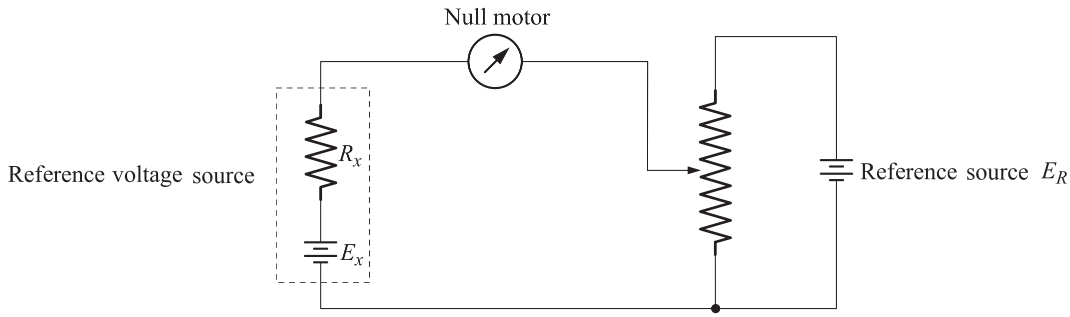


Figure 1.71 Differential voltmeter

source. To detect small differences in unbalanced potentials, a sensitive meter movement is required. The reference source consists of a low-voltage DC standard such as a 1 V DC supply on low-voltage Zener-controlled precision supply.

To measure high voltages, a high-voltage reference supply can be used. However, normally a potential divider network is used to reduce the unknown voltage supply to a sufficiently low value for direct comparison against a low-voltage DC standard (Fig. 1.72).

The AC differential voltmeter is a modification of DC differential voltmeter.

The AC to DC converter gives an average value of the voltage. Therefore, this voltmeter gives an average value of the AC input. The meter can be calibrated to give the *rms* value. The scale factor is 1.11.

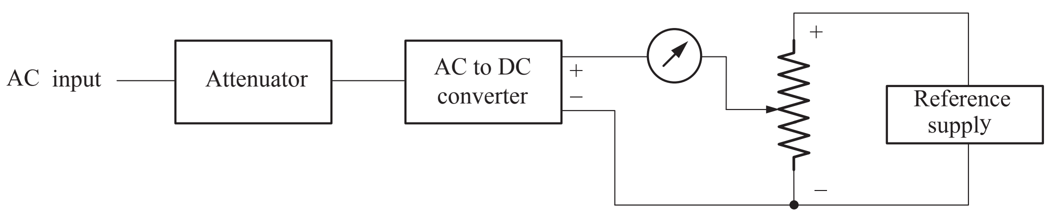


Figure 1.72 AC Differential voltmeter

Example 1.15

Design a range switch for the DC volt section of a balanced-bridge VTVM or (FET input VM). The total resistance of the attenuator should be 11 MΩ. The attenuator should be so arranged that input voltages from 3 to 1000 V can be accommodated in the customary 1–3–10 sequence (gate of FET). The bridge circuit requires 1 V at the gate of the balancing circuit to cause full-scale meter deflection.

Solution

Consider Fig. 1.73. The input to the grid of amplifiers should be only 1 V. Therefore, when the selector switch is at the 1000 V position, and the mass input is 1000 V, the drop across R_7 should be 1 value.

$$\therefore\ R_{\text{Total}} = R_1 + R_2 + \cdots + R_7 = 11\ \text{m}\Omega$$

$$\therefore\ R_7 = \left(\frac{1\ \text{V}}{1000\ \text{V}}\right) \times \text{m}\Omega = 11\ \text{k}\Omega$$

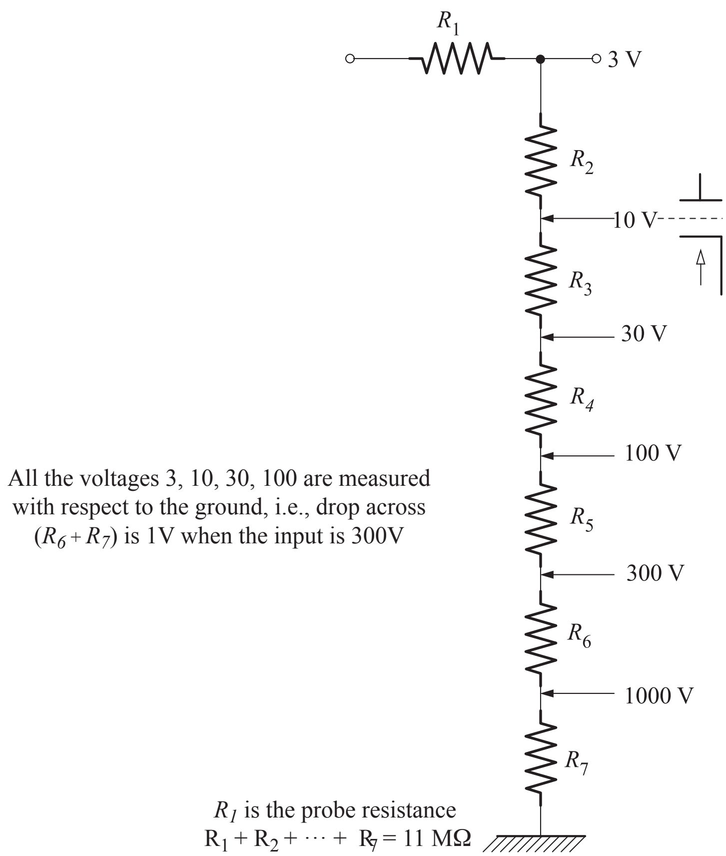


Figure 1.73 For Example 1.15

Similarly, when the selector switch is at the 300 V position, the drop across $(R_6 + R_7)$ should be 1 V.

$$R_6 = \left(\frac{1}{300} \times 11\ \text{m}\Omega\right) - 11\ \text{k}\Omega = 36.67 - 11 = 25.67\ \text{k}\Omega.$$

Similarly,

$$R_5 = \left(\frac{1}{100} \times 11\ \text{m}\Omega\right) - (R_6 + R_7) = 110 - 36.67 = 73.33\ \text{k}\Omega$$

$$R_4 = \left(\frac{1}{30} \times 11\ \text{m}\Omega\right) - 110\ \text{k}\Omega = 256.7\ \text{k}\Omega$$

$$R_3 = \left(\frac{1}{10} \times 11\ \text{m}\Omega\right) - 366.7\ \text{k}\Omega = 1100 - 366.7 = 733.3\ \text{k}\Omega$$

$$R_2 = \left(\frac{1}{3} \times 11\ \text{m}\Omega\right) - 1.1\ \text{m}\Omega = 2.57\ \text{m}\Omega$$

$$R_1 = 11\ \text{m}\Omega - 3.67\ \text{m}\Omega = 7.53\ \text{m}\Omega$$

Example 1.16

A differential voltmeter uses a reference source with an internal resistance of 200 Ω and a terminal voltage of 3.0 V. The galvanometer has a current sensitivity of 1 mm/μA and an internal resistance of 100 Ω.

(a) Calculate the *emf* of the unknown source, neglecting its internal resistance, if the galvanometer deflection is 250 mm.
(b) Calculate the resolution of the measurement setup if the galvanometer deflection can be read to 1 mm.
(c) For the differential voltage measurements in the above problem, a second galvanometer with a current sensitivity of 5 mm/μA and an internal resistance of 1000 Ω is available. Calculate which of the two galvanometers provides the greatest sensitivity to imbalance, expressing the result in millimeter per millivolt.

Solution

(a) Consider Fig. 1.74

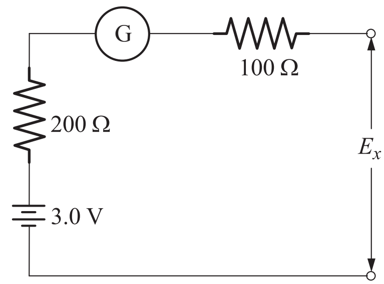


Figure 1.74 For Example 1.16

Sensitivity = 1 mm/μA
Deflection = 250 mm

∴ Current through the galvanometer = 250 μA

$$R_{total} = 200 + 100 = 300\ \Omega$$

$$\text{Voltage drop across } R_{total} = 300 \times 250 \times 10^{-6} = 75 \text{ mV}$$

$$\therefore\ E_k = 3.0 - 75 \text{ mV} = 2.925 \text{ V}$$

(b) 1 mm corresponds to 1 μA

$$\text{Resolution} = 1\ \mu\text{A} \times 300\ \Omega = 300\ \mu\text{V/mm}$$

(c) Galvanometer A: 1 mm deflection for 300 mV.

$$\text{Sensitivity} = 1/300 \text{ mm/mV}$$

Galvanometer B:

$$R_{total} = 1000 + 200 = 1200\ \Omega$$

A 5 mm deflection is caused by 1 μA × 1200 Ω or 1200 μV.

$$\text{Sensitivity} = 5 \text{ mm}/1200\ \mu\text{V} = 1/1240 \text{ mm}/\mu\text{V}$$

$\therefore$ Galvanometer B provides amplifier sensitivity.

$$\therefore \frac{1}{240} > \frac{1}{300}$$

1.31 DIFFERENTIAL AMPLIFIER

The temperature drift terms are major limits to amplifier sensitivity. The method of compensating these is to balance out these effects with an equal but opposite drift signal. This is the principle of operation of the differential amplifier shown in Fig. 1.75. If a signal is applied to the input terminals of Q_1 and Q_2 it appears as an amplified signal at the collector of Q_1 and Q_2. Kirchhoff's Voltage Law loop equations are written for the two input signals V_{i1} and V_{i2} as follows.

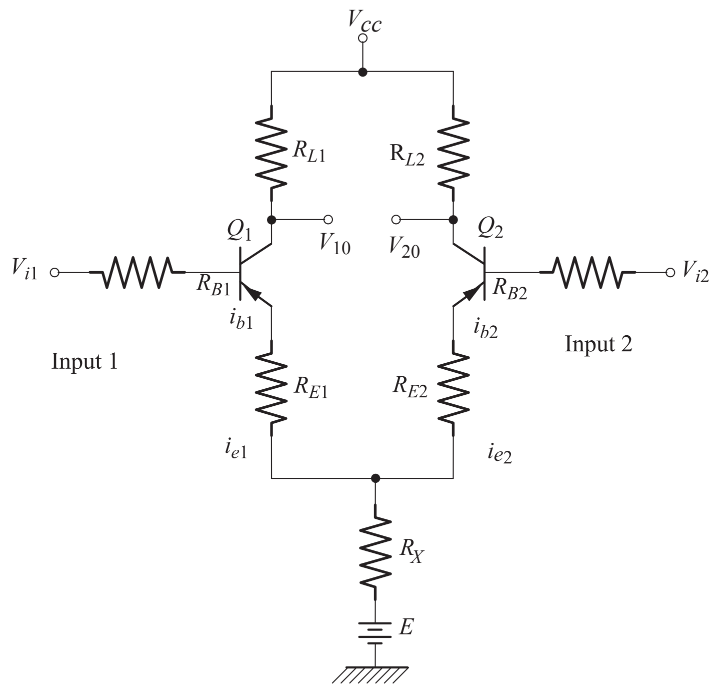


Figure 1.75 Differential amplifier

$$V_{i_1} - V_{Be1} + V_{EE} = R_{B1}I_{B_1} + (R_{E1} + R_X)\, I_{e1} + R_X I_{e2}$$

$$V_{i_2} - V_{Be_2} + V_{EE} = R_{B2}I_{B_2} + R_X I_{e_1} + (R_{E_2} + R_X)\, I_{e_2}$$

In addition,

$$I_{e_1} = I_{B_1} + \alpha_1 I_{e_1} + I_{co_1}$$

$$I_{e_2} = I_{B_2} + \alpha_2 I_{e_2} + I_{co_2}$$

The output voltage

$$V_{10} = V_{cc} - (\alpha_1 I_{e1} + I_{co1})\, R_{L1}$$
$$V_{20} = V_{cc} - (\alpha_2 I_{e2} + I_{co2})\, R_{L2}$$

After considerable algebraic manipulation, the gain of the differential amplifier is

$$A_{dm} = \frac{V_{10} - V_{20}}{V_{i_1} - V_{i_2}} = \frac{2\alpha R_L}{R_e + \left(\dfrac{R_B}{\beta}\right) + (R_B^2 / 2B^2 R_X)}$$

If the drift voltages are equal, they will get cancelled and will not be applied in the output. Operation with identical signals is called the *common mode*. A measure of how a differential amplifier cancels undesirable signals is called the *common mode rejection.* The parameter is called CMRR.

$$CMRR = \frac{Ad_m}{Ac_m} = \frac{2R_X \beta_1 \beta_2}{R_B(\beta_1 - \beta_2)}$$

It is expressed in decibels (dB).

Thus, in the ratio of differential gain to common mode gain, R_{B_1} and R_{B_2} are to limit the base currents. By means of R_X, we can make the output zero at ambient temperature, with no input.

1.32 ALTERNATING CURRENT INSTRUMENTS (AC METERS)

Various AC instruments are as given in Table 1.4.

Table 1.4 Different types of AC instruments

Types of Meter Movement	Use	
	DC	AC
Electrodynamometer	Yes	Yes
Iron vane	✓	✓
Electrostatic (Gold leaf out-dated)	✓	✓
Thermocouple	✓	✓
D'Arsonval	✓	With rectifiers

1.32.1 D'Arsonval Meter Movement for AC Circuit

If you are applying 10 V DC to the D'Arsonval meter (see Fig. 1.76(a))

$$I_{fs} \quad \frac{10\text{ V}}{10\text{ k}\Omega} \quad 1\text{ mA}$$

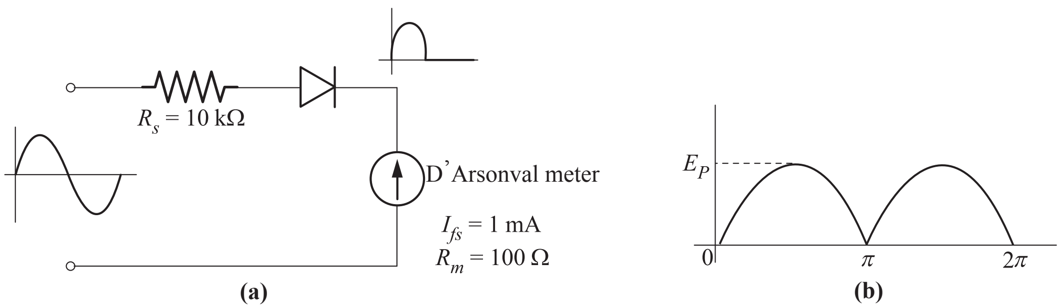


Figure 1.76 (a) D'Arsonval meter and (b) $E_{av} = 0.636\ E_p$ for FWR

Therefore, sensitivity

$$S = \frac{1}{I_{fs}} = \frac{1}{1\text{ mA}}$$
$$= 1\text{ kW/V}$$

If you apply 10 V AC current, 10 V being *rms* values for the above circuit, AC is half-wave rectified (HWR) by the diode (for AC, the value given is *rms*)

Peak value of the 10 V_{rms} sine wave E_p

$$E_p = 10\ V_{rms} \times 1.414 = 14.14\text{ V}$$
$$\left(= \sqrt{2}E_{rms}\right)$$

The DC meter movement will respond to the average value of the FWR signal:

$$\therefore\ E_{av} = E_p \times 0.636$$
$$= 14.14 \times 0.636$$
$$= 8.99\text{ V}$$

For a HWR circuit, the average value is half of this value.

$$E_{av}\text{ for HWR signal} = 0.318\ E_p$$
$$= 0.318 \times 14.14$$
$$= \frac{8.99}{2} = 4.5\text{ V}$$

If the meter reads 10 V for DC, and if we apply 10 V_{rms} to the HWR circuit, the meter reads 4.5 V.

Therefore, the sensitivity of the meter is only 45% of the sensitivity of the DC meter (HWR circuit).

To determine the multiplier resistance,

$$R_s = \frac{E_{dc}}{I_{dc}} - R_m = \frac{0.45E_{rms}}{I_{dc}} - R_m$$
$$S_{ac} = 0.45\ S_{dc}$$
$$S = \text{Sensitivity}$$

Example 1.17

Determine the value of the multiplier resistance for 10 V_{rms} AC range.

Solution

Consider Fig. 1.77.

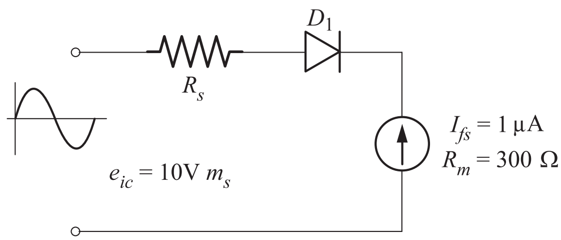


Figure 1.77 For Example 1.17

$$R_s = \frac{0.45 E_{rms}}{I_{de}} - R_m$$

$$= \frac{0.45 \times 10}{1\ \text{mA}} - 300\ \Omega$$

$$= 4.5\ \text{k}\Omega - 300\ \Omega = 4.2\ \text{k}\Omega$$

1.32.2 Modified Circuit for AC Measurements

Two diodes D_1, D_2 are employed in the circuit itself as shown in Fig. 1.78.

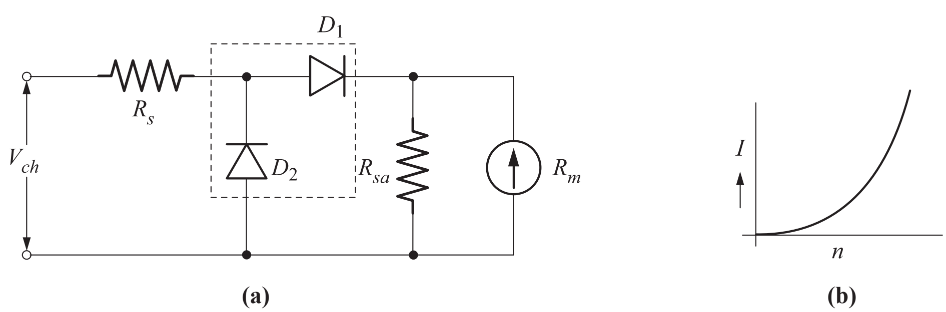


Figure 1.78 D'Arsonval meter with HWR circuit

D_1 is the diode for half wave rectification.

D_2 is reverse biased in the positive half cycle and forward biased in the negative half cycle. Therefore, the reverse leakage current of D_1 will flow through D_2 in the negative half cycle, instead of the meter. R_f of $D_2 < R_m$. Hence, the errors in the measurement circuit get reduced. As the reverse saturation current is not flowing through the meter, the meter responds to the average value and the error in the average value reduces.

The purpose of R_{sg} is to prove the linearity of the circuit. The diode I–V is characteristic non-linear (it is exponential in nature). Therefore, the calibration of the meter will also be non-linear. To make it, if R_{sg} is connected, the net resistance seen by the circuit is $R_m||R_{sg}$, which is less than R_m. Hence linearity increases

$$\therefore \; I_m = \frac{E_w}{R_s + (R_{sg} \,||\, R_m)} < \frac{E_w}{R_s + R_m}$$

Therefore, more current flows through the meter. Hence, linearity improves.

1.32.3 D'Arsonval Meter Movement Circuit (FWR)

Consider Fig. 1.79

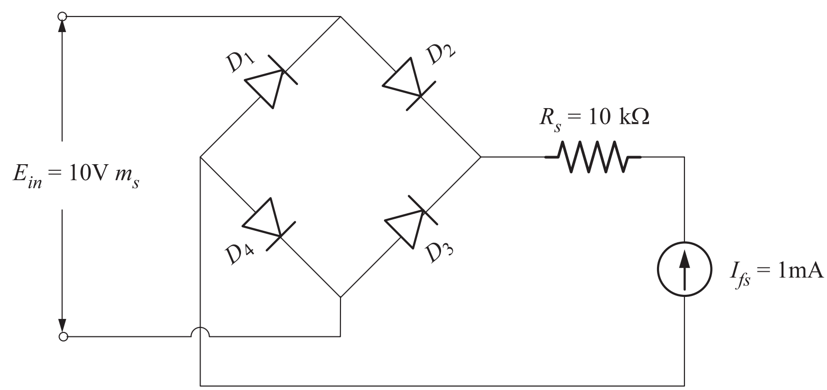


Figure 1.79 D'Arsonval meter with bridge circuit

$$E_p = 1.414\, E_{rms}$$

If

$$E_{rms} = 10 \text{ V}, \; E_p = 1.414 \times 10 = 14.14 \text{ V}$$

$$E_{qv} = 0.636\, E_p = qV$$

Therefore, if you give 10 V_{rms} AC it reads as equivalent to 9 V DC (average value is equivalent to DC).

Therefore, sensitivity with the FWR circuit will be 90% of S_{dc}

$$S_{FWR} = 0.9\, S_{dc}$$

Example 1.18

For a given meter, I_{fs} = 1 mA, R_m = 500 Ω, E_m = 10V_{rms}. Determine the value of R_s (Fig. 1.80).

$$I_{fs} = 1 \text{ mA}, \quad R_m = 500\ \Omega, \quad E_m = 10V_{rms}, \quad R_s = ?$$

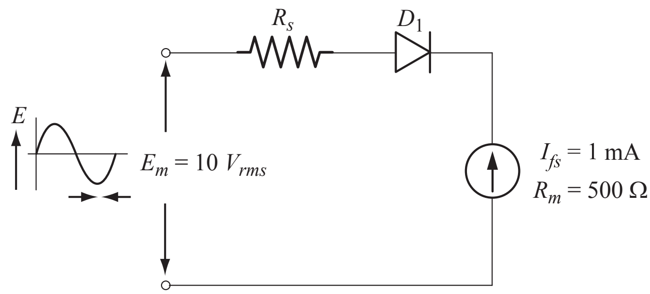


Figure 1.80 For Example 1.18

Solution

$$S_{dc} = \frac{1}{I_{fs}} = 1\ \text{k}\Omega/\text{V}$$

$$S_{ac} = 0.9\ S_{dc} = 900\ \Omega/\text{V}$$

$$R_s = S_{ac} \times \text{Range} - R_{ms}$$

$$= \frac{900\ \Omega}{V} \times 10\ V_{rms} - 500\ \Omega$$

$$= 8.5\ \text{k}\Omega$$

1.33 ELECTRODYNAMOMETER MOVEMENT

This type of deflecting system is used for standard voltmeters, ammeters, and also transfer instruments.

Standard meters: As the name implies, these meters are used as a standard for calibrations.

1.33.1 Transfer Instruments

These are the instruments calibrated for DC and are used to measure AC quantities to determine the errors or change in sensitivity.

The electrodynamometer deflection system consists of two fixed coils and a moving coil. All these coils are connected in series as shown in Fig. 1.81.

The magnetic field setup in the fixed coil interacts with the magnetic field setup in the moving coil and the moving coil deflects. This is the principle of measurement.

The current handling capacity of an electrodynamometer-type movement is greater than the D'Arsonval meter movement as the magnetic field setup here is weak. Since air is the coupling medium for magnetic field, more current is required to cause the same deflection. Therefore, a larger diameter coil must be used here as more current-handling capacity is required.

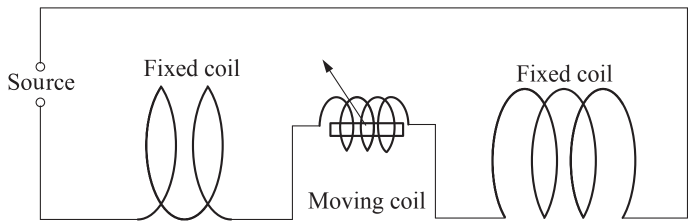


Figure 1.81 Electrodynamometer movement

$$\therefore \quad S = \frac{1}{I_f}, \quad S = \text{Sensitivity}$$

S of the electrodynamometer type is less than that of the D'Arsonval type and is typically 20–100 Ω/V.

1.33.2 Iron Vane-Meter Movement

The iron vane meter consists of a fixed coil of many turns, and two iron vanes are placed inside the fixed coil. It is rugged and the accuracy is more for AC than DC. It is frequency sensitive. Hence, it is used in the range 25–125 Hz. The current passing through the coils sets up a magnetic field. This field magnetizes the two iron vanes with the same polarity. Hence, the two iron vanes will repel. One of the iron vanes is fixed. Therefore, the other vane repels and the coil fixed to the movable vane deflects. This is the principle of the instrument.

Deflection is proportional to I^2 (square of the current passing through the coil).

Although this instrument can be used for DC applications, hysteresis or magnetic lag causes errors. Therefore, its application is confined to AC circuits.

1.34 THERMOCOUPLE METER

This setup (see Fig. 1.82) consists of a heater element of a nichrome wire, a thermocouple, and D'Arsonval meter.

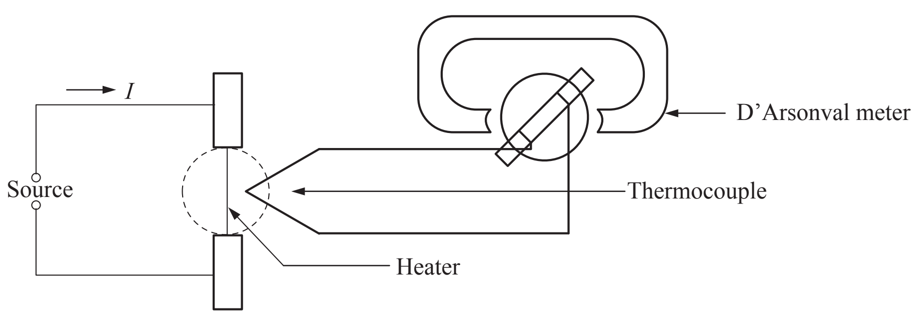


Figure 1.82 Thermocouple instruments

This can be used for both AC and DC. For AC, it can be used over wide frequency ranges even well above 50 MHzs. There is no other deflecting instrument type that can measure current up to that frequency.

The temperature T to which the wire is heated is proportional to the current passing through it.

The thermocouple output V_T is proportional to T.

The D'Arsonval meter deflection is proportional to V_T.

Therefore, frequency has no effect on the deflecting mechanism due to L or C.

The Thermocouple and the meter are thermally connected but electrically separated.

Example 1.19

Design a thermocouple voltmeter for three ranges 5, 10, and 25 V, given that, for D'Arsonval Meter,

$$I_{fs} = 50 \text{ mA}$$
$$R_m = 200\ \Omega$$

For a heater,

$$I_{\max} = 5 \text{ mA}$$

Resistance of heater (R) $= 200\ \Omega$.

Solution

To get a FSD for 5, 10, and 25 V, current through the heater must be limited to 5 mA.

5–V range:

Series resistance

$$R_s = \frac{E}{I_{max}} - R_{meter}$$
$$= \frac{5 \text{ V}}{5 \text{ mA}} - 200\ \Omega = 1\text{ k}\Omega - 200\ \Omega$$
$$= 800\ \Omega$$

10–V range:

$$R_s = \frac{E}{I_{max}} - R_m = \frac{10 \text{ V}}{5 \text{ mA}} - 200\ \Omega$$
$$= 2\text{ k}\Omega - 200\ \Omega = 1.8\text{ k}\Omega$$

25–V range:

$$R_s = \frac{E}{I_{max}} - R_{meter} = \frac{25 \text{ V}}{5 \text{ mA}} - 200\ \Omega = 4.8\ \Omega$$

Therefore, the circuit is as shown in Fig. 1.83.

1.34.1 Constant Voltage Source

A source that maintains the voltage across its output terminals constant irrespective of the current drawn from it is called a constant voltage source (Fig. 1.84). Example: Zener diode circuit.

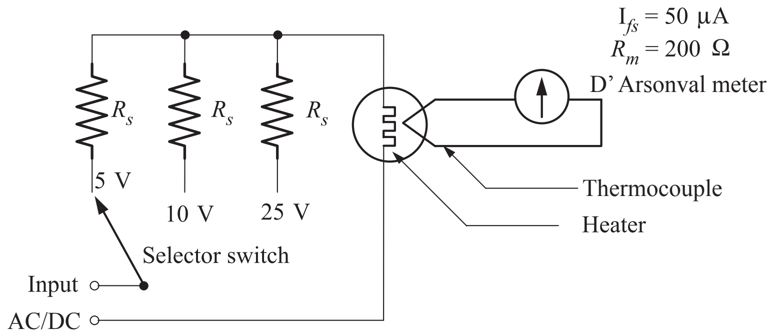


Figure 1.83 For Example 1.19

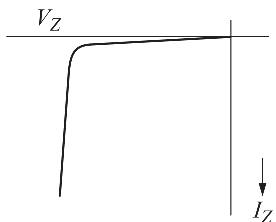


Figure 1.84 Zener reverse characteristics

1.34.2 Constant Current Source

When the current drawn from the source remains constant irrespective of the voltage across the terminals it is called a constant current source (Fig. 1.85). Example: BJT.

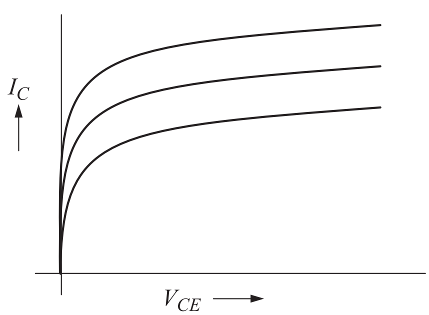


Figure 1.85 BJT output characteristics

Example 1.20

Design a standard DC voltage source of 12 V when the supply voltage is 20 V, using a Zener diode. Choose a Zener diode with a breakdown voltage of 12 V. Let the current through Zener I_z be 10 mA at the breakdown voltage of V_z = 12 V.

Solution

The circuit is shown in Fig. 1.86.

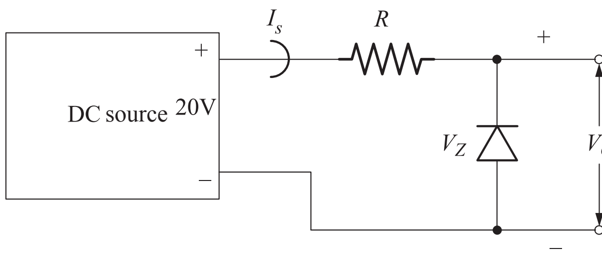


Figure 1.86 For Example 1.20

$$V_o = 12\text{ V}, \qquad V_s = 20\text{ V}, \quad I_z = 10\text{ mA}$$

$$\therefore \quad R = \frac{V_s - V_o}{I_z} = \frac{20 - 12\text{ V}}{10\text{ mA}} = \frac{8\text{ V}}{10\text{ mA}} = 800\ \Omega$$

Power rating of Zener :

$$V_{max} I_{max} = 12\ V \times 10\text{ mA} = 120\text{ mW}$$

$$120 \times 2 \simeq 250\text{ mW}$$

Factor 2 is the safety factor. The power rating of a resistor taking a safety factor of 2 is

$$P_{(Walton)} = \frac{2(V_{DC} - V_z)^2}{R} = \frac{2(20-12)}{800\ \Omega}$$

$$= \frac{16}{800} = \frac{1}{50} = 0.02\text{ W}$$

$\therefore \ \frac{1}{16}$ W resistor curve serves the purpose.

A JFET circuit is also used as a voltage reference source. The circuit is as shown in Fig. 1.87.

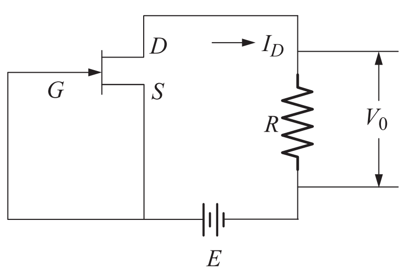


Figure 1.87 JFET circuit as voltage reference source

1.34.3 Volt Box

It is a potential divider network (see Fig. 1.88) with one resistance R_1 being large and the other R_2 being small. A potentiometer is connected across R_2 and the variable output is taken. The input is a large voltage.

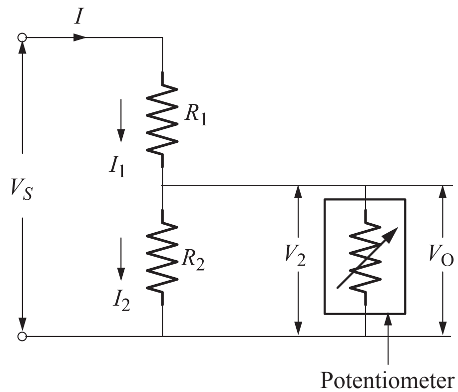


Figure 1.88 Volt box (potential divider network)

$$I = \frac{V_s}{R_1 + R_2}$$

$$R_1 >> R_2$$

$$\therefore \quad I \simeq \frac{V_s}{R_1} \simeq I_1 \simeq I_2$$

It is assured that the potentiometer draws no current. Therefore, R_2 is small, $I_1 \simeq I_2$.

Example 1.21

Design a volt box such that when a supply voltage of 100 V is applied to the input terminals, an output voltage of 5 V is available at the volt box terminals. It is desired that $(R_1 + R_2)$ must be 10 MΩ.

Solution

$$\frac{R_2}{R_1 + R_2} = \frac{V_2}{V_s} = \frac{5}{100} = \frac{1}{20}$$

$$\frac{R_1 + R_2}{R_2} = 20; \quad (R_1 + R_2) = 10\ \text{M}\Omega$$

$$\therefore R_2 = \frac{10\ \text{M}\Omega}{20} = 50\ \text{k}\Omega = 50\ \text{k}\Omega; \qquad R_v = 9.95\ \text{M}\Omega$$

1.34.4 Factors to be Considered in the Selection of an Analog Voltmeter

The type of voltmeter to be chosen depends on the measurement to be made, application, and performance. Some factors to be considered in choosing a voltmeter are given below:

1. Input impedance of the meter: 10 M or in that range.
2. Voltage ranges: 1–3–10 sequence or 1.5–5–15 sequence mV, Volts, 10s of V, 100s of V.

3. Decibels: For wide range of voltages.
4. Sensitivity: 1 mV, 100 μV.
5. Bandwidth: 10 Hz to 10 MHz.
6. Battery operation.

Example 1.22

If a symmetrical square wave is applied to an average responding AC voltmeter calibrated in terms of the *rms* value of a sine wave, calculate the percentage error in the meter indication.

The *rms* value of a square wave

Solution

$$E_{rms} = \sqrt{\frac{1}{T}\int_0^T e^2 dt} = E_m$$

The average value, $$E_{av} = \frac{2}{T}\int_0^{T/2} edt = E_m$$

$$\text{Form factor} = \frac{E_{rms}}{E_{aV}} = \frac{E_m}{E_m} = 1.0$$

Form factor of a sine wave = 1.11.

Since the meter is calibrated for a sine wave, for a 1.0 V *rms* value of square wave, it indicates 1.11 V.

$$\therefore \quad \%\ \text{error} = \frac{1.11 - 1.0}{1.0} \times 100$$

$$\%\ \text{error} = 11\%$$

1.35 DIGITAL VOLTMETERS

In the previous units, analog meters for the measurement of *V, I,* and *R* are described. Some of these instruments may use electronic amplifier circuits, but the deflecting system is analog in nature. Therefore, even though some of the instruments are electronic meters, they have the disadvantages of analog instruments. With digital display, some of the disadvantages associated with analog instruments such as reading with parallax error can be eliminated. Digital instruments can be μP based and can be made compatible with computers. In this section, such DVM principles are explained. Commercial DVMs will have the facility to measure current and resistance also. Such instruments are called digital multimeters (DMMs). With the advancement in IC technology, the size and cost have reduced, and versatility and reliability have improved. The power requirement of these instruments has also reduced considerably during recent times.

1.35.1 General Specifications

Some of the general specifications and typical values of instruments are given below. The numerical figures given may vary from instrument to instrument, but they give an idea about the range of the values.

1. Input range: From ±1.000000 to ±1000.000 V with an automatic range selection and overload indication.
2. Absolute accuracy: ±0.005%.
3. Stability: Short term: 0.002% for 24 hr.
4. Long term: 0.008% for 6 months.
5. Resolution: 1 μV on 1 V input range.
6. Input impedance: R = 10 MΩ; C = 40 pF.
7. Output signal: Compatibility with computer, printer, or recorder.

Broadly, there are four different types of DVMs:

(i) Ramp-type DVM.
(ii) Integrating-type DVM.
(iii) Continuous balance DVM.
(iv) Successive approximation DVM.

1.36 RAMP-TYPE DVM

Principle: Voltage is converted into time, and the time period is measured with an electronic counter. The reading is displayed as a voltage after processing. The time taken by a linear ramp voltage to rise from 0 V to the level of the input voltage or to decrease from the level of the input voltage to zero is measured with a counter. This time interval is proportional to the voltage to be measured. The waveforms shown in Fig. 1.89 indicate the method. At the start of the measurement cycle, a ramp voltage is initiated. This voltage can be positive going or negative going.

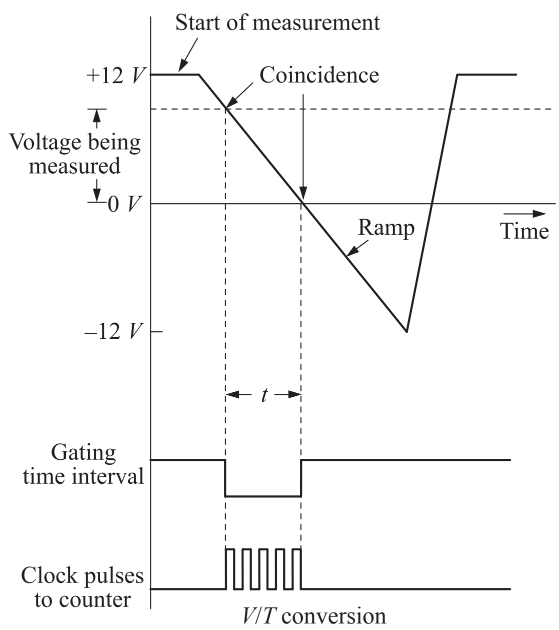


Figure 1.89 Ramp-type voltmeter-graphs

In the figure, a negative-going ramp is shown. It is continuously compared with the unknown input voltage. At the instant that the ramp voltage equals the unknown voltage, a comparator or coincidence circuit generates a pulse. This pulse opens a gate. The ramp voltage continues to decrease with time until it finally reaches 0 V. A second comparator generates an output pulse that closes the gate. An oscillator generates clock pulses that are allowed to pass through the gate to a number of decade-counting units (DCUs). They totalise the number of pulses passed through the gate. The decimal number displayed by the indicator tubes associated with DCUs is a measure of the magnitude of the input voltage. A block schematic of the ramp-type DVM is shown in Fig. 1.90.

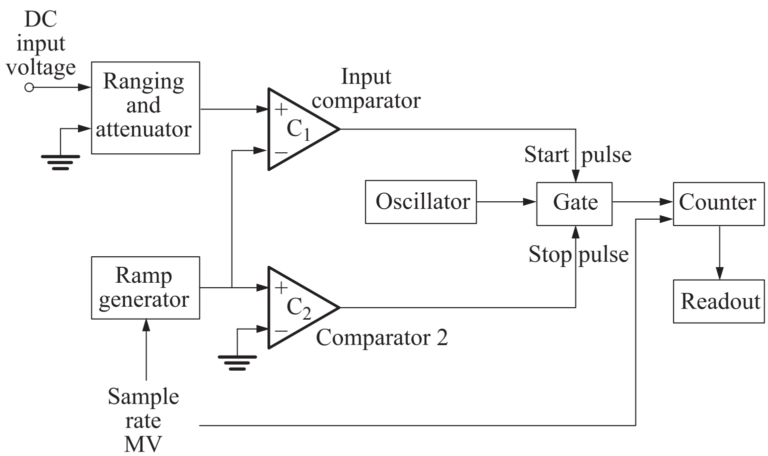


Figure 1.90 Block diagram of a ramp-type DVM

The DC input voltage to be measured is first given to the ranging and attenuator section. If the magnitude of the voltage is large, it is attenuated. If it is small, it is amplified. The sample rate multivibrator determines the slope of the ramp. It also determines the period of measurement or the duration of the measurement cycle, which can be in the range from a few cycles/sec to 1000 cycles/sec. The sample rate circuit provides an initiating pulse for the ramp generator to start its next ramp voltage. At the same time, a reset pulse is generated, which returns all the DCUs to their zero state. The measuring cycle involves

1. Sampling the input and holding it at the same value.
2. Measuring the value of the input.
3. Display.

The display remains at the precision value till the next measuring cycle is completed so that there is no flicker in the display and the user finds continuous display. The ramp voltage is compared with the input voltage and when the equality is reached, a start pulse is applied to the logic gate. If it is the AND gate, it is enabled. The pulses from the clock oscillator are counted in the counter. When the ramp voltage becomes zero, the ground comparator senses the same and the AND gate is disabled. The counting will stop.

The number of pulses counted during this time interval is a measure of the input voltage. These counts are totalised and indicated as voltage in the readout.

1.37 STAIRCASE RAMP-TYPE DVM

This is a modification of the ramp-type DVM. Its features are:

1. Simpler overall design.
2. Low cost.

Owing to these reasons, it is used in

(i) Laboratories.
(ii) Production test stands.
(iii) Repair shops.
(iv) Inspection stations.

The block schematic of this instrument is shown in Fig. 1.91. It consists of typically a 10 MΩ input attenuator. It provides five input ranges from 100 mV to 1000 V, full scale. The DC amplifier with a gain of 100 delivers 10 V to the comparator at any of the full-scale voltage settings of the input divider. When coincidence occurs between the amplified input voltage and the staircase-ramp voltage, display of the input voltage is obtained.

When the measurement cycle is first initiated, the clock that is usually a 4.5 kHz relaxation oscillator provides pulses to three DCUs in cascade. The units counter provides a carry pulse to the tens decade, counts the carry pulses from the units decade, and provides its own carry pulse after it has counted ten carry pulses. This carry pulse is fed to the hundreds decade counter, which provides a carry pulse to an over range circuit. The over-range circuit causes a front panel indicator to light up warning that higher voltage than the maximum range of the instrument in that particular range switch position has been applied. The operator has to change the setting to the next range.

Each decade counter unit is connected to a digital to analog converter (D/A). The outputs of D/As are connected in parallel and provide an output current proportional to the current count of the DCUs. The staircase amplifier converts the D/A current into a staircase voltage, which is applied to the comparator. When coincidence of the input voltage and the staircase voltage occurs, it provides a trigger pulse to stop the oscillator. The current content of the counter is then proportional to the magnitude of the input voltage. The display circuit stores each reading until a new reading is completed, eliminating and blinking or counting during the computation.

The accuracy of measurement depends on the linearity of the ramp voltage. Therefore, a precision resistor and stable capacitors must be used in the integrator. In addition, the offset voltages and currents of the op-amp used in the integrator are critical in the accurate ramp generator.

1.38 DUAL SLOPE INTEGRATING-TYPE DVM

In this method, the accuracy of conversion will not depend on the precision of the resistor and capacitor of the ramp generator current, or the op-amp. In this technique, an integrator is used to integrate an accurate reference voltage for a fixed period of time. The same integrator is then used to integrate with the reverse slope, the input voltage. The time required to return to the starting voltage is measured. The block schematic of this type or DVM is shown in Fig. 1.92.

The input voltage to be measured V_x is applied to the integrator by means of an electronic switch, usually a JFET device. The integration is done for a fixed amount of time, as determined by the counter. As soon as integration starts, the counter is also initiated by the control logic. When the counter reading

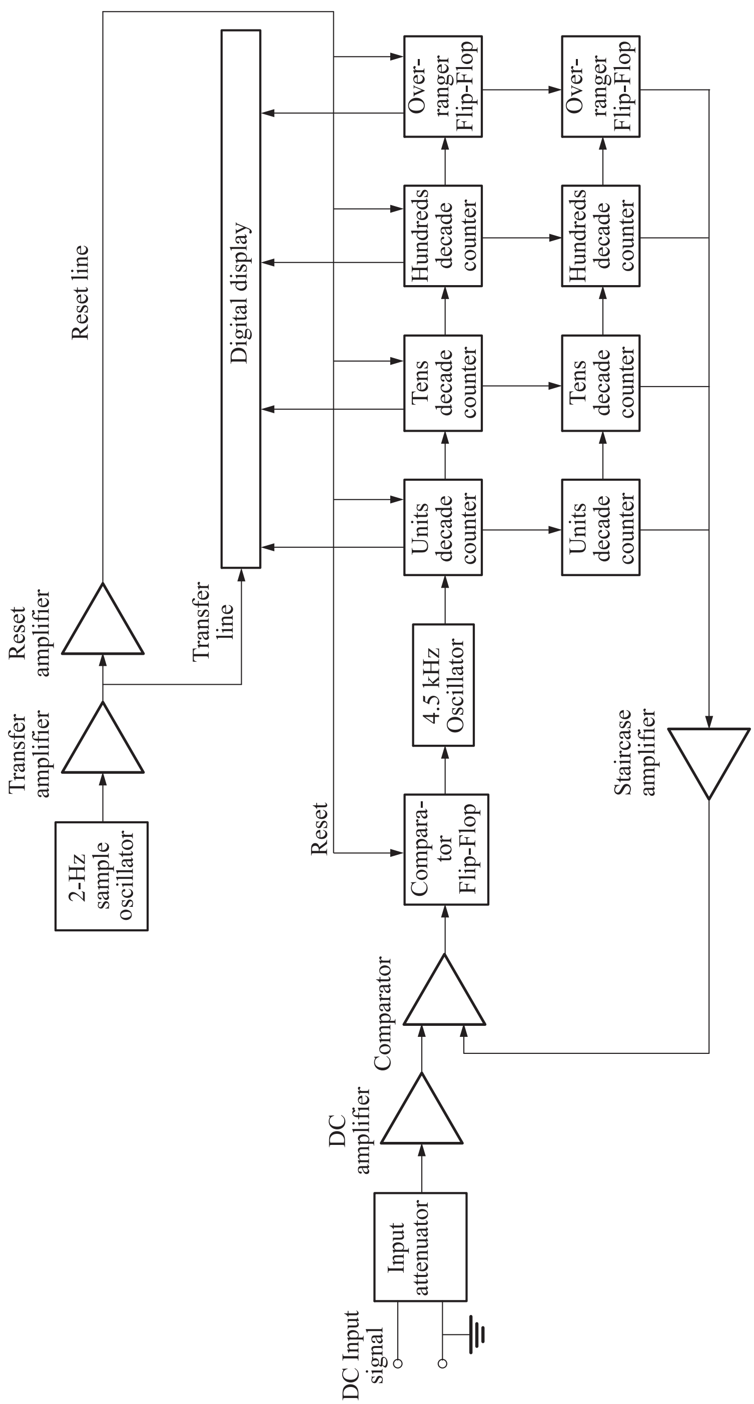


Figure 1.91 Block diagram of a staircase-ramp DVM

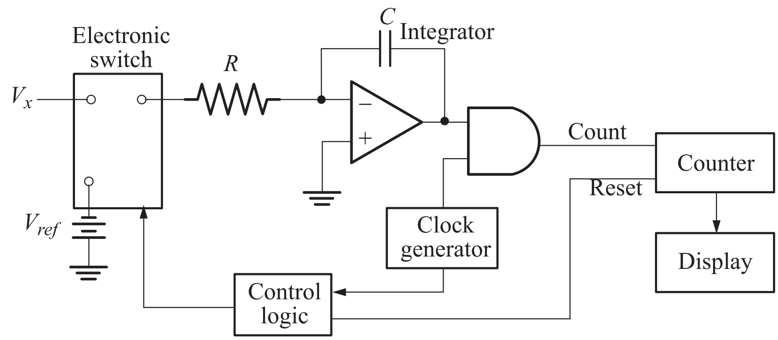


Figure 1.92 Block diagram of a dual slope integrating-type DVM

reaches a predetermined value, integration is stopped. The output of the integrator will have opposite polarity to that of the input voltage V_x. The counter is reset. Reference voltage V_{ref}, which is of opposite polarity to the input voltage, is now connected to the integrator. The integrator output voltage will now go in the opposite direction. At the same time, the counter is also initiated. Integration is done till the output voltage of the integrator becomes 0 V. The counts accumulated during this period are a measure of the input voltage. The counts are accumulated and the digital display or readout is given. The waveforms are as shown in Fig. 1.93.

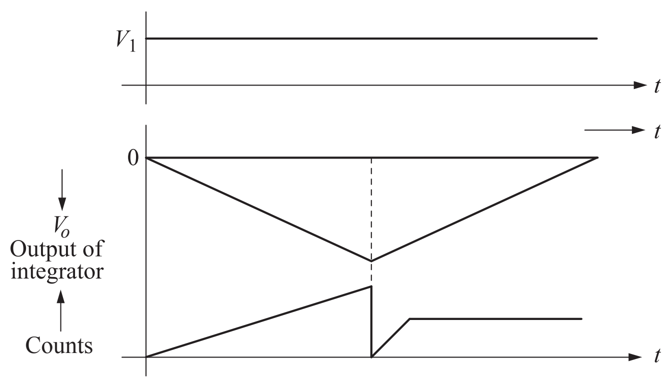


Figure 1.93 Block diagram of a dual slope integrating-type DVM

The integrator circuits used are shown in Fig. 1.94. Output of the integrator

$$|V_o| = \frac{1}{CR} V_i \, T_{integrate}$$

where $T_{\text{integrate}}$ is the period of integration when V_i is connected to the integrator. This time period depends on the counter, to count a prefixed count, at the rate determined by the clock. This $|V_o|$ is also the same when integration is being done for the second time with V_{ref} connected to the integrator.

$$\therefore \; |V_o| = \left(\frac{1}{C_1}\right)\left(\frac{V_{ref}}{R_1}\right) T_{discharge}$$

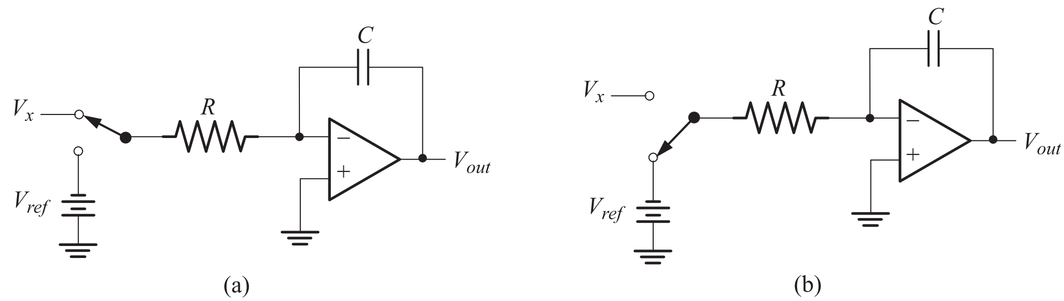


Figure 1.94 Functional schematic of the integrator of an integrating DVM

For a 5-decade counter,

$$T_{integrate} = \frac{10^5}{f_c}$$

$$T_{discharge} = \text{Accumulated counts in the counter}/f_c$$

where

$$f_c = \text{clock frequency}$$

$$\therefore \text{ Accumulated counts} = \left(\frac{V_i}{V_{ref}}\right)10^5$$

$$\frac{V_i}{R_1C_1}T_{integrate} = \left(\frac{V_{ref}}{R_1C_1}\right)T_{discharge}$$

$$T_{integrate} = \frac{10^5}{f_c}$$

$$Vi\left(\frac{10^5}{f_c}\right) = V_{ref}T_{discharge}$$

$$\therefore T_{discharge} = \left(\frac{V_i}{V_{ref}}\right)\left(\frac{10^5}{f_c}\right)$$

$$\text{Accumulated counts} = T_{discharge}f_c$$

$$= \left(\frac{V_i}{V_{ref}}\right)\left(\frac{10^5}{\cancel{f_c}}\right)\cancel{f_c} = \left(\frac{V_i}{V_{ref}}10^5\right)$$

$$\therefore \text{ Accumulated counts} = \frac{V_i}{V_{ref}}10^5$$

Therefore, V_{ref} and the decade counter prefixed count (10^5) are constants, and the counter reading is proportional to the input voltage V_i. Note that the digital readout is *independent of R and C of the integrator and f_c*. Therefore, the precision of *R* and *C* will not affect the measurement, so also the accuracy of the clock frequency f_c. The readout depends on V_{ref}. Hence, a precision reference supply must be used. The clock frequency determines the period of conversion or the conversion time.

The dual slope-type A/D technique is very popular for DVM applications. It is slow but adequate for laboratory instruments. With IC technology advancing, IC chips are available to simplify the construction of DVMs of this type. For data acquisition applications, where a number of measurements are to be made, faster techniques are used. Automatic zero correction can be made with dual slope A/D also. As with any analog system, amplifier offset voltages, onset currents, and bias currents can cause errors. In addition, in the dual slope A/D, the leakage current of the capacitor can cause errors in integration. This results in DVM giving a finite reading, when no input voltage is applied. This can be rectified by modifying the input circuit. The input to the converter is grounded and an autozero capacitor is connected via an electronic switch to the output of the integrator. The feedback of the circuiting is such that the voltage at the integrator output is zero. This effectively places an equivalent offset voltage on the automatic zero capacitor so that there is no integration.

1.39 SUCCESSIVE-APPROXIMATION CONVERSION (SAC)

Principle: The successive-approximation A/D converter (ADC) compares the analog input to a D/A conversion (DAC) module (resistor network) reference voltage, which is repeatedly divided in half. A comparator is used to compare the input voltage and a sequence of voltages, which are binarily related. Initially, the input is compared with $\frac{V_R}{2}$ where V_R is a full-scale input voltage. If the input is greater than $\frac{V_R}{2}$, it is changed. The reference value is now made to $\left(\frac{V_R}{2}+\frac{V_R}{4}\right)$. If the input is less, then it is changed to $\left(\frac{V_R}{2}-\frac{V_R}{4}\right)$. One more comparison is now made and the voltage is now changed depending on the result of the comparison. This procedure is repeated till the desired accuracy is obtained. Suppose the full-scale reference voltage $V_R = 1000 = (8)_{10}$ (for a 4-bit A/D).

Then

$$\frac{V_R}{2} = 100 = (4)_{10} = \left(\frac{8}{2}\right)_{10}$$

The input V_A to be measured is compared with the analog equivalent of 100. If the input is greater than $\frac{V_R}{2}$, then it is compared with $\left(\frac{V_R}{2}+\frac{V_R}{4}=\frac{3V_R}{4}\right)$. If the input is greater than $\frac{3V_R}{4}$, then the next comparison is made with $\frac{3V_R}{4}+\frac{V_R}{8}=\frac{7V_R}{8}= (111)_2$. If the input is greater than $\left(\frac{7V_R}{8}\right)$ comparison is made with V_R itself. If the input is still greater, over voltage indication is given. If the input is less than $\left(\frac{7V_R}{8}\right)$, it is indicated as V_R. If the input is less than $\frac{3V_R}{4}$, the comparison made with $\left(\frac{3V_R}{4}-\frac{V_R}{8}\right)=\frac{5V_R}{4} = (101)_2$. If the input voltage is less than $\frac{5V_R}{8}$, it is compared with $\frac{V_R}{2}$, and the result is displayed by the display units. When the first comparison is made with $\frac{V_R}{2}$ if V_A is less, the second comparison is made with $\left(\frac{V_R}{2}-\frac{V_R}{4}\right)=\frac{V_R}{4} = (0100)_2$.

If the input is still less, a fourth comparison is made with zero and the input is given as zero, since the comparator cannot detect voltages less than $\frac{V_R}{8}$ in this specific case. If in the fourth comparison, V_A is greater than 0000, the next comparison is made with $\frac{V_R}{8} = (0001)_2$. The process is illustrated in Fig. 1.95, for a 4-bit A/D.

After every comparison, for increasing V_R, the next most significant bit (MSB) is incremented. For reducing V_R, the immediate significant bit is decremented.

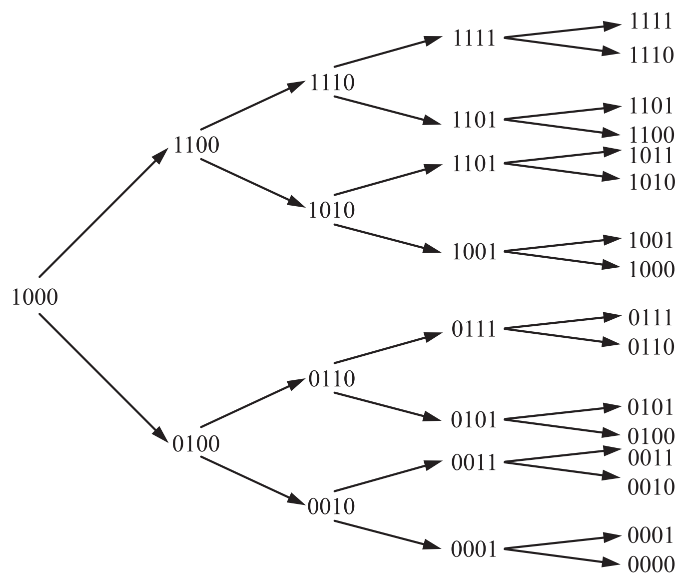


Figure 1.95 Comparison with binary equivalent of V_R with input voltage to be measured

1.39.1 Block Schematic

The schematic of SAC is shown in Fig. 1.96. A D/A converter is used to provide the estimates. The *equal to*, *greater than,* or *less than* decision is made by a comparator. The D/A converter provides the estimate and is compared with the input signal.

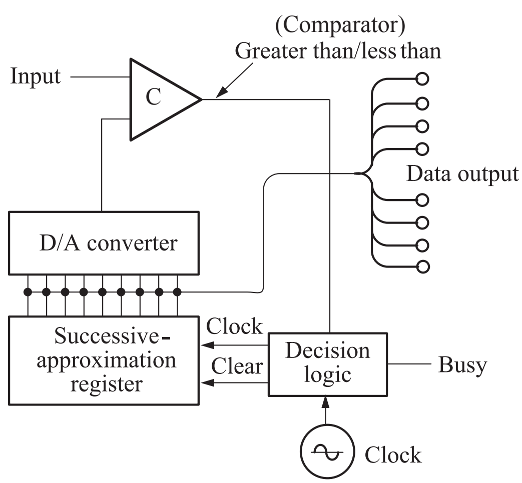


Figure 1.96 Block diagram of a successive-approximation DVM

The special shift register called successive-approximation resistor (SAR) is used to control the D/A converter. At the beginning of conversion, all the outputs from the SAR are at logic zero. If the estimate is greater than the input, the comparator output is high and the first SAR output reverses state and the second output changes to a logic *one*. If the comparator output is low indicating that the estimate is lower than the input signal, the first output remains in the logic one state and the second output assumes the logic *one* state. This continues to all the states until the conversion is complete. An estimate is made on the edge of the SAR clock. For an N-bit conversion, after N clocks the actual value of the input is known. The least significant bit (LSB) is the state of the comparator. In some systems the clock is used to store the last bit in the SAR, and thus (N+1) clocks are required for conversion.

In the ramp-type conversion technique, the input V_i or V_A has to be compared with the ramp voltage. Therefore, for higher-end inputs, the conversion time is more. However, in the successive-approximation technique, comparison is starting from half the maximum input value. Hence, for higher-end input voltages, the conversion time is less. This technique (SAC) is less expensive. Many A/D conversion ICs are available using this technique.

1.40 CONTINUOUS BALANCE-TYPE DVM

This method is also known as the *Servo balancing potentiometer-type DVM*. This is based on servo mechanism and is mostly electromechanical in nature. As such it is not a pure electronic DVM. A mechanical counter is used to give the voltage reading in numerical figures. Hence, it was named DVM in the early days of electronics. The schematic diagram is shown in Fig. 1.97.

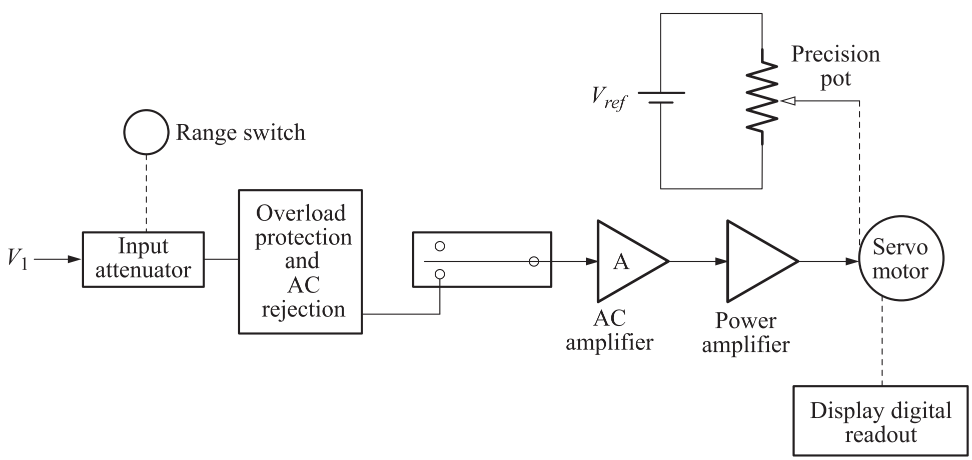


Figure 1.97 Continuous balance-type DVM

The accuracy of the instrument is ±0.1%. The input impedance is 10 MΩ. A DC input is applied to the input attenuator, providing suitable range switching. This is the front panel control, which also causes a decimal point indicator to move on the display area in accordance with the input range selected. After passing through an overload protection circuit and an AC rejection filter, the input voltage is applied to one side of a mechanical chopper comparator. The opposite side of the comparator is connected to the wiper arm of the motor-driven precision potentiometer. The potentiometer is connected across a precision reference supply. The output of the chopper comparator, which is driven by the line voltage and vibrates at the line frequency rate, is a square wave signal.

The amplitude of the square wave is a function of the difference in magnitude and polarity of the DC voltages connected to the opposite sides of the chopper. The square wave signal is amplified by the high impedance, low noise preamplifier, and is fed to a power amplifier. The servomotor drives the arm of the precision potentiometer in the direction required to cancel the difference voltage across the chopper comparator. The servomotor also drives a drum-type mechanical indicator, which has the digits from zero to nine imprinted about the periphery of the drum segments. The position of the servomotor shaft corresponds to the amount of feedback voltage required to null the chopper input and the position is indicated by the drum-type indicator on which digits are marked. This instrument derived its name DVM because of the indicator. The position of the shaft is an indication of magnitude of input voltage.

Modern DVMs have the facility of autoranging. The electronic control circuit's present reading is less than the next lower range or higher than the full scale. If the present reading is less than the full scale of the next lower range, the attenuation is reduced. The attenuation continues to be reduced until the reading is between the next lower range and the full scale of the present range.

1.41 AUTOMATIC POLARITY INDICATION FOR DVM

An automatic polarity indicator that can be used with the DVM is shown Fig. 1.98. The circuit comprises two voltage comparators, both with zero reference and connected to handle signals of opposite polarity. The input terminals of the comparators can be connected to the outputs of the buffer amplifier.

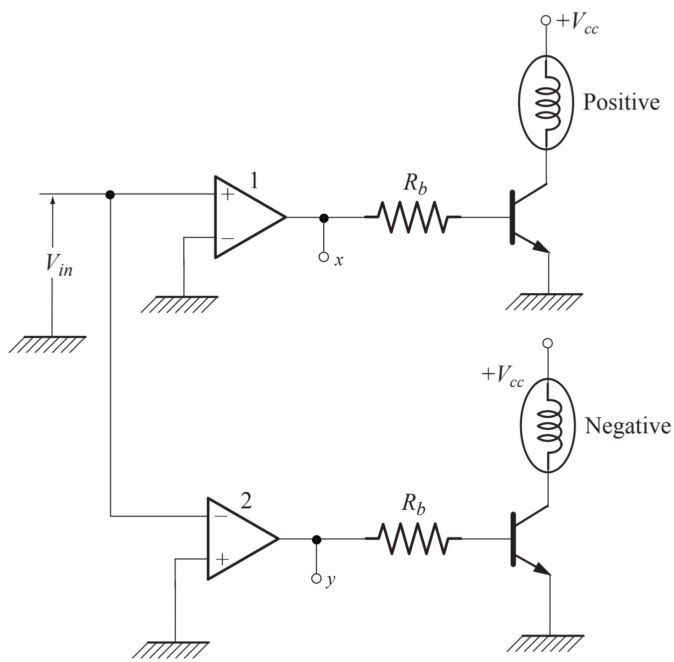


Figure 1.98 Circuit for autopolarity indication in DVMs

The output of comparator 1 will be in the '1' state for positive input signals, whereas the comparator 2 will go to the '1' state for negative inputs. Thus, either a positive or a negative indicator signal results at the output of these voltage comparators depending on the input signal polarity. The high sensitivity of IC voltage comparators —(μA710) or Op-amps used for the purpose (μA741)—ensures excellent performance. The comparator outputs are fed to lamp driver circuits, which facilitate automatic polarity indication. Alternatively, light-emitting diodes (LEDs) or any other single-element display device can be used for this purpose.

1.42 AUTORANGING FOR DVM

The basic DVM has a fixed voltage range of 1 V. The performance of this DVM can be improved by the addition of an input attenuator, which can be controlled to change the range of the instrument. The circuit is shown in Fig. 1.99.

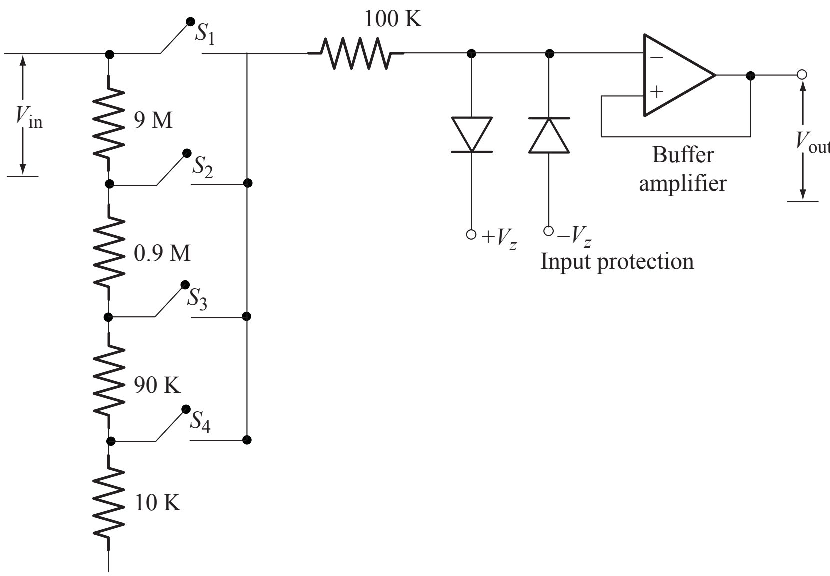


Figure 1.99 Circuit for autoranging in DVMs

A typical attenuator suitable for use at the input terminals of the DVM, that is, proceeding the buffer amplifier is shown.

The attenuator ensures a constant R_{in} for the DVM, example, 10 MΩ in this case, so long as R_{in} of the buffer is much higher than that. In addition, the voltage input to the buffer can be controlled in decade stops by the setting of the switches $S_1 - S_4$, which will be 'on' one at a time. Thus, when S_1 is 'on', the input voltage range is the basic range of the DVM, that is 1 V. The input voltage range is 1000 V when S_4 is 'on'. For ensuring accurate division of V_{in}, it is necessary to have precision resistors, such as metal-film type, low temperature coefficient, and high stability resistors for this application.

The switches $S_1 - S_4$ may be controlled manually or automatically. The latter offers advantages for the user, especially from the point of view of operating convenience.

The autoranging involves the generation of suitable signals for controlling the switch matrix.

1.42.1 Typical Case

Let the DVM be considered as in Fig. 1.98 in its proper range if the input to the basic DVM block is 0.1–1.2 V. This corresponds to a range of *t* of 0.5 m – 6 m or to the counts accumulated in this range, that is, between 100 and 1200. If the counts accumulated in any sample are outside this range, the DVM has to be switched to other ranges.

For example, if the counts are less than 100 at the end of a sample, the instrument should select the next lower range (down-ranging) provided the instrument is not already in this range.

Similarly, if the counts are greater than 1200, the instrument should go to its next higher range (up-ranging) provided it is not already in this range. This range-changing operation has to continue until the instrument locks on to the proper range of measurement (Fig. 1.100).

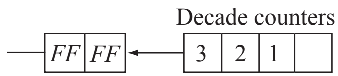


Figure 1.100 Autoranging in DVMs

As four ranges are involved in the present case, a maximum of four sampling intervals are necessary for the instrument to choose the proper range under the worst case. Therefore, the transfer of counts from the counters to auxiliary storage FFs has to be done only after the range switching is complete. Up-range or down-range signals have to be generated using the information of overflow as well as the state of the MSB in the decade counter.

The < 100 counts information is obtained by reading the state of the overflow *FF* (first *T*-flip-flop at the output of the decade counter chain) designated as *Q* and *BCD* output of the counter $\bar{Q}\ \bar{A}\ \bar{B}\ \bar{C}\ \bar{D} = 1$.

Whenever the above equation is satisfied, along with the information that the DVM is in its lower range (which is obtained from the switch driver of S_1), the point *X* changes from 1to > 0. This information resets the down-range *FF* and the up-down counter counts up by one digit.

The >1200 counts information is also obtained similarly. In this case, the MSB counter should read '2', and the overflow or *Q* is in '1' state. This logic is generated by considering all the combinations followed by simplification and the following relationship is the final result:

$$Q(B+C)D = 1$$

Whenever the equation is satisfied along with the information that the DVM is not in its highest range (which is obtained from the switch driver of S_4), the point *Y* changes from '1' to '0'. This information resets the up-range *FF* and up–down counter counts down by one digit.

Thus, the up–down counter counts up or down depending on the presence of a down-range or an up-range signal at its input terminals. Since only four ranges are involved, a 2-bit up–down counter is sufficient for this application. The outputs of the up-down counter have to be combined suitably to yield the four range signals as required (Fig. 1.101).

If *P* is the LSB output and *Q* is the MSB output of the up–down counter, then the following four combinations yield the signals for range switching. At any time, only one of these will be in the '1' state. Hence, only one switch 5 is 'on'.

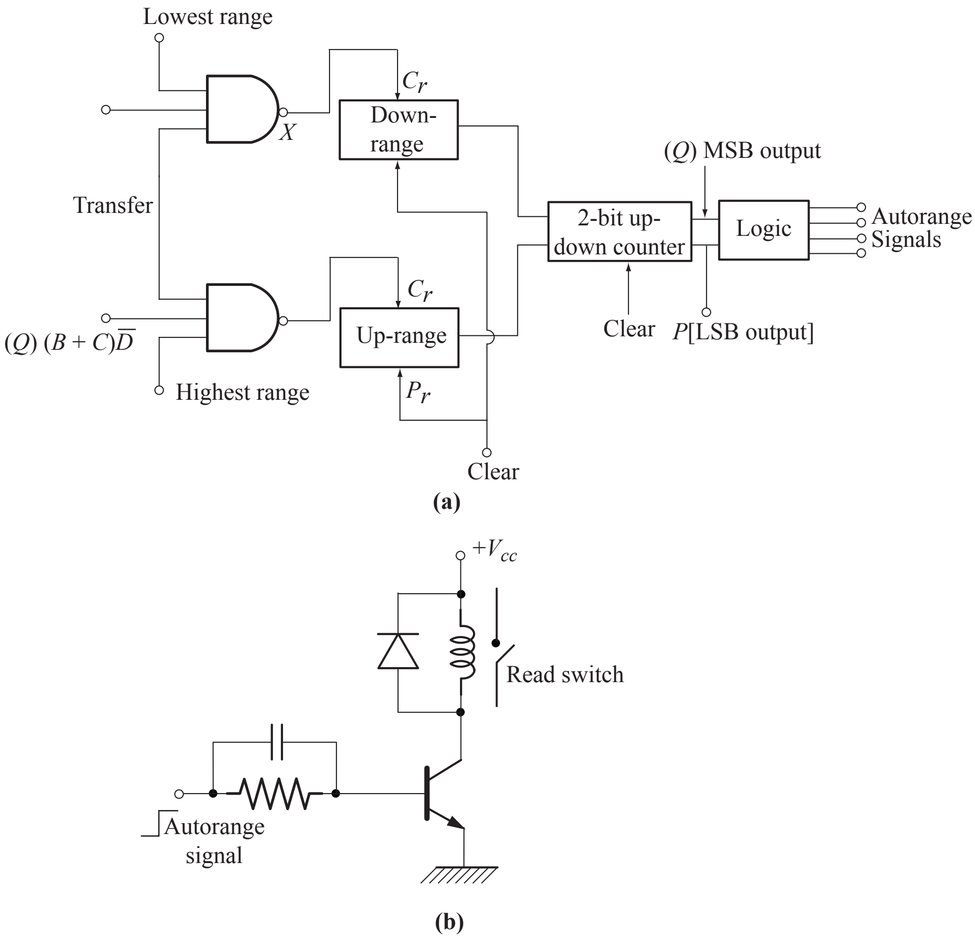


Figure 1.101 Circuit for autoranging in DVMs

Lowest range: PQ
Second range: PQ
Third range: PQ
Highest range: PQ

Since device breakdown voltage rates and R_{off} values for these switches are less in common practice, electromechanical switches such as reed switches are used for this application.

1.43 $3\frac{3}{4}$ DIGIT DISPLAY

Four thousand counts are made in this type of display counter. Accuracy is more compared to a $3\frac{1}{2}$ digit meter. The maximum voltage that can be measured is 1000 V. A $3\frac{3}{4}$ digit DMM can measure voltages typically in the ranges 400 mV, 4 V, 40 V, and 1000 V, with a resolution of 0.1 mV, 1 mV, 10 mV, 100 mV and 1 V, respectively, i.e., with 0.1% resolution, approximately. DC current is measured in the ranges of 4 mA, 40 mA, 400 mA, and 10 A with a resolution of 1, 10, 100, and 10 μA, respectively.

Resistance is measured in the ranges of 400 Ω, 4 kΩ, 40 kΩ, 400 kΩ, 4 MΩ, and 40 MΩ with a resolution of 0.1 Ω, 1 Ω, 10 Ω, 100 Ω, and 1 kΩ, as in the respective ranges. APLAB 2835 DMM is a 4000 counts meter with a $3\frac{3}{4}$ digit display.

1.44 PICOAMMETER

Generally DMMs lack the sensitivity required to measure currents less than 100 mA. Even at higher currents, an input voltage drop of hundreds of mV a cross DMM can make accurate current measurements impossible. Electrometers can measure low currents accurately, but the circuit needed to measure extremely low currents combined with functions like measuring voltage, resistance, and charge will increase the cost of the instrument significantly. Picoammeters are economical and are easy to operate like DMMs.

DMMs employ short ammeter circuitry to measure current. By employing feedback circuitry, the voltage drop can be reduced by several orders of magnitude resulting in a voltage burden of less than 200 μV. This principle is employed in picoammeters. This makes the function of the meter more like an ideal ammeter than a DMM. Therefore, the picoammeters can perform current measurements with high accuracy, even in circuits with a very low source voltage.

Commercially available picoammeters can perform measurements at a speed of about 1000 readings per second. A time-stamped data buffer of 2500 readings provides minimum, maximum, and standard deviation statistics. These instruments can measure current from 20 fA (20×10^{-15}A) to 20 mA.

Some of the features are as follows:

- Transition of measurement results in devices such as DMMs data acquisition boards, oscilloscopes, or strip chart recorders.
- 220 V overload protection.
- Built-in trigger-like interface that simplifies synchronisation of this instrument with other instruments.
- It can be integrated easily into automated test and measurement systems.
- While making measurements in research and on light-sensitive components, such as measuring dark currents of photodiodes, the front panel display can be switched off to avoid introducing light that could reduce the accuracy of the results.
- Using the instrument resistance can also be measured by driving a sourced voltage value by the measured current.

1.44.1 Applications

- Beam monitoring and radiation monitoring.
- Leakage current testing in insulators, switches, relays, and other components.
- Scanning electron microscope (SEM) beam currents.
- Optoelectronic device testing and characterization.
- *I–V* measurements of semiconductor devices and other devices.
- Nanoelectronic device characterization.

Keithly model 485 and 6485 are picoammeters.

1.45 LOW-CURRENT AMMETER APPLICATIONS

1.45.1 Wafer-Level Photodiode Testing

A picoammeter-voltage source can be used with a calibrated light source and photodiodes can be tested at the wafer level in semiconductor device manufacturing. The circuit schematic is as shown in Fig. 1.102.

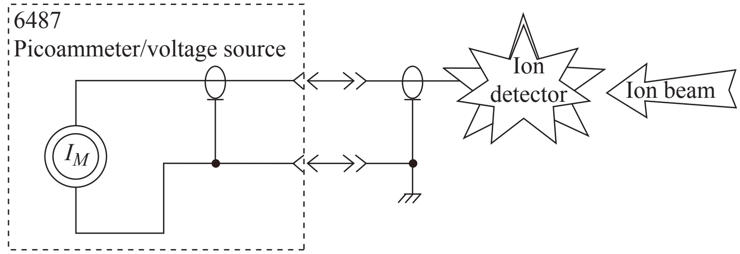


FIGURE 1.102 Diode testing at the wafer level

Using a switch matrix, a picoammeter can be enabled to take readings from multiple pods. The first step of the measurement process is done in total devices. The instrument produces a voltage sweep and then measures the resulting dark current. In the second step, a voltage bias is applied and the resulting photocurrent is measured while the light level is increased in calibrated steps. Similarly, PIN diodes and avalanche photodiodes can also be tested.

1.45.2 Monitoring and Control of Focused Ion Beam Currents

In the manufacture of semiconductor devices, focused ion beam systems are used for manometers, scale imaging, micromapping, and mapping. The magnitude of an ion beam current using an ion detector is important in the system. The ion detector generates a secondary current that is proportional to the current of the primary beam. When the secondary current is measured, it can be used to control the intensity of the primary beam. However, the secondary current is of the order of picoamperes. Therefore, the picoammeter should have good resolutions. Usually a $5\frac{1}{2}$ digit resolution is required. The schematic is shown in Fig. 1.102.

1.46 HIGH-RESISTANCE MEASUREMENTS

Picoammeters can also be used to measure high resistances, greater than 1 GΩ in applications such as insulation resistance testing. A constant voltage source is placed in series with the unknown resistance and the picoammeter. The voltage drop across the picoammeter is negligible. Therefore, all the voltage appears across the unknown resistance. The resulting current is measured by the picoammeter, and the resistance R can be calculated as the ratio of V to I ($R = V/I$). The unknown resistance must be housed in a shielded test fixture to prevent electrostatic interference-generated current affecting the measurement. A small series resistance can be added to reduce noise, if the unknown resistor has high stray capacitance across it. The circuit schematic is shown in Fig. 1.103.

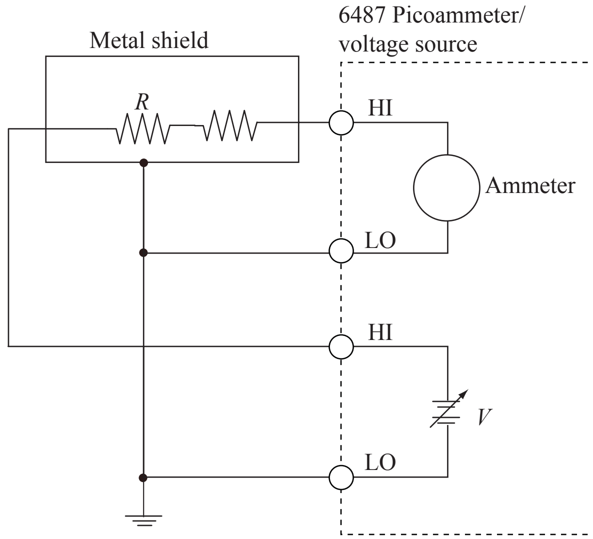


FIGURE 1.103 Circuit for high-resistance measurements

Example 1.23

Determine the period of integration of dual slope integrating-type DVM, which has an integrating capacitor of 0.1 μF and R = 10 kΩ if the reference voltage is 2 V and the output of the integrator is not to exceed 10 V.

Solution

$$\begin{aligned}
\text{Integrator time constant} &= CR \\
&= 0.1\ \mu\text{F} \times 10\ \text{k}\Omega = 10\ \text{msec} \\
\text{Reference voltage } V_R &= 2\ \text{V} \\
\text{Integrator output} &= \frac{2\ \text{V}}{10\ \text{m sec}} = 200\ \text{V/sec} \\
\text{Maximum output of integrator} &= 10\ \text{V} \\
\therefore\ \text{Period of integration} &= \frac{10\ \text{V}}{200\ \text{V/sec}} = 0.05\ \text{sec} \\
&= 50\ \text{ms}
\end{aligned}$$

Example 1.24

The capacitance is specified as 20F ± 5%. Determine the limits of capacitance that it is guaranteed.

Solution

$$\begin{aligned}
\varepsilon_r &= \frac{\text{Percentage error}}{100} \\
&= \frac{5}{100}
\end{aligned}$$

$$
\begin{aligned}
A &= A_m \pm \delta_A \\
&= A_m \pm \delta_r A_m \\
&= A_m(1 \pm \delta r) \\
A &= 20(1 \pm 0.05) \\
&= 20 \pm 1 \text{ F} \\
&= 21, 19\text{F}
\end{aligned}
$$

Example 1.25

A 0–250 milliammeter has an accuracy of 2% FSR. The ammeter measures 150 mA. Determine the limiting error.

Solution

$$
\begin{aligned}
\delta V &= \varepsilon_r V \\
&= 0.02 \times 250 = 5 \text{ mA}
\end{aligned}
$$

$$\%\text{ Limiting error} = \frac{5}{150} \times 100 = 3.33\%$$

Example 1.26

The current passing through a resistor 50 ± 0.2Ω is 4 ± 0.02A. Determine the limiting error.

Solution

$$
\begin{aligned}
\%\text{ Limiting error (resistor)} &= \frac{0.2}{50} \times 100 \\
&= \pm 0.4\% \\
\%\text{ Limiting error (current)} &= \pm \frac{0.02}{4} \times 100 \\
&= \pm 0.5\% \\
P &= I^2R \\
&= 2 \times 0.5 + 0.4 \\
&= 1.4\% \\
P &= I^2R \\
&= (4)^2 \times 50 \\
P &= 800 \text{ W}
\end{aligned}
$$

Example 1.27

Find the resolution of a moving-coil ammeter that has a scale of 100 divisions and gives FSRs of 10 A. The instrument can read up to $\frac{1}{2}$ of the full-scale division.

Solution

$$\text{Full-scale reading} = 10\text{ A}$$

$$\text{No. of div.} = 100$$

$$\text{1-scale div.} = \frac{10}{100} \times 1000 = 100\text{ mA}$$

$$\text{Resolution} = \frac{1}{2}\text{ of a division}$$

$$= \frac{100}{2} = 50\text{ mA}$$

Example 1.28

Two capacitors are in series and also in parallel. Determine the limiting error:

$$C_1 = 99 \pm 1\text{ Mf}$$

$$C_2 = 49 \pm 1\text{ Mf}$$

Solution

In parallel	In series
$y = C_1 + C_2$	$\frac{1}{y} = \frac{1}{C_1} + \frac{1}{C_2}$
148 ± 2 Mf	$\frac{1}{(99 \pm 1\text{Mf})} + \frac{1}{(49 \pm 1\text{Mf})}$

Limitation error

In parallel = ±2 Mf.

In series = 0.25 Mf.

Example 1.29

Three resistors have the following values:

$$R_1 = 200\ \Omega \pm 5\%$$

$$R_2 = 100\ \Omega \pm 5\%$$

$$R_3 = 50 \pm 5\%$$

Solution

Resistor R_{se} connected in series and parallel

(a) Series

$$R_{series} = R_1 + R_2 + R_3$$

$$= 3\text{ n}\Omega$$

$$\text{Relative error} = \frac{R_1}{R_{sc}}\frac{SR_1}{R_1} + \frac{R_2}{R_{sc}}\frac{SR_2}{R_2} + \frac{R_3}{R_{sc}}\frac{SR_3}{R_3}$$

$$= \frac{200}{300} \times 5 + \frac{100}{350} \times 5 + \frac{50}{350} \times 5$$

$$= \pm 5\%$$

$$\text{Relative error} = 350 \times \frac{5}{100} = 17.5\ \Omega$$

(b) Parallel

$$\frac{1}{R_p} = \frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}$$

$$= \frac{R_2R_3 + R_1R_3 + R_1R_2}{R_1R_2R_3}$$

$$= \frac{R_1R_2R_3}{R_2R_3 + R_1R_2 + R_1R_3}$$

$$R_P = 28.5\ \Omega$$

Let RP = $\frac{x}{y}$

$$\text{Error in } x = \frac{\delta R_1}{R_1} + \frac{\delta R_2}{R_2} + \frac{\delta R_3}{R_3}$$

$$= 5 + 5 + 5 = \pm 15\ \%$$

$$\text{Error in } y_1 = \frac{\delta R_1}{R_1} + \frac{\delta R_2}{R_2}$$

$$= 5 + 5 = 10\ \%$$

$$\text{Error in } y_2 = \frac{\delta R_2}{R_2} + \frac{\delta R_3}{R_3}$$

$$= 5 + 5 = 10\%$$

$$\text{Error in } y_3 = \frac{\delta R_3}{R_3} + \frac{\delta R_1}{R_1}$$

$$= 5 + 5 = \pm 10\%$$

$$\%\ \text{Error} = \frac{y_1}{y}\frac{\delta y_1}{y_1} + \frac{y_2}{y}\frac{\delta y_2}{y_2} + \frac{y_3}{y}\frac{\delta y_3}{y_3}$$

$$= \frac{20{,}000}{35{,}000} \times 10 + \frac{5000}{35{,}000} \times 10 + \frac{10{,}000}{35{,}000} \times 10$$

$$= \pm 10\ \%$$

$$\%\ \text{Error} = 15 + 10 = \pm 25\%$$

$$\text{Error} = 28.57 \times \frac{25}{100} = 7.142\ \Omega$$

Example 1.30

The following readings are obtained in the measurement of an inductance: 1.003, 0.998, 1.001, 0.991, 1.009, 0.996, 1.005, 0.997, 1.008, 0.994 MH. Calculate

(a) The arithmetic mean.
(b) The average deviation.
(c) The standard deviation.

Solution

(a)

$$\bar{X} = \frac{1.003 + 0.994}{10} = 0.999 \text{ MH}$$

(b)

$$\begin{aligned}
d_1 &= 1.003 - 0.999 = +0.0038 \\
d_2 &= -0.0012 \\
d_3 &= +0.0018 \\
d_4 &= 0.0098 \\
d_5 &= -0.0082 \\
d_6 &= +0.0058 \\
d_7 &= -0.0032 \\
d_8 &= -0.0022 \\
d_9 &= -0.0088 \\
d_{10} &= -0.0052
\end{aligned}$$

(c)

$$\begin{aligned}
\text{Average deviation} &= d_1 + \cdots + d_{10}/10 \\
&= 0.005 \text{ MH} \\
\text{Standard deviation, } \sigma &= \sqrt{\frac{(d_1)^2 + \ldots (d_{10})^2}{10 - 1}} \\
&= 0.006 \text{ MH}
\end{aligned}$$

Example 1.31

The following 10 observations were recorded: 41.7, 42, 41.8, 42, 42.1, 41.9, 42, 41.9, 42.5, 41.8 for an ammeter. Find

(a) The mean.
(b) The standard deviation.
(c) The probable error of one reading.
(d) The probable error of mean.
(e) The range.

(a) Mean:

$$\bar{X} = 41.7 + \ldots 41.8/10 = 41.97 \text{ A}$$

(b) Standard deviation:

$$d_1 = -0.27,\ d_2 = +0.03,\ d_3 = -0.17,\ d_4 = +0.03,\ d_5 = +0.13$$
$$d_6 = -0.07,\ d_7 = +0.03,\ d_8 = -0.07,\ d_9 = +0.53,\ d_{10} = -0.17$$
$$= \sqrt{d_1^2 \ldots + d_2^2 / 10 - 1} = 0.22 \text{ A}$$

(c) Probable error of one reading:

$$\gamma = 0.6745\sigma = 0.148 \text{ A}$$

(d) Probable error of mean:

$$Vm = \frac{\gamma}{\sqrt{n-1}} = \frac{0.148}{\sqrt{9}} = 0.049 \text{ A}$$

(e) Range:

$$= 42.5 - 41.7 = 0.8 \text{ A}$$

Example 1.32

If Young's modulus for phosphor is 1.2×10^4 kg per mm^2, estimate the approximate torque produced by the string.

Solution

Length of the strip,

$$l = 400 \text{ mm}$$
$$\text{Width,} \quad w = 0.5 \text{ mm}$$
$$\text{Thickness,} \quad t = 0.08 \text{ mm}$$
$$E = 1.2 \times 104 \text{ kg/mm}^2$$
$$\theta = 90°$$
$$T = \frac{Ewt^2}{12l}$$
$$= 100.5 \times 10^{-8} \text{ kg-m}$$

Example 1.33

In gravity-controlled instruments, the controlling weight is 0.005 kg and acts at a distance of 2.4 cm. Determine the deflection in degrees corresponding to the deflection torque of 10.5×10^{-4} kg-m.

Solution

$$W = 0.005 \text{ kg}$$
$$\text{Distance of controlling weight} = 2.4 \text{ cm}$$
$$T_d = 1.05 \times 10^{-4} \text{ kg-m}$$
$$T_d = Wl \sin\theta$$
$$T_d = 0.005 \times 0.024 \times \sin\theta$$
$$\theta = \sin^{-1}\left(\frac{1.05 \times 10^{-4}}{0.005 \times 0.024}\right) = 61°$$

Example 1.34

The torque of an ammeter varies as a square of the current. If a current of 10 A produces a deflection of 90°, what deflection occurs for a current of 1 A when the instrument is (a) spring controlled and (b) gravity controlled?

(a)

$$Tc \propto \theta$$
$$\theta \propto I^2$$
$$\frac{\theta_2}{\theta_1} = \left(\frac{I_2}{I_1}\right)^2$$
$$= 90° \times \left(\frac{5}{10}\right)^2 = 22.5°$$

(b)

$$Tc \propto \sin\theta$$
$$\theta \propto I^2$$
$$\frac{\sin\theta_2}{\sin\theta_1} = \left(\frac{I_2}{I_1}\right)^2 = 14.5°$$

Example 1.35

A circuit consisting of an unknown coil, resistance, and variable capacitance counted in series is found to resonate using a Q-meter. If the frequency is 450 kHz and the resonating capacitor is 250 pf, Determine the effective resistance, R.

Solution

$$Q = 105, \quad R_{sh} = 0.75\ \Omega$$
$$f = 450 \text{ kHz}$$
$$C = 250 \text{ pf}$$
$$L = \frac{1}{(2\pi f)^2 c}$$
$$= 500 \text{ MH}$$
$$R = \frac{WL}{Q} - \text{R}_{\text{sh}} = 12.76\ \Omega$$

Example 1.36

A coil is tuned to a resonance frequency of 1 MHz with a resonating capacitance of 480 pf. At 2 MHz, the resonance frequency with a resonating of the coil of 120 pf. Determine C_d and Q of the coil. $R = 10\ \Omega$.

Solution

Given

$$R = 10\ \Omega,$$
$$f_1 = \text{MHz},$$

$$C_1 = 480 \text{ pf},$$
$$f_2 = 2 \text{ m},$$
$$C_2 = 120 \text{ pf}.$$

(1) $C_d = \dfrac{C_1 4 l_2}{3} = 40 \text{ pf}$

(2) $Q = 1/2\pi f(C_1 + d)R$
$= 30.6$

Specification of a DMM

Type: $3\frac{3}{4}$ Digit 4000 counts LCD Bench Multimeter.

Model No. xxxx.

Sl. No.	Description	Typical Values
1.	DC volts	
	Ranges	400 mV, 4 V, 40 V, 400 V, 1000 V
	Resolution	0.1 mV, 1 mV, 10 mV, 100 mV, 1 V
	Accuracy	
	Input impedance	10 MΩ
2.	AC volts	For ranges, resolution and accuracy, same as AC volts range
	Input impedance	10 MΩ < 100 pF
3.	DC current	
	Ranges	4 mA, 40 mA, 400 mA, 10 A
	Resolution	1 µA, 10 µA, 100 µA, 10 µA
	Accuracy	0.75% + 2 digits
	Overload protection	1 A, 600 V
4.	AC current	
	Ranges	4 mA, 40 mA, 400 mA, 10 A
	Resolution	1 µA, 10 µA, 100 µA, 10 µA
	Basic accuracy	1.5% + 5 digits
	Overload protection	600 V DC or AC peak
5.	Resistance	
	Ranges	400 Ω, 4 kΩ, 40 kΩ, 400 kΩ, 4 MΩ, 40 MΩ
	Resolution	0.1 Ω, 1 Ω, 10 Ω, 100 Ω, 1 kΩ, 1
	Basic accuracy	0.75% rdg + 4 digits

1.47 SUMMARY

The basic deflecting systems of D'Arsonval and an electrodynamometer type of deflecting systems are described. The construction of the meter movement and the principle of working have been explained. Materials used and typical parameter values are also given. The DC and AC instruments based on this principle are explained. Extending the range of a given ammeter and voltmeter is also explained. Frequency compensatory elements and determination of their values have also been given. Voltage, current and resistance measurements, and working of a multimeter are explained. Numerical problems on the determination of value of resistance to extend the range and for a selector switch are also given.

Points to Remember

- Transducer is a device that converts a physical quantity into electrical quantity or vice versa.
- Quantity to be measured is called 'measurand'.
- Amplification, filtering, modulation, demodulation, A/D, and D/A conversion are referred to as signal conditioning.
- The terms accuracy, precision, sensitivity, resolution, repeatability, responsibility, drift, error, and limiting error describe the performance characteristics of instruments.
- Fidelity and speed of response pertain to dynamic characteristics.
- The errors that occur in instruments are classified as gross errors, systematic errors, and random errors.
- When random errors are predominant, the statistical analysis method is employed to determine the quantum of error in measurement.
- Probable error = $\pm 0.6745\sigma$, where σ is the standard deviation.
- Expression for torque T in a permanent magnet moving- coil (PMMC) system is $T = BAIN$.
- In electrodynamometer movement, the deflecting torque T_d is proportional to I.
- Electrical shunts are used to extend the range of ammeters.
- By connecting a resistor of suitable value in series, the range of voltmeter can be extended.

Objective-type Questions

1. The quantity to be measured is called ________.
2. An example of a signal conditioner is ________.
3. High-precision measurement can be done by ________ type of instruments.
4. The smallest change in the measured value to which the instrument can respond is ________.
5. The different types of drifts are ________.
6. The quantity of indication by the instrument with regard to the change in input is ______.
7. The delay in response is known as ________.
8. The relationship between standard deviation σ and probable error is ________ .
9. The equation for the Gaussian law for probability of occurrence of deviation ω is ________ .
10. The expression for torque produced in a permanent magnet moving-coil movement T is ________.
11. Shunts are used to extend the range of ________ meters.
12. For a dual slope integrating-type DVM with five decade counters and clock frequency fc the expression for T, the time to integrate is ________ .
13. Two meters A and B require 30 mA and 40 mA, respectively, to give a full-scale deflection. The one that is more sensitive is ________ .

14. The last interval between two adjacent readings that can be distinguished from the other is called __________ .
15. The change in the first movement reading when the input is first increased and then decreased is called __________ .
16. The maximum variation in the ammeter zero due to temperature variations is called ______ .
17. In a moving iron or moving-coil type of instrument, generally the meter range is related so that the readings are obtained near the __________ .
18. The number 1.5×10^5 has __________ significant figures.
19. In portable instruments, the controlling torque is provided by __________ .
20. A DC ampere-hour meter registers__________ if connected to a voltage that is lower than rated.

Review Questions

1. Describe the construction details and working principle of D'Arsonval type of instrument. What are its advantages and disadvantages?
2. Explain the principle and working of an electrodynamometer type of instrument. Describe the expression for the deflecting torque.
3. Compare D'Arsonval and electrodynamometer movements.
4. How do you extend the range of a given ammeter and voltmeter? Describe the expressions used.
5. Using necessary circuits, explain how *V*, *I*, and *R* measurements are made with a VOM meter.
6. What are shunts? What are the materials used for their construction? Give the construction details of shunts.
7. Draw the circuit for an electronic DC meter and explain its working.
8. Explain the principle of a chopper-stabilised DC amplifier used in meters. What are its advantages and disadvantages?
9. Draw the circuits and explain how average, peak, and peak-to-peak values of a given signal are determined.
10. With the help of a schematic diagram, explain how the true *rms* value of a given input is determined by the meter. What are its salient features?
11. Draw the circuit of an electronic multimeter and explain how DC and AC currents and voltages and resistance are measured.
12. With the help of a block schematic, explain the principles and working of a ramp-type DVM.
13. What are the salient features of a dual slope integrating-type DVM? Describe its working.
14. What is the principle of successive-approximation-type DVM? Give the block schematic and explain its working.
15. Compare the performance and applications of different types of DVMs.

Unsolved Problems

1. The capacitance of a capacitor is specified as 30 F ± 6%. Determine the limit of capacitance to which it is guaranteed.
2. One universal shunt has a total resistance of 5 kΩ. A galvanometer has a resistance of 1 kΩ. Determine the multiplying power of the shunt for 2, 3, 4, and 5 kΩ tappings.
3. An AC voltammeter has a maximum scale reading of 100 V. It has an inductance of 0.1 H and a total resistance of 1000 Ω. By connecting a capacitor across the non-conductive series resistance of the voltammeter, the meter can be made to read correctly on both DC and AC. Determine the value of the capacitor. The meter resistance $R = 50\ \Omega$.
4. The full-scale deflection torque of 20 A, moving iron ammeter is 5×10^{-5} N-m. Determine the rate of change of self-inductance of the meter in μH/radians.

5. A moving-coil voltmeter spring control produces a deflection of 1° for a torque of 25×10^{-7} N-m. The resistance of the moving-coil voltammeter is 10 kΩ. The coil has 100 turns, 5 cm long and, 2 cm wide. The flux density of the air gap is 5×10^{-2} wb/m^2. Determine the deflection produced for a 200 V scale reading.
6. A 0–250 mA range ammeter has an accuracy of 2% of full-scale reading. The ammeter measures 150 mA. Determine the limiting error.
7. The current passing through a resistor of 50 ± 0.2 Ω is 4 ± 0.02 A. Determine the limiting errors.
8. Find the resolution of a moving-coil ammeter that has a scale of 100 divisions. Its full-scale reading is 10 A. The instrument can read upto $\frac{1}{2}$ the division.
9. Two capacitors of values C_1 = 99 ± 1 μF and C_2 = 49 ± 1 μF are connected in series once and in parallel next. Determine the limiting error in both the cases.
10. Three resistors R_1, R_2, and R_3 have the following values. R_1 = 200Ω ± 5%, R_2 = 100Ω ± 5%, R_3 = 50 Ω ± 5%. Determine the relative error, if the resistors are:

    (a) Connected in series,
    (b) Connected in parallel.
11. The following readings are obtained in the measurement of an inductor: 1.003, 0.998, 1.001, 0.991, 1.009, 0.986, 1.005, 0.997, 1.008, and 0.994 μH. Determine the
    (a) Arithmetic mean.
    (b) Average deviation.
    (c) Standard deviation.
12. The following 10 observations were received in the case of an ammeter: 41.7, 41.8, 42, 42, 42.1, 41.9, 42, 41.9, 42.5, 41.8 mA. Determine
    (a) The mean.
    (b) The standard deviation.
    (c) The probable error of one reading.
    (d) The probable error of mean.
    (e) The range.
13. If Young's modulus for Bronze is 1.2×10^4 kg/mm^2, estimate the torque produced by the string. Given L = 400 mm, width = 0.5 mm, thickness t = 0.08 mm.
14. In a gravity-controlled instrument, the controlling weight is 0.005 kg, and acts at a distance of 2.4 cm. Determine the deflection in degrees corresponding to the deflection torque of 1.05×10^{-4} kg-m.
15. In the case of a spring-controlled ammeter, a current of 10 A produces a deflection of 90°. If the current is 5 A, what deflection occurs in degrees?

2

Waveform Generators

Introduction • Considerations in choosing an oscillator or signal generator • Sine wave generator • Oscillator circuit • Attenuator • Frequency-synthesised signal generator • Sweep-frequency generator • Pulse and square wave generator • Function generator • Arbitrary waveform generator • Video signal generator • Summary

2.1 INTRODUCTION

Generation of signals is an important aspect in electronic circuits and troubleshooting. Signal generators are widely used in testing electronic circuits, systems, and equipments. A signal generator must be capable of producing stable signals over a wide range of frequencies from a few Hz to even in the GHz range. The amplitude must be variable, and attenuators are usually provided to change the amplitude. The amplitude must also be variable from a low value to a high value. The signal generated by the instrument must be free from distortion. Amplitude and frequency stability with variation in temperature must be good. These are the desirable features of signal generators.

There are various types of signal generators:

1. *Standard signal generator:* This instrument produces sinusoidal waveforms in audio frequency (AF) and radio frequency (RF) ranges. Continuous wave and modulated RF signals can also be produced. This is an oscillator with a modulation capability.
2. *Oscillators:* These are available in the AF and RF ranges separately with variable amplitude and frequency. Modulated signals will not be available.
3. *Test oscillator:* It is also an oscillator circuit with a calibrated attenuator and an output monitor.
4. *Function generator:* If the instrument is capable of generating square, triangular, ramp, pulse, and sine waves or some of these different types of waveforms in addition to the sine wave, then the instrument is called *function generator*. In this instrument, the sine wave is synthesised. Distortion will be more in the sine wave compared to that produced in oscillator instruments.
5. *Pulse generator:* Rectangular waveforms with variable duty cycles, variable frequency, and amplitude are produced in this instrument.
6. *Sweep generator:* Ramp waveforms with variable slopes are produced by this instrument.

Function generators, pulse, and sweep generators are specialised signal generators. The term oscillator is used if the instrument produces only a sine wave with least distortion employing an electronic oscillator circuit. The term function generator is used if the instrument produces square and triangular waveforms in addition to the sine wave. The sine wave is usually derived by synthesising the triangular wave.

Signal sources can be broadly classified as:

1. Fixed.
2. Variable.

In fixed signal generators, the amplitude of the waveform or the frequency or both may be fixed. But such instruments have limited applications.

In the variable type, the amplitude of the waveform can be varied from '*mv*' to 'volts'. The frequency is also variable over a wide range. Most of the commercial instruments are of this type.

2.2 CONSIDERATIONS IN CHOOSING AN OSCILLATOR OR SIGNAL GENERATOR

The following factors must be considered in selecting or comparing signal generators.

1. *Frequency range:* The instrument must be capable of giving a wide range of frequency signals, from low frequency signals to very high frequency signals typically from 1 Hz to 30 MHz or even higher.
2. *Output voltage:* Variable from low to a sufficiently high value.
3. *Resolution:* Small variation in frequency and amplitude could be made.
4. *Accuracy:* The deviation from the set values must be minimum.
5. *Frequency stability:* Ability to maintain the selected frequency over a period of time must be good. Component ageing, and power supply fluctuations will affect this parameter.
6. *Amplitude stability:* Amplitude must remain constant at the set value when frequency is changed.
7. *Distortion:* It should be minimum. The waveform generated will be usually distorted at very low frequency and very high frequency ranges.

2.3 SINE WAVE GENERATOR

The sine wave is very important and is widely used in electronic circuits. Therefore, most of the signal generators are sine wave generators. These are commercially known as *Oscillators*. This instrument covers a frequency range of few Hzs to few MHzs and in some cases to the GHzs range. In the simplest form, the instrument can be as shown in Fig. 2.1.

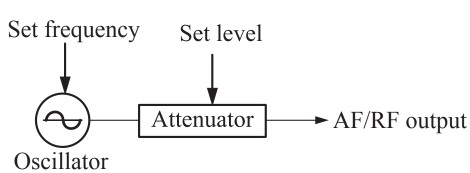


Figure 2.1 Block diagram of a sine wave generator

Depending upon the frequency range, specifications, and cost, the instrument may employ the RC phase shift oscillator circuit or Wien bridge oscillator or Hartley oscillator or Colpitts oscillator circuits to generate the sine wave. The last two are the commonly preferred circuits. An attenuator is used to vary the amplitude levels in the low and high ranges.

2.4 OSCILLATOR CIRCUIT

The block diagram of the oscillator circuit is shown in Fig. 2.2.

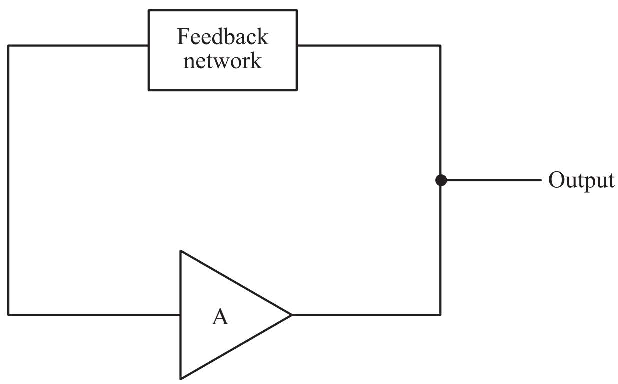


Figure 2.2 Block diagram of an oscillator circuit

The resonant characteristic of an LC-tuned circuit is made use of, to provide a low-distortion sine wave. The frequency of oscillation is given by the expression:

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

where

L = circuit inductance (Henry)

C = circuit capacitance (Farad)

f = resonant frequency (Hertz)

Various considerations of L, C, and R are used to make the frequency variable. The oscillator circuit consists of an amplifier and a feedback network. When the Barkhausen criteria are satisfied, oscillations are produced. Figure 2.3 shows the actual circuit of a Hartley oscillator and the equivalent circuit showing the amplifier and feedback components.

Figure 2.4 shows a Colpitts oscillator circuit.

These two circuits can be used up to a high-frequency range of even 1 GHz. For microwave frequencies, specialised oscillators are used. For lower frequencies, the size of the inductors required for the tuned circuits becomes prohibitive.

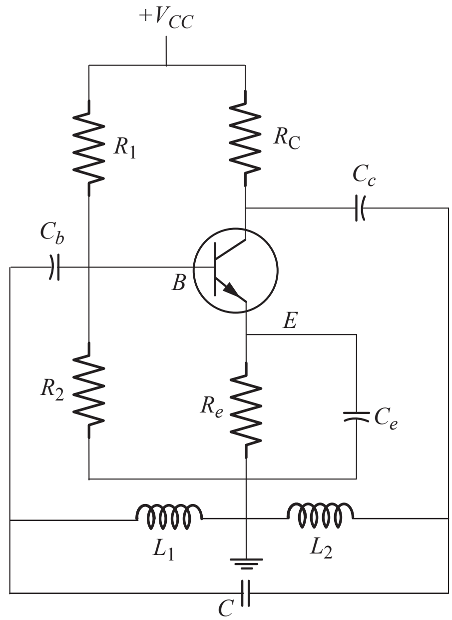


Figure 2.3 Hartley oscillator circuit

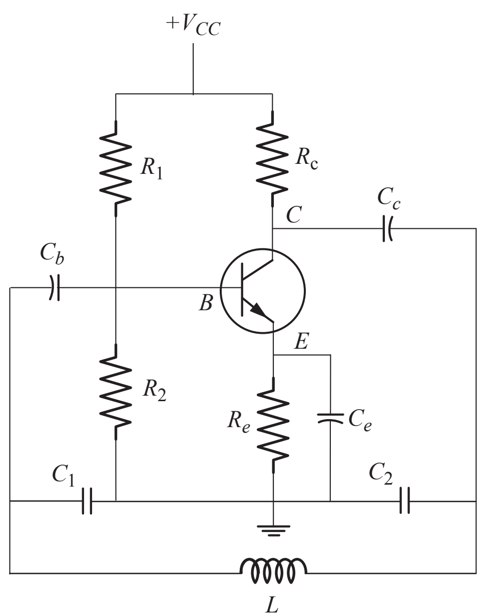


Figure 2.4 Colpitts oscillator

2.5 ATTENUATOR

The attenuator reduces the power of an input so that the ratio of the input power to the output power is a constant. The attenuator is the second part of the sine generator.

$$A(\text{dB}) = 10 \log \frac{P_i}{P_o}$$

where

A = attenuation in decibels
P_i = power input
P_o = output power

If a signal is passed through two attenuators in cascade (see Fig. 2.5), the total reduction is the product of the two attenuations.

$$\begin{aligned} A(\text{dB}) &= 10 \log \left(\frac{P_i}{P_o}\right)\left(\frac{P_i}{P_o}\right) \\ &= 10 \log \left(\frac{P_i}{P_o}\right) + 10 \log \left(\frac{P_i}{P_o}\right) \end{aligned}$$

or $A(\text{dB}) = A_1 + A_2$ in decibels.

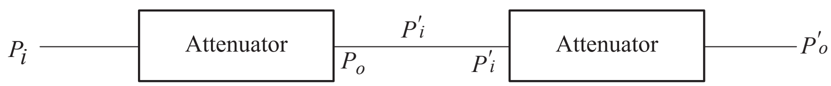


Figure 2.5 Two attenuators cascaded

The *Pi* (π) attenuator is shown in Fig. 2.6. Attenuation of 20 dB can be produced and this circuit can be used up to 100 MHz.

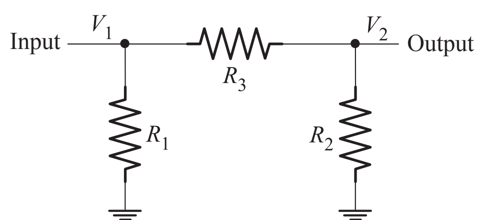


Figure 2.6 π (*Pi*) attenuator

If the parasitic reactances are also taken into consideration, then the *Pi* attenuator appears as in Fig. 2.7.

When the attenuator connected to the oscillator circuit is set to values of 20 dB or more, the isolation requirements are met. However, if a signal of higher amplitude is required, attenuation is to be reduced. Therefore, the isolation of the signal generator from the other circuit load is also reduced. This problem can be solved by providing an isolation amplifier between the oscillator and the attenuator.

An amplifier having a gain of 10 dB can be made to have an isolation of 20 dB or greater. Signal generators require a precise frequency readout. Precision crystal calibrators are included in more expensive instruments to periodically check the dial calibration. A built-in frequency counter in the signal generator instrument makes frequency measurement accurate and simple. Figure 2.8 shows the block diagram of a modern signal generator with a frequency counter display, an isolation amplifier, and an automatic level control system.

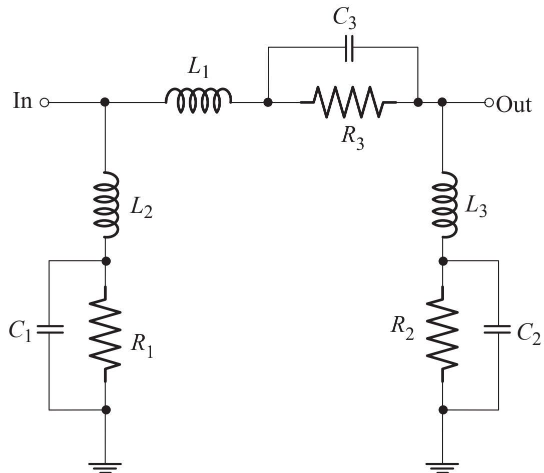


Figure 2.7 π (*Pi*) attenuator considering parasitic reactances

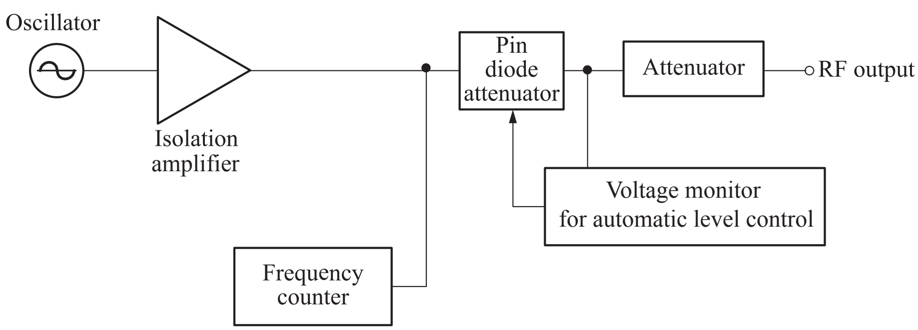


Figure 2.8 Modern sine wave signal generator

2.6 FREQUENCY-SYNTHESISED SIGNAL GENERATOR

The block schematic of this instrument is shown in Fig. 2.9. This method of frequency synthesis is called the *indirect method*. Here the phase locked loop (PLL) technique is employed.

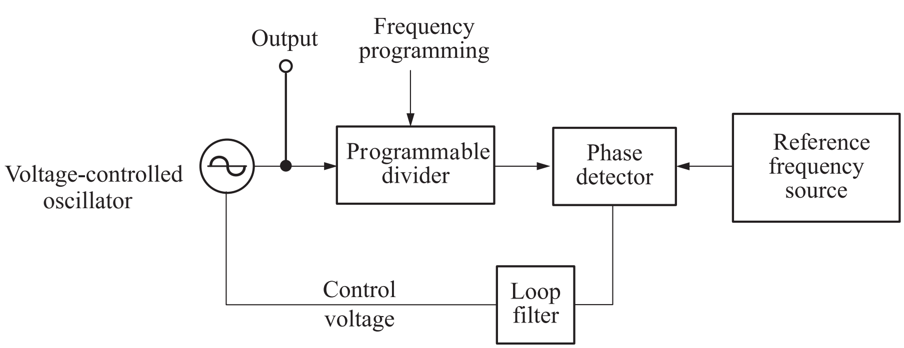


Figure 2.9 PLL-based signal generator

The different blocks are as follows:

Voltage-controlled oscillator (VCO): It is the source of output frequency. It can be tuned electronically.

Programmable divider: It is a logic element that divides the frequency of the VCO by an integer that can be entered through a microprocessor or programming switches.

Phase detector: It provides an analog output that is a function of the phase angle between the two inputs.

Reference source: It is a very accurate and stable frequency source such as a quartz crystal oscillator. The crystal operator works in the frequency range of 1–10 MHz.

Loop filter: It is an analog filter and ensures stable and noise-free operations.

The output of the programmable divider is fed to the phase detector and is compared to the phase of the reference frequency. The output of the phase detector is returned to the VCO and any variation in phase could be corrected so that the frequency of VCO would be equal to the reference frequency.

2.7 SWEEP-FREQUENCY GENERATOR

Solid-state variable capacitance diodes contribute towards the development of sweep-frequency generators. The block diagram of a simple sweep generator is shown in Fig. 2.10. A sweep generator oscillator can be electronically tuned. A sweep-voltage generator is supplied within a generator to provide the frequency sweep. The relationship between the sweep voltage and the frequency of the oscillator is not linear. Therefore, a compensating circuit is provided between the sweep-frequency voltage and the oscillator tuning voltage. A typical linearising circuit is shown in Fig. 2.11.

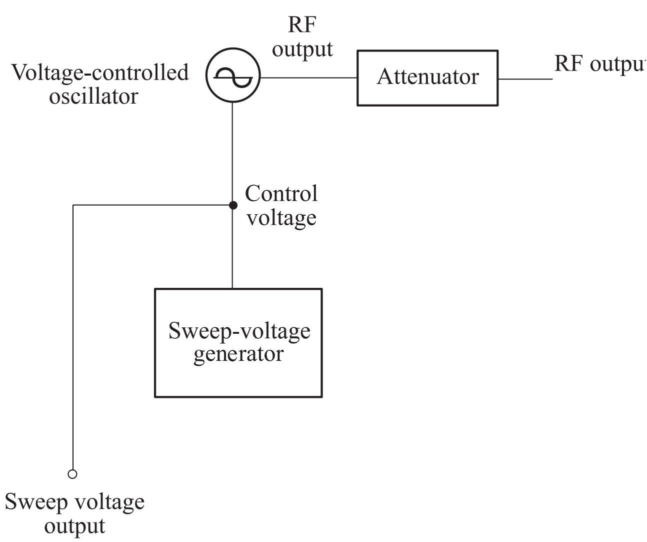


Figure 2.10 Sweep signal generator

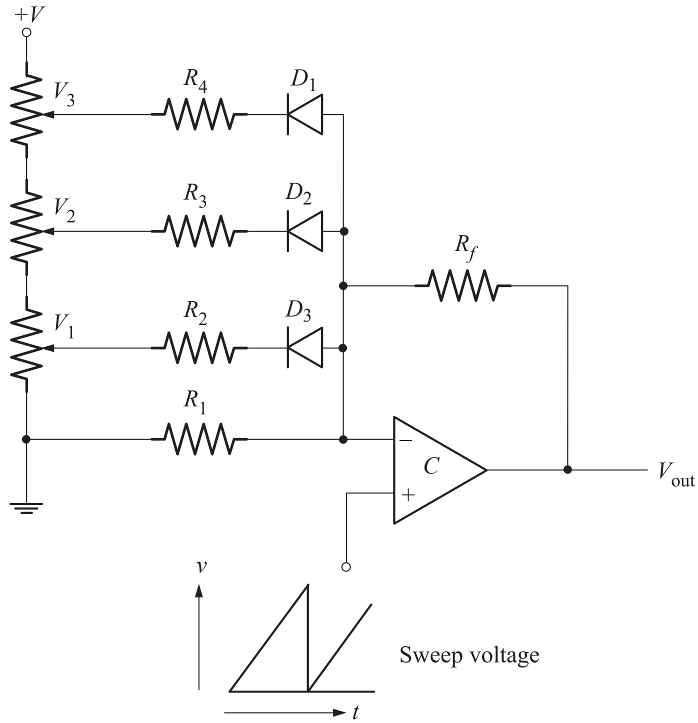


Figure 2.11 Linearising circuit for a sweep generator

The block schematic of a wide band sweep generator is shown in Fig. 2.12.

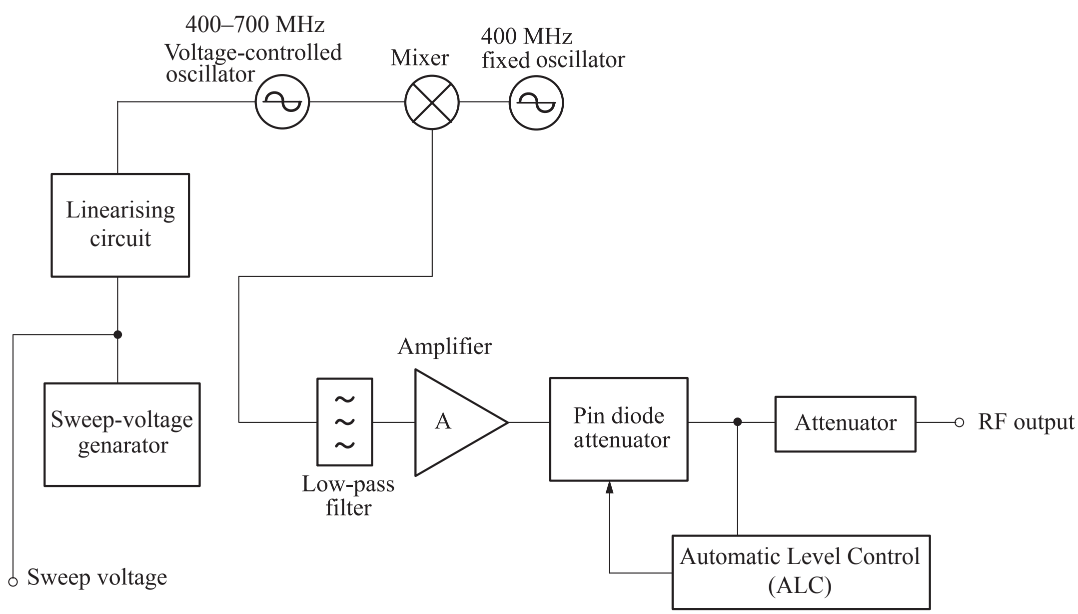


Figure 2.12 Block diagram of a wide band sweep generator

2.8 PULSE AND SQUARE WAVE GENERATOR

The difference between the pulse and square waveform is with respect to the duty cycle. The duty cycle is 50% for the square wave. The duty cycle is defined as the ratio of the average value of the pulse over one cycle to the peak value of the pulse. The average value and the peak value are inversely related to the time duration.

$$\text{Duty cycle} = \text{Pulse width/Period} = \frac{T_{ON}}{T_{ON} + T_{OFF}}$$

The duty cycle of a pulse waveform varies. It is not 50%. Very short duration pulses give a low duty cycle. Pulse generation can supply more power during its ON period than a square wave generator can. A stable or free-running multivibrator circuit is widely used to generate square wave and pulse waveforms.

The frequency range of the pulse generator (Fig. 2.13) is usually covered in seven decade steps from 1 Hz to 10 MHz. The duty cycle can be varied from 25% to 75%. Two independent outputs are available. The rise and fall time of the pulse is 5 ns at the 5 V peak amplitude. A 600 Ω source supplies pulses with a rise and fall time of 70 ns at the 30 V peak amplitude. The instrument can be operated as a free-running generator or it can be synchronised with external signals. The two current sources provide constant current for charging and discharging the ramp capacitor.

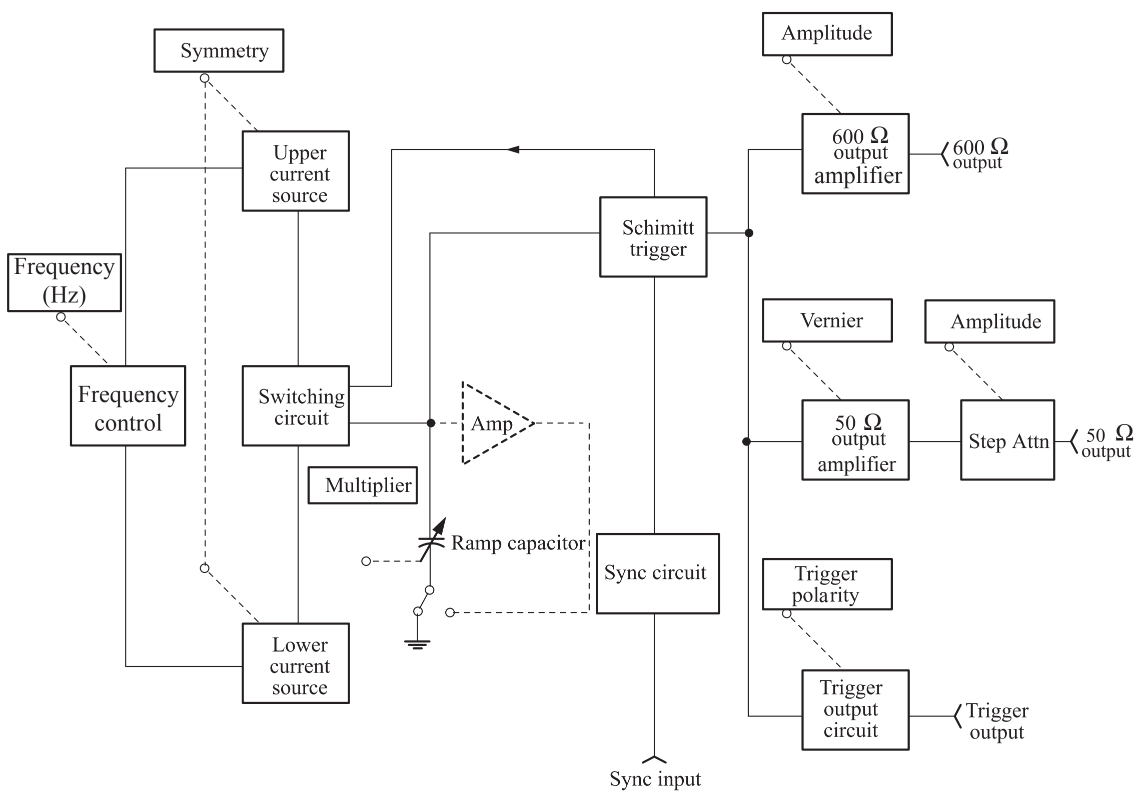


Figure 2.13 Block diagram of a pulse generator

2.9 FUNCTION GENERATOR

A function generator delivers a choice of different waveforms. The most common output waveforms are the sine, triangular, square, and saw-tooth waves. This equipment can usually supply output waveforms at very low frequencies. Since the low frequency of a simple RC oscillator is limited, a different approach is used in the circuit. The instrument can deliver sine, triangular, and square waves with a frequency range of 0.01 Hz to 100 kHz. The schematic diagram is shown in Fig. 2.14.

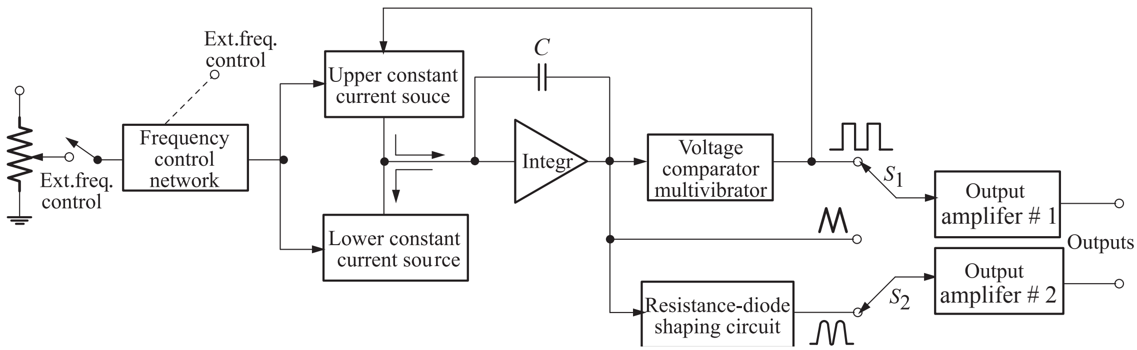


Figure 2.14 Block schematic of a function generator

The upper current source supplies a constant current to the integrator, whose output voltage increases linearly with time. The slope of the output voltage depends on the increase or decrease of the current. The voltage comparator multivibrator changes its state at a predetermined level. This change of state cuts off the upper current supply to the integrator and switches on the lower current supply. The lower current source supplies a reverse current to the integrator. Therefore, its output decreases linearly with time. When the output voltage reaches a predetermined level on the negative slope of the output waveform, the voltage comparator again switches and cuts off the lower current source while at the same time the upper current source is switched on. The integrator delivers the triangular waveform. The comparator gives the square wave output from the triangular waveform, and the sine wave is synthesised by a diode resistance network.

2.10 ARBITRARY WAVEFORM GENERATOR

An arbitrary waveform generator delivers signal fidelity at 2.7 Gb/sec to solve measurement challenges. The instruments can support two channels. The instrument combines world-class signal fidelity with high-speed mixed signal simulation, a powerful sequencing capability, and a graphical user interface with a flexible waveform editor to solve the toughest measurement challenges in the disk drive, communications and semiconductor design, and test industries.

Signal applications are built-in, which enable one to easily create standard waveforms for disk drive read channels, communications upto 2.7 Gb/S. The sample rate is of the order of 2.7 Gb/S, and real-world signals upto 1.35 GHz are generated. Analog signal bandwidths upto 2 GHz are provided the highest signal fidelity. There is provision for a direct external clock input, which allows jittered and non-jittered signals for a high-speed data stream timing margin test upto 2.7 Gb/sec. A synchronous operation mode supports two outputs. A waveform quick editor with 400 FS edge timing resolution delivers an output edge control with near real-time precision.

Two signals can be mixed digitally to support the disk drive noise performance test and the pre/de-emphasis serial data communication test. Real-time sequencing creates infinite waveform loops, jumps, patterns, and conditional branches.

2.10.1 Applications

1. Disk drive read/write design and test.
2. Communication design and test.
3. Arbitrary IF base band signals and pulse generation.
4. For mixed signal design and test.

2.11 VIDEO SIGNAL GENERATOR

This instrument is a multiformat analog and digital precision signal generator platform. It offers synchronizing pulse generation and test signal generation for a wide array of analog serial digital and high deformation formats.

Specification of AM/FM Generators

(Signal generators/programmable signal generators.)

Sl. No.	Parameter	Typical Values
1.	Frequency range	100 kHz to 260 MHz in eight ranges
2.	Output voltage	100 mV Max into 75 Ω
3.	Output control	0 db/10 db × 2 and fine control
4.	Frequency indication	Digital, 6 digits, 0.3^{11} LED
5.	Modulation	
	(a) Internal modulation	
	(i) Modulating frequency	1 kHz, 400 Hz selectable
	AM	Upto 80% adjustable
	FM	Dev. 0–100 kHz adjustable at 260 MHz
	(b) External modulation:	
	(i) Frequency	50 Hz to 20 kHz
	AM	0%–80% AM
	FM	0–100 kHz FM

	Level	15 V p-p
6.	Input supply	230 V AC ± 10% 50 Hz
	Model No./Type No.	xxxxx

Specification of a Function Generator

Model No./Type No. xxxxxx

Sl. No.	Description	Specifications (Typical Values)
1.	Waveforms	Sine (~), square (Ω), triangle (Λ), Ramp up (Λ), Ramp down (Λ), and AC
2.	Frequency characteristics	
	Sine	10–3 MHz
	Square	10–3 MHz
	Triangle	10–3 MHz
	Ramp	10–3 MHz
	Resolution	10–3 MHz
	Accuracy	< 50 ppm
	Temperature efficient	< 5 ppm (ppm: parts per million)
	Sine wave harmonic distortion	
	square wave	Upto 100 kHz: –50 db typical 100 MHz
		–40 db typical
	Overshoot	Rise time/Fall time : < 35 nsec
	Linearity	1%
	Triangle/ramp	Rise time
	Jitter	< 0.1% n peak output
	Power consumption	< 35 nsec
	Operating environment	30 W (average)
		10–50°C, 80% relative humidity

Example 2.1

In a Wien bridge oscillator $R_1 = R_2 = 55$ kΩ. $C_1 = C_2 = 800$ pf with usual notation. Determine the frequency of oscillations.

Solution

$$f = \frac{1}{2\pi RC}$$

$$f = \frac{1}{2\pi \times 55 \times 10^3 \times 800 \times 10^{-12}} = 3617.2 \text{ Hz}$$

Example 2.2

Design R and C elements of Wien bridge oscillators, for f_0 = 1 MHz.

Solution

$$f_0 = 1 \text{ MHz}$$

Let $$R = 200\ \Omega$$

$\therefore$ $$C = 1/2\pi \text{ fR} = \frac{1}{2 \times \pi \times 1 \times 10^3 \times 200 \times 10^3}$$

$\therefore$ $$C = 53 \text{ pf}$$

Example 2.3

In an ordinary phase shift oscillator $R_1 = R_2 = R_3 = 800\ \text{k}\Omega$, $C_1 = C_2 = C_3 = 100$ pf, with usual notation. Find the frequency of oscillations.

Solution

$$f_0 = \frac{1}{2\pi Rc\sqrt{6}}$$

$$= \frac{1}{2\pi \times 10^5 \times 8 \times 10^{-6} \times \sqrt{6}}$$

$$= 812 \text{ Hz}$$

Example 2.4

Determine the frequency of the transistor Colpitts oscillator with L = 100 mH, C_1 = 0.005 μF, C_2 = 0.01 μF.

Solution

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

$$= 3.33 \times 10^{-9}$$

$$L = 100 \text{ mH}$$

$$f_0 = 1/2\pi\sqrt{LC} = 275 \text{ kHz}$$

Example 2.5

Determine the oscillation frequency of a Hartley oscillator with L_1 = 100 mH, L_2 = 1 mH, and M = 50 mH (with usual notation), C = 100 pf.

Solution

$$C = 100 \text{ pf}$$
$$L = L_1 + L_2 + 2\text{ M} = 1200 \text{ mH}$$
$$C = 100 \text{ pf}$$
$$f = 1/2\pi\sqrt{LC} = 450 \text{ kHz}$$

Example 2.6

An amplifier with feedback has a voltage gain of 40. To produce specified output, the input voltage required without specified feedback is 0.1, with feedback I/P as 2.4 V to produce the same O/P. Calculate the value of the feedback ratio.

Solution

$$A_{fb} = 40$$
$$A = A_{fb} \times \frac{\text{I/P voltage rev}}{\text{I/P voltage without feedback}}$$
$$= 40 \times \frac{2.4}{0.1} = 960$$
$$A_{fb} = \frac{A}{1 - BA}$$
$$\beta = 1 - A/A_{fb}/A = -0.23958$$

2.12 SUMMARY

The different types of signal generators used in electronic measurements and instrumentation systems are standard signal generators, which produce a sine wave in audio frequency(AF) and radio frequency(RF) ranges. In these instruments, continuous wave (CW) and modulated RF signals can also be produced. Oscillator instruments also generate sine waves in the AF range and also in the RF range separately. The test oscillator is an instrument also generating sinusoidal waveforms with a calibrated attenuator and an output monitor. The function generator is an instrument that generates square, triangular, ramp, pulse, and sine waveforms. The sine wave generated in the function generator will have more distortion compared to the sine wave generated in oscillator instruments, because in function generator instruments, the sine wave is derived from a triangular wave through a diode resistor network. Pulse generator instruments generate pulses or rectangular waveform within variable duty cycle. Frequency and amplitude in these instruments can also be varied. A sweep generator instrument produces ramp waveforms within variable shapes. In selecting these instruments, the factors to be considered are as follows:

1. Frequency range.
2. Output voltage.
3. Resolution.
4. Accuracy.
5. Frequency stability.
6. Amplitude stability.
7. Distortion.

Points to Remember

- A standard signal generator instrument produces sinusoidal waveforms in AF and RF ranges. CW and modulated RF signals can also be produced.
- A standard signal generator is an oscillator with modulation capability.
- Oscillator instruments produce sine waves in the AF and RF ranges separately.
- A test oscillator is an instrument with a calibrated attenuator and an output monitor.
- The function generator instrument produces square, triangular, ramp, pulse and, sine waves. In this instrument, the *sine wave is synthesised from a triangular wave*.
- The sine wave produced in a function generator is not a pure sine wave compared to the one produced in an oscillator instrument. Distortion will be more in the sine wave generated in the function generator instrument.
- A pulse generator is an instrument generating pulses with variable duty cycles, frequency, and amplitude.
- A sweep generator instrument produces a *ramp waveform* with variable slopes.
- The term oscillator is used if the signal generator produces only sine waves.
- The term function generator is used if the instrument produces triangular, square waveforms, etc. in addition to sine waves.
- The specifications of these instruments are frequency range, output voltage, resolution, accuracy, distortion, frequency stability, and amplitude stability.
- In a sweep-frequency generator, a sweep-voltage generator is supplied internally. It provides the required frequency sweep of the waveforms.

Objective-type Questions

1. A standard signal generator produces a sine wave in __________ ranges.
2. Continuous wave (CW) and modulated RF signals can also be produced in __________ instruments.
3. Oscillator instruments are available in __________ ranges.
4. The oscillator instrument within a calibrated attenuator and an output monitor is __________.
5. A function generator instrument produces __________ waveforms.
6. A pulse generator produces rectangular waveforms with variable __________, __________, __________.
7. The resolution of a function generator instrument denotes __________.
8. The amplitude stability of a signal generator means the ability of the instrument to keep the amplitude of a signal constant at the set value when __________ is changed.
9. The term oscillator is used if the instrument produces __________.
10. The sine wave produced in a function generator instrument is __________ from __________ waveform.
11. Function generator instruments can produce waveforms with a frequency as low as __________.
12. The sine wave is generated from a triangular wave in the function generator through __________ network.
13. The harmonic content in an oscillator instrument must be __________.
14. Low spurious output is one of the desirable characteristics of an __________ instrument.
15. The output power level obtainable from a general-purpose audio-signal generator is usually of the order of __________.
16. The type of waveform generated by an oscillator instrument is __________.
17. The type of feedback used by an audio oscillator instrument is __________.
18. Wien bridge and phase shift oscillators are __________ type of oscillator circuits.

19. The number of R–C networks used in a standard R–C phase shift oscillator circuit is ________.

20. The main drawback of a crystal oscillator circuit is ________.

Review Questions

1. What are the considerations to be made in choosing a signal generator?
2. What are the different types of attenuators used in signal generators? Explain their operation.
3. With the help of a block schematic, explain the working of a frequency-synthesised signal generator.
4. Give the complete block diagram and explain the working of a pulse generator.
5. Draw the block diagram of a function generator and explain its working.

Unsolved Problems

1. With usual notation, in a Wien bridge oscillator, $R_1 = R_2 = 100$ kΩ. $C_1 = C_2 = 1000$ pF. Determine the value of frequency of oscillations, f_o.
2. Determine the component values in a Wien bridge circuit for $f_o = 10$ kHz where f_o is the frequency of oscillations.
3. In the case of an R – C phase shift oscillator circuit, the component values are $R_1 = R_2 = R_3 = 10$ kΩ. $C_1 = C_2 = C_3 = 150$ pF. Determine the value of frequency of oscillations f_o.
4. Determine the frequency of oscillations of Colpitts oscillator of BJT version, if $L = 120$ µH, $L_{RFC} = 0.5$ mH, $C_1 = 0.005$ µF, $C_2 = 0.02$ µF, $C_c = 20$ µF.
5. In the case of Hartley oscillator circuit, the values of the components are given as $L_1 = 200$ µH, $L_2 = 3$ mH, $M = 150$ µH, $C = 200$ pF. Determine the frequency of oscillator f_o.
6. Determine 'f_o' of a Wien bridge oscillator if $R_1 = R_2 = 100$ kΩ. $C_1 = C_2 = 500$ pf.
7. Determine the frequency of oscillations 'f_o' of Hartley oscillator, given $L_1 = 120$ µH, $L_2 = 5$ µH, $M = 70$ µH, and $C = 150$ pf.
8. The gain of an amplifier with feedback is 60. To get a particular output, the input voltage required without feedback is 1 V and with feedback is 2.4 V. Determine the value of the feedback ratio.

3

Signal Analysers

Introduction • Wave analyser • AF wave analyser • High-frequency wave analyser • Harmonic distortion • Heterodyne wave analyser • Tuned circuit harmonic analyser • Heterodyne harmonic analyser or wavemeter • Fundamental suppression harmonic distortion analyser • Spectrum analyser • Low-frequency spectrum analyser • Power analyser • Capacitance–voltage analysers • Oscillators • Summary

3.1 INTRODUCTION

In the previous chapters, different types of instruments that are used to generate various waveforms are described. To evaluate the quality of the waveform generated, distortion, and stability of the output of such a signal generator, instruments are required. Harmonic distortion (HD) analysers, wave analysers, and spectrum analysers are such instruments. The earlier instruments measured total HD without any indication of the frequency of the harmonic component. Wave analysers are more sophisticated analysers. They separate harmonic and non-HDs and evaluate each one.

3.2 WAVE ANALYSER

This instrument measures the relative amplitude of single-frequency components in a complex or distorted waveform. This can also be regarded as a frequency component that can be selected and its amplitude can be determined. The circuit is tuned to the particular frequency and all other components are rejected.

The signal to be analysed is applied first to the input attenuator. Here the meter range switch is set on the front panel. Then the signal is given to a driver amplifier. The output of the driver amplifier is given to a high-Q active filter. The filter consists of a cascaded arrangement of RC resonant sections and filter amplifiers. The pass band of the total filter section is covered in decade steps over the entire audio range by switching capacitors in the RC sections. Polystyrene capacitors with good tolerance are used for selecting the frequency ranges. Precision potentiometers are used to tune the filter to any desired frequency within the selected pass band.

The signal is further amplified in the final amplifier stage and is given to the meter circuit and to an untuned buffer amplifier. The buffer amplifier can be used to drive a recorder or an electronic counter. The meter is driven by an average-type detector and usually has several voltage ranges as well as a decibel scale. The attenuation characteristic of the instrument is shown in Fig. 3.1. The initial attenuation at one half is about 600 dB/octave. The attenuation at one half and twice the selected frequency is about 75 dB.

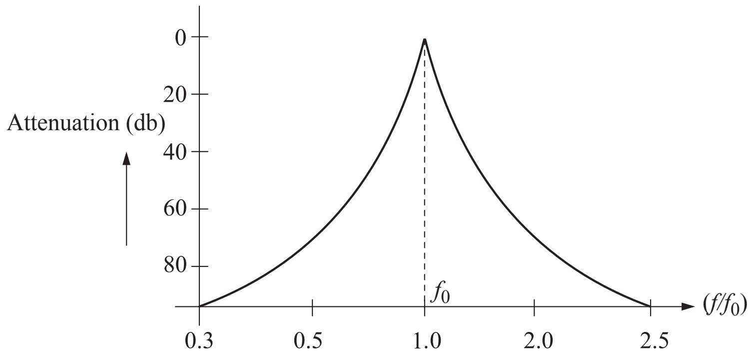
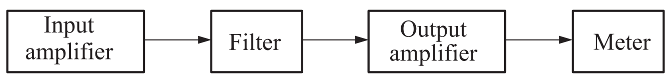
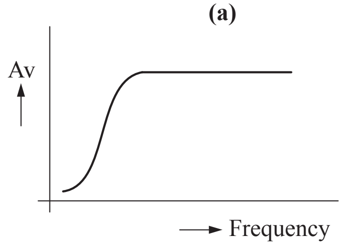


Figure 3.1 Attenuator characteristic

The signal input is given to an attenuator. A driver amplifier feeds the attenuated waveform to a *high-pass* (*active*) *filter*. It consists of a cascaded arrangement of *RC resonant sections* and *filter amplifiers*. The pass band of the total filter section is covered in decade steps over the entire audio frequency range (Fig. 3.2 (a) and (b)) by switching capacitors in the *RC* section. *Close-tolerance polystyrene capacitors* are usually used.



(a)



(b)

Figure 3.2 (a) Block schematic of an attenuator and (b) frequency response of a high-pass filter

A signal amplifier supplies the selected signal to the meter and to an untuned buffer amplifier. The buffer amplifier can be used to drive a recorder or an electronic counter. The bandwidth of the instrument will be 1% of the selected frequency.

Attenuation curve of the wave analyser: Attenuation at (0.707 or $\frac{E_2}{E}$) frequency (f_0) of the centre is about 75%.

3.3 AF WAVE ANALYSER

The block diagram is shown in Fig. 3.3. By varying *R* and *C*, the tuning frequency can be varied. Measurement in the MHz range is usually done with a wave analyser particularly suited for higher frequencies. The input signal to be analysed is heterodyned to a higher intermediate frequency (IF) range by an internal local oscillator. Tuning the local oscillator shifts the various signal frequency components into the pass band of the IF amplifier. The output of the IF amplifier is rectified and is applied to the metering circuit.

The operating frequency range of this instrument is from 10 kHz to 18 MHz. The input signal enters the instrument through a probe connector, which contains a unity gain isolation amplifier. After appropriate attenuation, the input signal is heterodyned in the mixer stage with the signal from a local oscillator. The output of the mixer is in the IF range and it is amplified by the IF amplifier. The amplified IF signal is then mixed again with a 30 MHz crystal oscillator signal, which results in information centered around the zero frequency (0–1500 Hz). An active filter with a controlled bandwidth then passes the selected components to the meter amplifier detector circuit.

A wave analyser is used in industrial applications, for example, in the reduction of sound and vibrations generated by machines. The noise level emitted by the machine is measured.

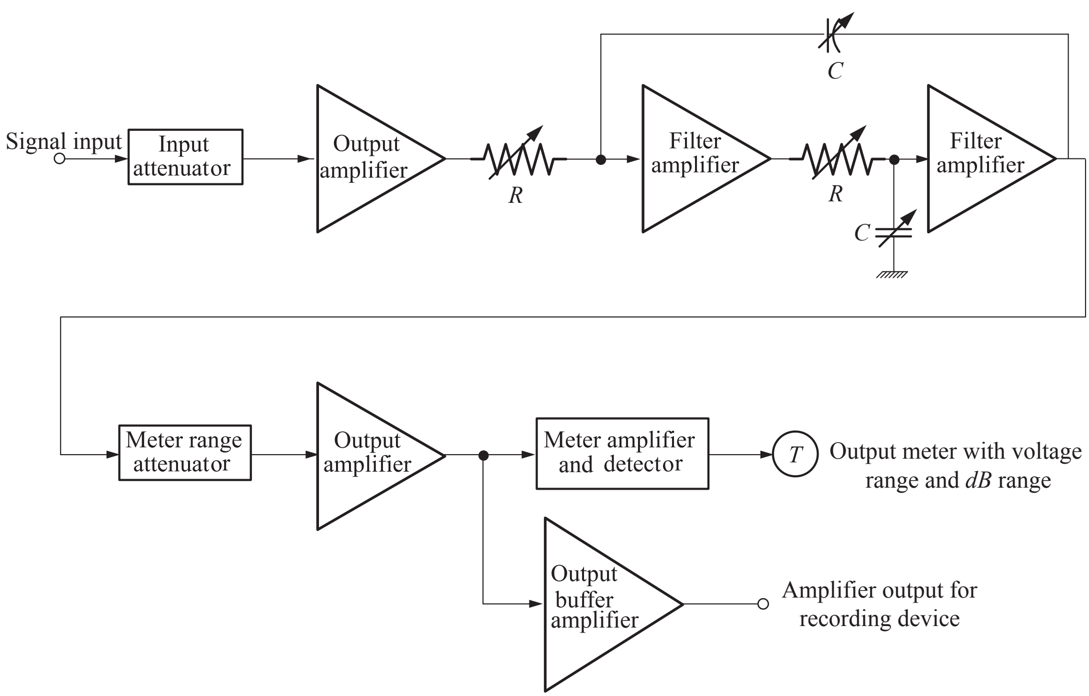


Figure 3.3 Block schematic for audio frequency wave analyser

3.4 HIGH-FREQUENCY WAVE ANALYSER

3.4.1 Frequency Mixers

In a super heterodyne receiver, the input amplitude modulated (AM) and radio frequency (RF) signal carriers have to be combined with a locally generated RF signal in order to produce a signal at a new carrier frequency (IF). The block diagram in as shown in Figs. 3.4 and 3.5.

Two methods are available:

1. Multiplication type in a wave device.
2. Addition type in a non-wave device.

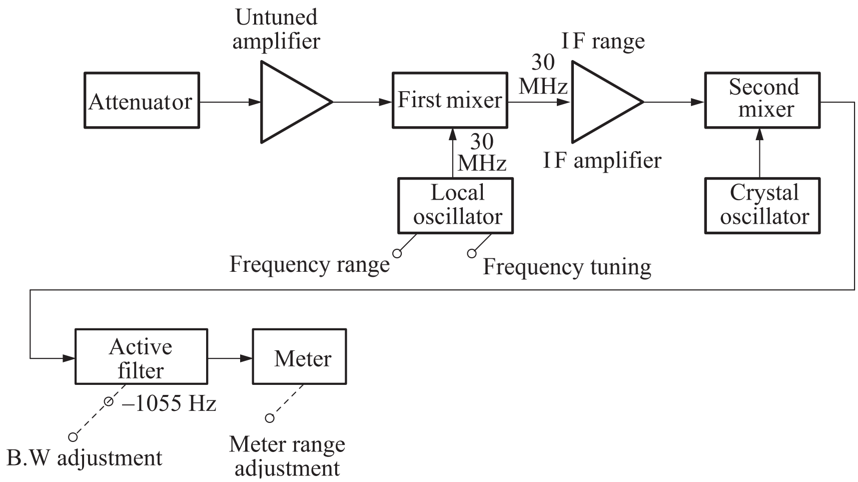


Figure 3.4 Block diagram for a high-frequency wave analyser

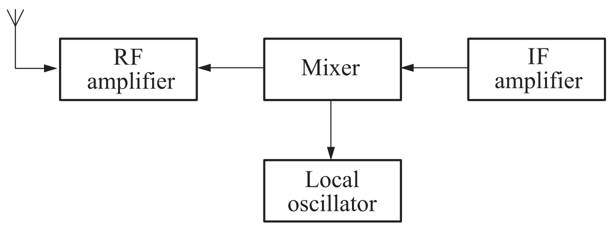


Figure 3.5 Simplified block diagram

Consider the second one. Let the local oscillator voltage be $V_o \cos \omega_0 t$. Using a device with a square law characteristic,

$$E_c \propto V_b^2$$

where E_c is the collector voltage and V_b the base voltage.

$$V_o = V_o \cos \omega_0 t + V_c(1 + m_a \cos \omega_m t) \cos \omega_c t$$

An amplitude-modified AM wave is the input to an RF amplifier.

The output of the device will give a large number of RF components including $(\omega_0 - \omega_c)$, $(\omega_0 + \omega_c)$, $(\omega_0 - \omega_c \pm \omega_p)$, $(\omega_0 + \omega_c \pm \omega_p)$ and so on. If the output load is tuned to $(\omega_c - \omega_c)$, the only significant voltage across is given by

$$i_t = KV_cV_o\left[\cos(\omega_0 - \omega_c)t \frac{\omega_0}{2} \cos(\omega_0 - \omega_c + \omega_p)t\right]$$

3.5 HARMONIC DISTORTION

There are three types of harmonic distortion (HD) circuits. They are as follows:

1. Tunable selective circuit.
2. Heterodyne type.

3. Distortion measuring component (which suppresses the product frequency and measures Root Mean Square (*RMS*) values of the distortion component).

3.5.1 Tunable Selective Circuit

The complex wave to be analysed is passed through an adjustable attenuator and is then applied to the selective amplifier, which is tuned to the frequency component to be determined (Fig. 3.6). (This is to find whether the component of that particular frequency is present or not). The output is indicated by a vacuum tube voltmeter (VTVM). If the amplifier has a constant gain over all frequencies, the attenuator can be set so that the VTVM gives 100% deflection for the fundamental frequency and the harmonics are expressed (magnitude) as a fraction of fundamental amplitude. The system can be calibrated using a standard signal generator. A tuned amplifier is generally a resonance-tuned amplifier. The advantage is that the gain can be totalised at a high frequency and can be made independent of frequency.

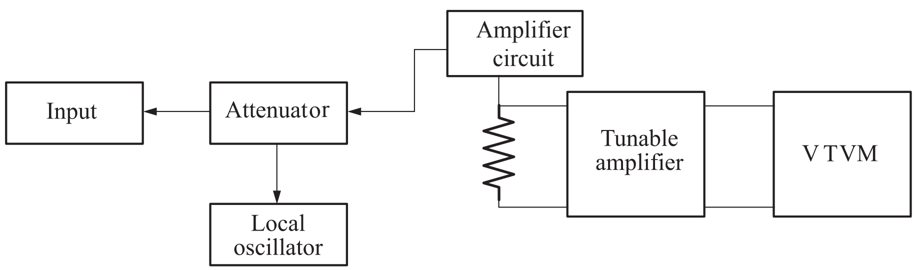


Figure 3.6 Block schematic of a tunable selective circuit

3.5.2 Disadvantages

1. At low frequencies the sizes of *L* and *C* are large.
2. Harmonics of signal frequencies are very close to each other. Therefore, it is difficult to distinguish between them.

3.5.3 Heterodyne Wave Analyser (Wavemeter)

The wave to be analysed is combined with a tunable local oscillator in a balanced mixer. The output of the mixer passes through a very highly selective multistage amplifier having a predetermined fixed response frequency, which is somewhat higher than any of the frequencies contained in the unknown wave (Fig. 3.7). In operation, the frequency of the local oscillator is adjusted so that the difference (heterodyne) in frequency produced by interacting with the desired component of the complex unknown wave is equal to the resonant frequency of the selective amplifier. Thus, the component to be determined has its frequency transformed to the predetermined value and is amplified and measured at this fixed frequency. Other frequency components present in the unknown wave transform to frequencies that are rejected by the selective amplifier. The frequency of the unknown component is determined from the local oscillator frequency that must be used to change the unknown component to the known fixed frequency.

The heterodyne wave analyser has an excellent frequency resolution, but the limitation in the mixer introduces spurious cross-modulation products.

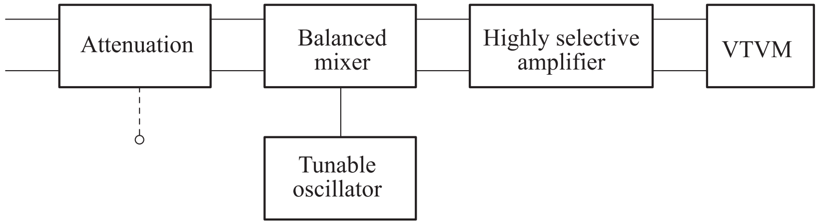


Figure 3.7 Block diagram of a heterodyne wave analyser

3.5.4 Fundamental Suppression Method of Distortion Measurement

RMS distortion of a wave is defined as the ratio of effective value of the harmonics to the *RMS* amplitude of the wave.

RMS distortion can be measured by *suppressing the fundamental frequency component* and then measuring the part of the wave that remains. Suppression of the fundamental frequency component can be done by the use of a high-pass filter, which is so designed that harmonics present in the pass band and fundamental frequency are severely attenuated. Another method is to use a bridge that is balanced for a fundamental frequency but is unbalanced for the harmonics.

A Wien bridge arrangement is employed in which the bridge is balanced for a fundamental frequency component and unbalanced for harmonics.

The distortion measuring instruments give only total distortion and not the amplitude of individual distortion components (Figs. 3.8, 3.9).

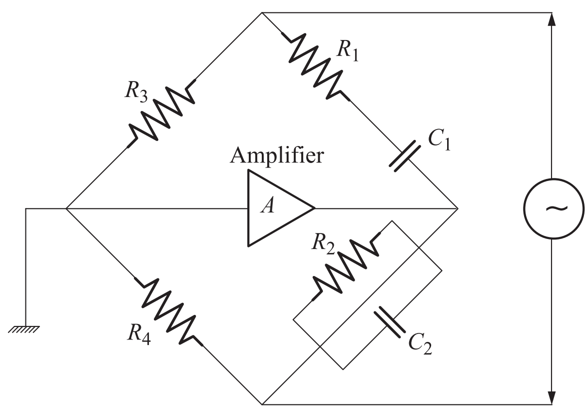


Figure 3.8 Bridge circuit for distortion measurement

The input to the Wien bridge circuit is along the distortion with the fundamental frequency completely suppressed.

Total HD or distortion factor

$$B = \sqrt{D_2^2 + D_3^2 + D_4^2 + \ldots}$$

where

$$D_2 = \frac{B_2}{B_1}, \quad D_3 = \frac{B_3}{B_1}, \quad D_4 = \frac{B_4}{B_1}$$

B_1 = Amplitude of the fundamental frequency

B_2 = Amplitude of the second harmonic

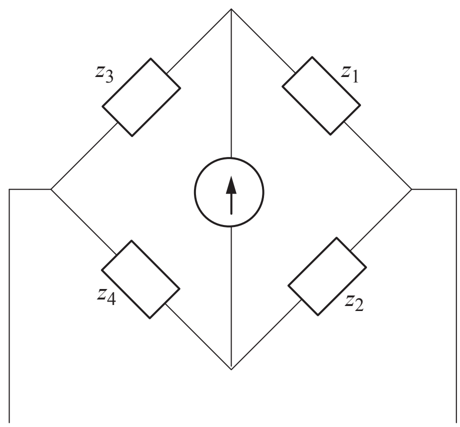


Figure 3.9 Bridge circuit for distortion measurement

3.6 HETERODYNE WAVE ANALYSER

This instrument is used in the MHz range. The input signal to be analysed is heterodyned to a higher IF by an internal local oscillator. Tuning the local oscillator shifts various signal frequency components into the pass band of the IF amplifier. The output of the IF amplifier is rectified and is applied to the metering circuit. The instrument using the heterodyning principle is called a *heterodyning tuned voltmeter*.

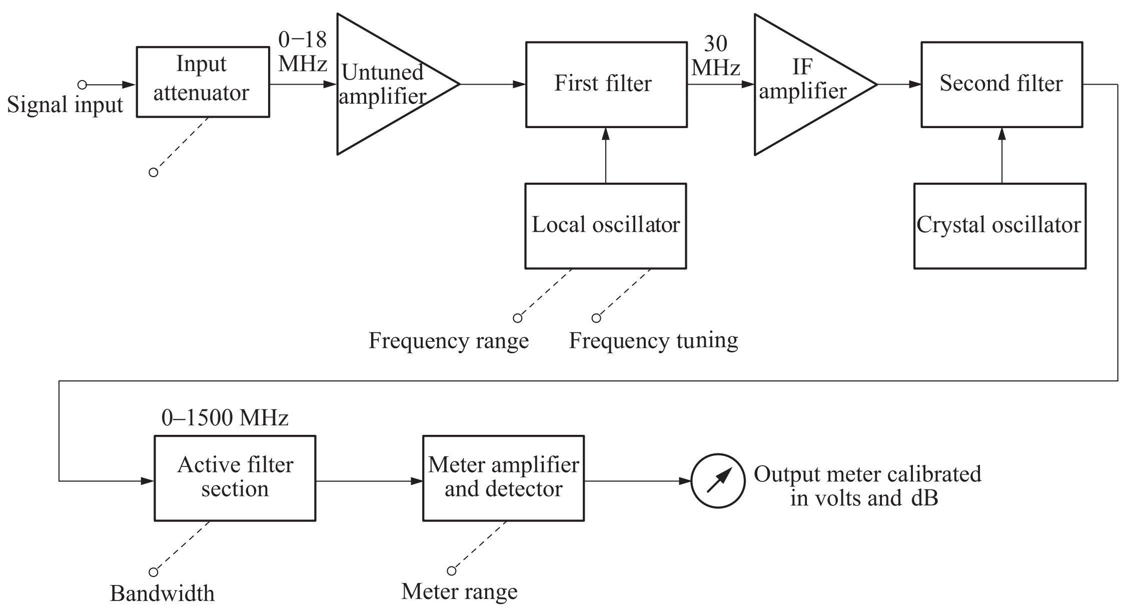


Figure 3.10 Block schematic of a heterodyne wave analyser

The block schematic of the wave analyser using the heterodyning principle is shown in Fig. 3.10. The operating frequency range of this instrument is from 10 kHz to 18 MHz in 18 overlapping bands selected by the frequency range control of the local oscillator. The bandwidth is controlled by an active filter and can be selected at 200, 1000, and 3000 Hz.

The input signal enters the instrument through a probe connector. The probe connector contains a unity gain isolation amplifier. After suitable attenuation, the input signal is heterodyned in the mixer stage with the signal from a local oscillator. The output of the mixer forms an IF that is uniformly amplified by the 30 MHz IF amplifier. This amplified IF signal is then mixed again with a 30 MHz crystal oscillator signal, which results in information centered around a zero frequency. An active filter with controlled bandwidth and symmetrical slopes of 72 dB per octave, then passes the selected component to the filter amplifier and the detector circuit. The output from the meter detector can be read on a decibel scale or can be applied to a recording device.

3.6.1 Applications of Wave Analysers

1. In electronic measurements.
2. Sound and vibration analysers of machines.
3. Harmonic distortion (HD) analysers.

3.6.1.1 HD analyser. If a sinusoidal input is given to an amplifier, the output must also be a true sine wave. However, because of the non-linear characteristics of active devices such as bipolar junction transistor, junction field-effect transistor, or MOSFET and other passive elements, the output will be distorted. The non-linear behaviour of the circuit elements introduces harmonics of the fundamental frequency in the output waveform. The resultant distortion is called as *Harmonic Distortion* (*HD*). Distortion is measured as a ratio of the amplitude of the harmonic to that of the fundamental frequency, expressed as a percentage:

$$D_2 = \frac{B_2}{B_1}; \quad D_3 = \frac{B_3}{B_1}; \quad D_n = \frac{B_{n3}}{B_1}$$

where

B_1 = amplitude of the fundamental frequency

B_2 = amplitude of the first harmonic

B_3 = amplitude of the second harmonic

B_n = amplitude of the *n*th harmonic

D_2 = distortion of the second harmonic

D_n = distortion of the *n*th harmonic

The total HD factor is defined as

$$D = \sqrt{D_1^2 + D_2^2 + D_3^2 + \cdots + D_n^2}$$

3.7 TUNED CIRCUIT HARMONIC ANALYSER

A series-resonant circuit consisting of L and C is tuned to a specific harmonic frequency. This harmonic component is transformer coupled to the input of an amplifier. The output of the amplifier is rectified and is applied to a meter circuit. The reading on the meter is noted. The resonant circuit is tuned to another frequency and the next reading is taken and so on. The parallel resonant circuit consisting of L_1, R_1, and C_1 provides compensation for the variation in the AC resistance of the series-resonant circuit and also for the variations in the amplifier caused over the frequency range of the instrument. The schematic of the instrument is shown in Fig. 3.11.

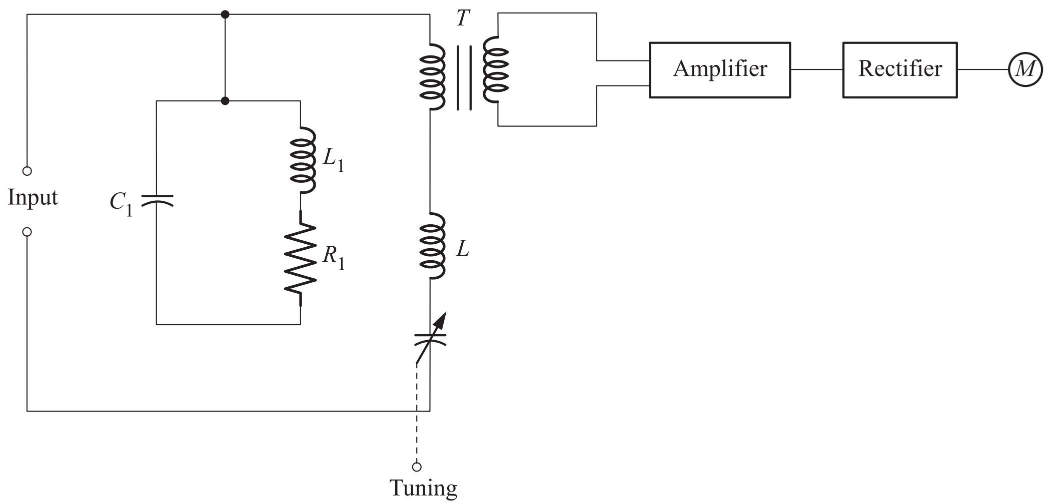


Figure 3.11 Tuned circuit harmonic analyser

The disadvantages of this instrument are as follows:

1. For low frequencies, very large values of L and C are required. Hence, physical size of the components is too large.
2. Harmonics of the signal frequency are very close in frequency. Hence, separating and distinguishing them is difficult.

3.8 HETERODYNE HARMONIC ANALYSER OR WAVEMETER

This instrument employs a highly selective fixed-frequency filter. The block schematic of the instrument is shown in Fig. 3.12.

The output of a variable frequency oscillator is heterodyned or mixed with each harmonic of the input signal and either the sum or the difference frequency is made equal to the frequency of the filter. Each of this harmonic frequency is converted to a constant frequency. Therefore, a highly selective filter of the quartz crystal type can be used. The mixer circuit is a balanced modulator. The balanced mixer produces low harmonic distortion (HD). This is another advantage of this instrument. Good selectivity is obtained by using a quartz crystal filter or inverse feedback filter. In the case of some heterodyne analysers, the meter reading is calibrated directly in terms of voltage. Direct reading instruments of the heterodyne type are also called *frequency-selective voltmeters.*

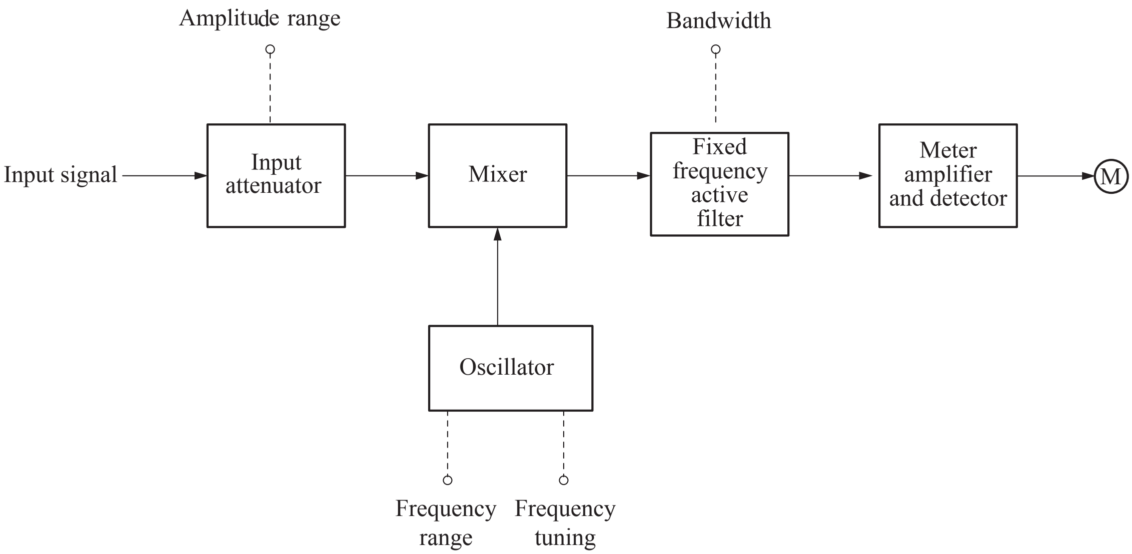


Figure 3.12 Heterodyne harmonic analyser

3.9 FUNDAMENTAL SUPPRESSION HARMONIC DISTORTION ANALYSER

This instrument is used when the total HD is to be measured rather than the HD of each component. The input is applied to a network of filters that suppresses or rejects the fundamental frequency but passes all the harmonic frequency components. The advantages are as follows:

1. The HD generated within the instrument is very small and can be neglected.
2. Selectivity requirements are not stringent. Only the fundamental frequency component must be suppressed.

The block diagram of the instrument is shown in Fig. 3.13. The instrument consists of four major sections:

1. Input circuit with impedance converter.
2. Rejection amplifier.
3. Metering circuit.
4. Power supply.

The impedance converter provides a low-noise high-impedance input circuit. The rejection amplifier rejects the fundamental frequency of the input signal and passes the remaining frequency components. The metering circuit provides the visual indication of total HD in terms of the percentage of total input voltage. The instrument can be operated in two modes:

1. Voltmeter.
2. Distortion meter.

In the voltmeter mode, the instrument works like a simple AC voltmeter. When the function switch is in the distortion position, distortion measurement can be made.

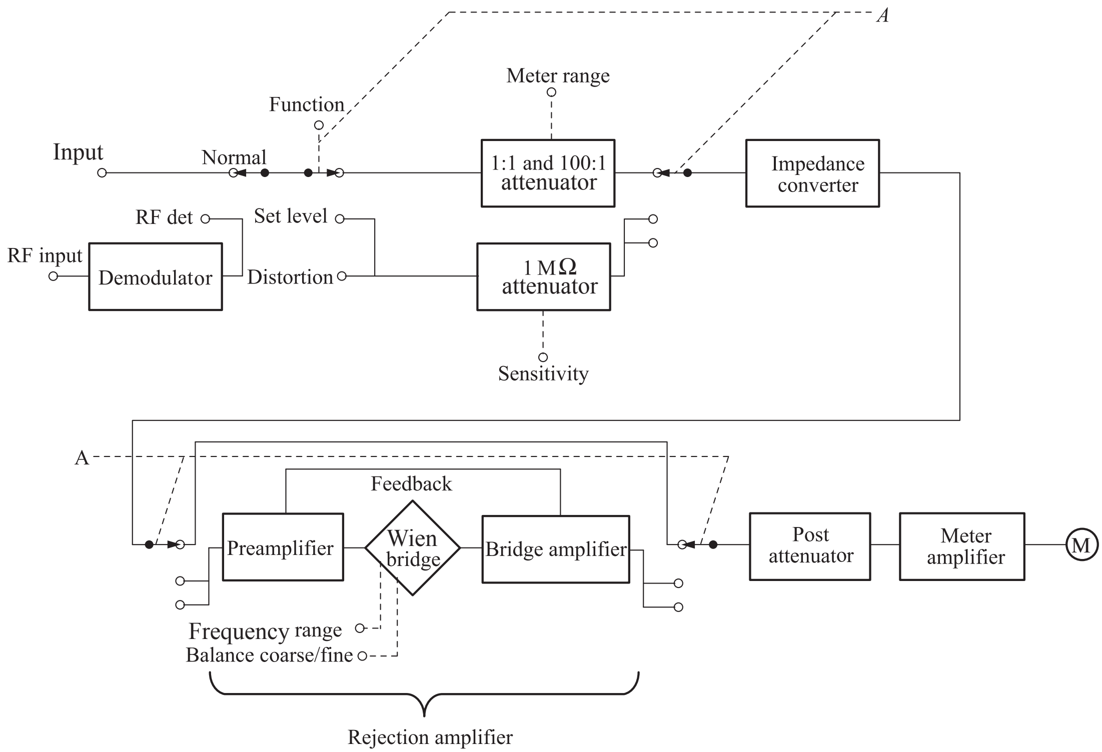


Figure 3.13 Fundamental suppression harmonic distortion analyser

3.10 SPECTRUM ANALYSER

The distribution of the amplitude of voltage or (voltage)^2 [energy] with frequency is defined as the spectrum of an electrical signal. A spectrum analyser is an instrument that gives a plot of such a distribution. Broadly, there are two types of spectrum analysers:

1. Low-frequency spectrum analysers.
2. High-frequency spectrum analysers.

Further, there are two types of high-frequency spectrum analysers:

(i) Audio frequency analyser.
(ii) Radio frequency analyser (Fig. 3.14).

The block diagram of a general-purpose spectrum analyser is shown in Fig. 3.15.

The typical frequency range of a spectrum analyser is from 10 kHz to 300 MHz. A spectrum analyser is similar to an up-converting super-heterodyne receiver. The input of the spectrum analyser is first converted to an IF higher than the highest input frequency. The band of frequencies of the input image is moved with a low-pass filter. The spectrum analysers may have 1 kHz selectivity at its narrowest setting. However, this selectivity will not be obtained at high frequencies. Therefore, first IF is heterodyned to a lower frequency.

The frequency of the first local oscillator is swept electronically using varactor diodes in a manner similar to the sweep generator. The span of frequency that is swept is called the *dispersion* of the analyser.

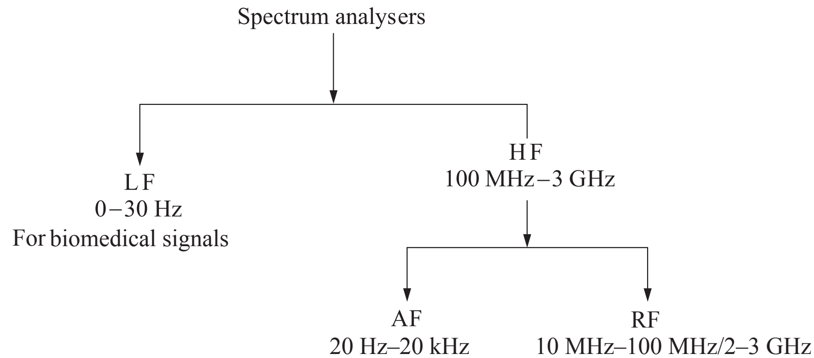


Figure 3.14 Classification of spectrum analysers

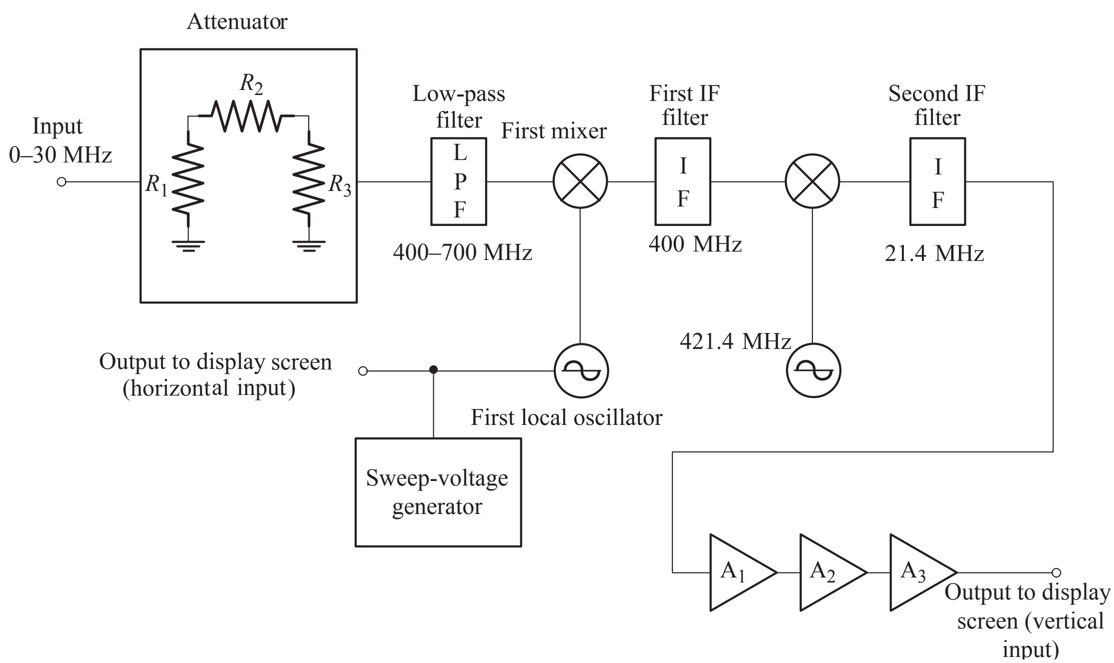


Figure 3.15 Block diagram of a general-purpose spectrum analyser

The voltage applied to the varactors must pass through a correcting circuit to cancel the non-linearities. The spectrum analyser is usually required to sweep narrow frequency ranges.

Two types of frequency instabilities will occur in the instrument. They are as follows:

(a) *Long-term instability*: This is because of the drift of the frequency of the first local oscillator. This will appear as the movement of the spectrum across the spectrum analyser screen.

(b) *Phase noise*: This is because of the noise voltage in the tuned circuit or noise voltages picked up by the varactor circuit. Owing to this, the frequency changes rapidly. The principle of operation of both low and high-frequency spectrum analysers is the same. The frequency ranges of different blocks of the instrument differ.

3.10.1 Characteristics of a Spectrum Analyser

Frequency resolution: To separate two signals closely spaced. This depends on the drift and noise of a local oscillator.

Bandwidth: The frequency range over which signal analysing can be done.

Sweep desensitisation: This is the maximum rate of scanning of the spectrum without affecting the resolution. This depends on the maximum slope of the saw-tooth waveform.

Sensitivity: It is the smallest amplitude of the signal that can be detected.

3.10.2 Applications of a Spectrum Analyser

Spectrum analysers are used in pulse studies, in the measurement of attenuation, and to view the burst of RF energy in radar applications. In an RF instrument covering 1 kHz to 100 MHz, the spectrum window ranges from 100 MHz down to 2 kHz, with signals that are 100 Hz apart capable of separate identification. Input Z is 50 Ω.

With spectrum analysers, we can study the bandwidth, spurious signal generation, effects of various types of modulation, etc.

3.10.3 Basic Spectrum Analyser

The block schematic is shown in Fig. 3.16. A low-pass filter with a cut-off level above 1 GHz at the input suppresses spurious signals.

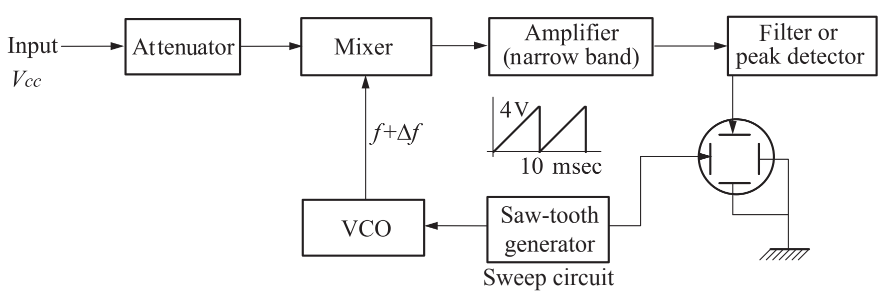


Figure 3.16 Block schematic of a spectrum analyser

A saw-tooth provides linear time axis, which is useful in many cases. A saw-tooth generator supplies a saw-tooth voltage to the frequency control element of the voltage-tuned oscillator, which then sweeps through its frequency band at a recurring linear rate. The same saw-tooth voltage is simultaneously applied to the horizontal deflection plates of the cathode ray tube (CRT) in the cathode ray oscilloscope (CRO). The RF signal under investigation is applied to the input of the mixer stage. The local oscillator is swept through its frequency band by the saw-tooth generator, amplified with the input signal to produce the required IF. An IF component is produced only when the corresponding component is present in the RF input signal. The resulting IF signals are amplified, detected, and then applied to the vertical deflection plates of the CRT producing a display of amplitude versus frequency.

When an RF carrier wave is modulated, the resulting waveform comprises not only the original carrier frequency but also side bands above and below the carrier frequency. The distribution of these frequency components plotted as a graph of amplitude (voltage versus frequency) versus frequency is termed spectrum. Spectrum analysers provide a plot on the screen of a CRT portraying a graph of amplitude versus frequency. The local oscillator is electronically swept back and forth between two frequency limits at a linear rate. At the same time, the spot on the CRT is caused to move horizontally across the face of the tube in synchronisation, so that the horizontal position is a function of the local oscillation frequency (Fig. 3.17). As the spot moves horizontally, it is deflected upwards in proportion to the amplitude of the voltage delivered by the detector and video amplifier. The local oscillator is a klystron at the microwave frequency and a BJT oscillator for low frequencies.

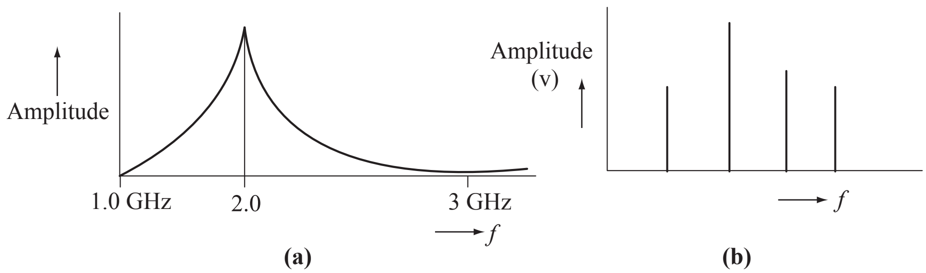


Figure 3.17 (a) Frequency response and (b) output

The input signal is fed into a diode mixer, which is driven to saturation by a strong signal from the local oscillator. This oscillator is linearly tunable over the range 2–3 GHz. The input mixer multiplies the input signal and the local oscillator signal together and so provides two signals at its output, which are proportional in amplitude to the input signal, but of frequencies that are the sum and difference between the frequency of the input signal and the local oscillator. The IF amplifier is tuned to a narrow band around 2 GHz. As the local oscillator is tuned over the range from 2 to 3 GHz, any input signals that are separated from the local oscillator frequency by 2 GHz will be converted to the IF band, passed through the IF amplifier, rectified in the detector, and will produce a vertical deflection on the CRT (suppose the input signal frequency is 1 GHz, local oscillator frequency is 2 GHz, mixer output frequency is 2 + 1 = 3 GHz or 2–1 = 1 GHz. Hence, both 1 and 3 GHz signals will be attenuated).

The input signal whose frequency spectrum is required is given to the attenuator if the amplitude of the signal is very large. A saw-tooth generator produces a waveform where the voltage rises to 4 V in 10 msec. (These are typical values.)

The output frequency voltage-controlled oscillator (VCO) varies with the applied voltage to the varactor diode. The capacitance of the varactor diode varies with the reverse bias applied to it. As V varies C also varies and, therefore, the frequency of oscillation of the LC oscillator circuit also varies. This variation is not linear over the entire range. Suppose that the variation of frequency VCO is linear in the range of 15–16 MHz corresponding to voltage variation 10–14 V of the supply. Therefore, the VCO has to be operated only in this range. Otherwise, calibration will be difficult, therefore, the frequency range of the spectrum analyser is 1 MHz. The amplitude of the different signals in the complex input up to a frequency of 1 MHz will be displayed on the CRO. Since the frequency variation is linear for a variation of 4 V input, the saw-tooth generator is designed to get a linear variation for 4 V. Since the actual voltage must be 10–14 V, the saw-tooth wave is superimposed over a 10 V AC to get a linear variation of voltage from 10 to 14 V. Therefore, the function of the sweep circuit is to automatically vary or sweep the voltage of VCO from 10 to 14 V linearly, so that the frequency of oscillation of VCO also varies linearly from 15 to 16 MHz.

Suppose that the input signal is 15 MHz fixed. The VCO gives a 15–16 MHz signal. The mixer gives output signals, which are the sum and difference of these two signals. That is, the output signals will be 15 + 15 = 30 MHz signal and 15 – 15 = 0 Hz signal. We want a spectrum analyser whose frequency range is 0–1 MHz. Therefore, the high-frequency signals of 30 MHz should be filtered; hence, after the mixer, we must have low-pass filters, which will alternate signals above 1 MHz. There is an amplifier to amplify and the signal is fed to the *Y*-plates of the CRO. To the *X*-plates the same sweep generator output is given. Another block diagram of a spectrum analyser is shown in Fig. 3.18.

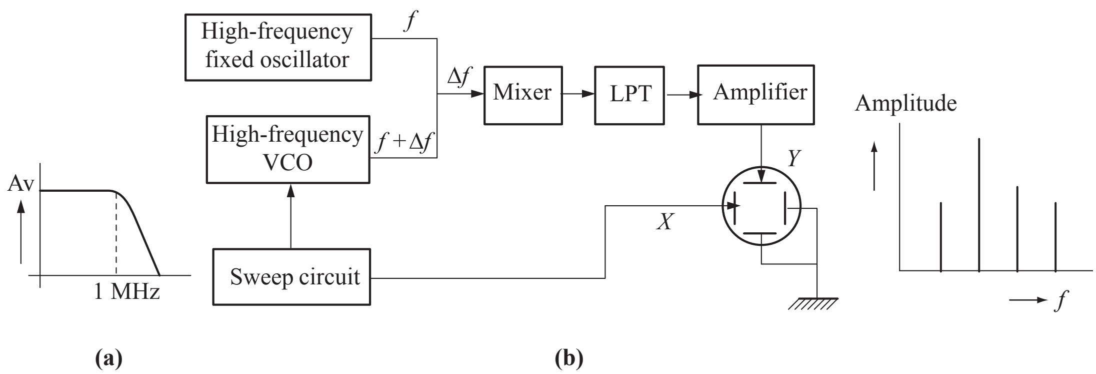


Figure 3.18 Spectrum analyser—another block schematic

The block schematic of an RF spectrum analyser is shown in Fig. 3.19.

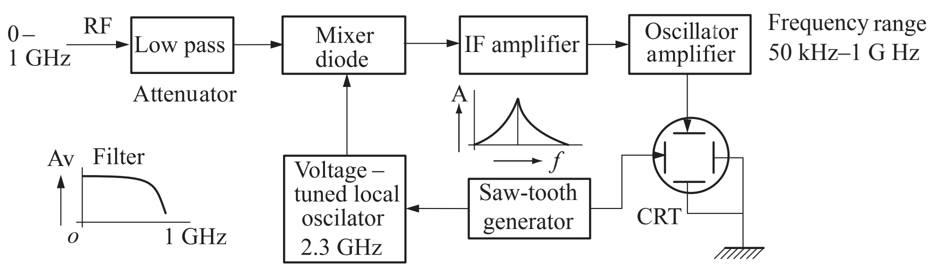


Figure 3.19 Block schematic

3.10.4 Factors to be Considered in a Spectrum Analyser

In order to secure proper operation of the spectrum analyser, attention must be given to certain design considerations:

1. *Frequency resolution*: Frequency resolution is the ability of the spectrum analyser to separate signals that are closely spaced in the frequency. This is determined by the bandwidth or the selectivity of the IF amplifier and the drift and noise of a local oscillator. The oscillator must have greater stability.
2. *Sweep desensitisation*: If the spectrum is being scanned at a very fast rate, then the sensitivity, frequency resolution, and the amplitude of the signal will suffer. This is because during a scan, the signal must lie in the band pass of the IF amplifier filter, long enough to allow the amplitude of the signal in the filter to build up to the proper value.

3. *Bandwidth*: The bandwidth of the IF amplifier should be narrow. Then the sweep speed should be very low in order to allow time to build up the voltage in the receiver circuit and flicker of the CRT image. (Since normal electrical supply frequency (50 Hz) is very less, flickering will be present).
4. *Frequency*: The centre frequency of the IF amplifier. There is no selectivity ahead of the mixer.

3.11 LOW-FREQUENCY SPECTRUM ANALYSER

The frequency range of a low-frequency spectrum analyser is typically 7–30 Hz. These are very uncommon. Such analysers are employed in the analysis of biological electrical outputs. The block diagram is as shown in Fig. 3.20.

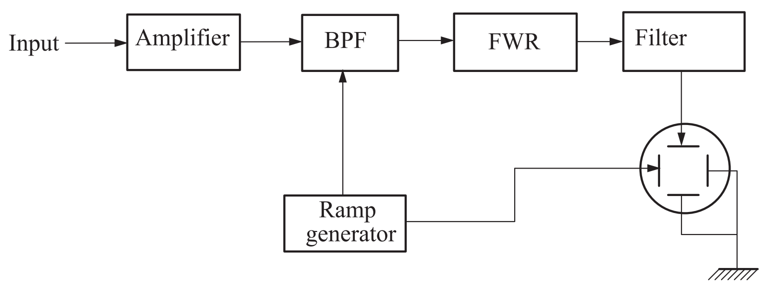


Figure 3.20 Block schematic of a low-frequency spectrum analyser

Because of difficulties in the mixing of low frequency, the RF spectrum analyser cannot be used for the analysis of low-frequency signals. The difference between the LF and the RF spectrum analyser is that there is no local oscillator in LF spectrum analyser. The amplifier can consist of a number of stages, with an overall gain of 200, depending upon the magnitude of the input signal. The centre frequency of the band pass filter (BPF) can be varied. The output of BPF is full-wave rectified and filtered to give an absolute value of the amplitude of the frequency component present at that moment. The centre frequency of BPF is varied at a rate by the ramp and the same is given to *X*-plates of a CRO.

Sensitivity: It is the ability of the analyser to measure small signals. It is limited by the internally generated noise. The noise frequency is 25 dB at the nodes.

Dynamic range: This expresses the ability of the spectrum analyser to display true spectra without any distortion. If the signal levels are within the dynamic range, the spurious signals that result from the distortion of the large signals can either weaken the small signals or erroneously appear as small signals. The dynamic range of the instrument is free of spurious signals.

3.11.1 Applications

1. Pulse width and repetition rate measurements.
2. FM deviation measurement.
3. FR interface testing.

3.12 POWER ANALYSER

These instruments are used for

1. Testing and verifying correct operation of motors.
2. Checking transformer efficiency.
3. Verifying power-supply performance.
4. Measuring the effect of neutral current.

These features include harmonic analysis, power measurements, and Pulse Width Modulation (PWM) motor drive triggering. These instruments can also be used as oscilloscopes, which enable troubleshooting and verification of complicated electronic control circuits controlling the high-voltage power electronic circuits.

The bandwidth of the oscilloscopes will be about 100 MHz. The sampling rate per channel is typically 500 samples/sec. These instruments can also be used as digital multimeters with a data logger. Some models in this category can measure harmonics upto the thirty-first harmonic with a fundamental frequency ranging from 30 to 450 Hz.

THS730A/THS720A/THS720P are oscilloscopes/DMMs/Power analysers from Tektronix.

3.12.1 Communications Signal Analyser

Communication signal analysers are used for design, evaluation, and testing of Datacom and Telecom components, transceiver subassemblies, and transmission systems. This instrument generates measurement results, not only with data but also with time and amplitude histograms, mask testing, and statistical measurements. The instrument can also provide a communication-tailored measurement set that includes jitter, noise, duty cycle, overshoot, undershoot *Q*-factor, mean optical power, and amplitude measurements for both return to zero and non-return to zero signals. Gigabit Ethernet and other optical and electrical standard support and conformance testing can be done. This instrument also supports a large family of optical and electrical plug-in modules.

Some of the general features are as follows:

- Wide bandwidth (typically DC to 70 GHz).
- Fast acquisition rate.
- Very low jitter.
- Support to wide range of optical and electrical standards.
- Automated communication measurements.

Applications of communication signal analysers are as follows:

1. Design/verification of Telecom and Datacom components and systems.
2. Manufacturing/testing for IEEE conformance.

Optical and electrical modules support clock recovery for a simple connection to a device under test with optical, electrical, or electrical differential signals. The optical modules provide complete optical test solutions for standards of Telecom from 155 Mbps to 43 Gbps.

The electrical modules include acquisition modules with bandwidths upto 7 GHz typically. These instruments ensure most accurate acquisition of high-speed communication signals. Multiprocessor architecture with dedicated per slot digital signal processors are used. CSA 8200 is a Tektronix communication signal analyser.

3.12.2 Logic Analysers

These instruments are used for real-time digital system analysis. A single synchronous base state and timing analysis can be done to debug and to verify multiple-base digital systems. For advanced processors and bases, some of the advanced analysers are 800 MHz state acquisition or a 1.25 Gb/S data rate. An oscilloscope and a single logic analyser probe is used to capture, display, and analyse time-correlated analog and digital measurements.

3.12.3 Network Monitoring System

It provides a centralised remote monitoring system that gives a broad view of the entire network. It offers complete network management and supervision capabilities. They can be used for GSM and GPRS-non-intrusive network monitoring. The protocol library of the system includes more than 150 protocols and protocol variants for both fixed and mobile 2G, 2.5G (second generation) and 3G (third generation) networks. The system architecture can range from a small single-site system with few links to a fully distributed multisite system with thousands of links.

The system capabilities include:

- Real-time status monitoring.
- Traffic monitoring and signal measurements.
- Signalling accounting.
- Call data record (CDR) and transaction data record.
- Billing verification.
- Call/procedure trace.
- Protocol analysis.
- Fraud detection.
- Carrier/destination traffic monitoring and analysis.
- Performance analysis of intelligent network services.
- Roaming supervision.
- Routing verification.
- GPRS monitoring.

3.12.4 System Architecture

The system architecture includes remote monitoring probes connected in a communications network (a wide area network (WAN) with TCP/IP) to control units. A large number of user sites can be supported from a single central unit. The system physically links to the network at the monitoring site interface located in a network node.

3.12.5 Features

- Revenue losses can be prevented because of fraud with proactive detection capabilities.
- Losses can be minimised because of billing errors with an accurate generation of customer cell detail records for every call and call attempt.
- Previously unbilled services can be located and billed.

- Quality service can be provided.
- Quality of service solution can be provided to operations and maintenance.
- Maintenance efficiency can be increased by centralising operations and enabling fast and accurate troubleshooting and planning.
- Revenue-customers can be increased by increasing the network uptime.

3.12.6 Applications

- Complete monitoring system management and network supervision.
- Billing verification.
- Signalling protocol and traffic accounting.
- Call trace.
- Fraud detection and management applications.

Monitoring sites-that can be distributed anywhere in the network contain one or more racks equipped with:

- Monitoring probes having different interface boards to access network lines.
- Mass storage of signalling raw data.
- WAN connectivity hardware.
- Global Positioning System (GPS) antenna.

The monitoring probes provide simultaneous monitoring of a number of signalling data links, collecting and decoding signalling information and alarms.

Statistical measurements and counters are provided at different intervals based on specific protocol messages to evaluate the performance of a signalling network. Call or transaction-selected information is provided by the system to allow CDRs for every call or call attempt in the network.

Data collected by different monitoring probes are to be synchronised across the entire network to ensure a correct interpretation of the information. A GPS is provided in each of the monitoring sites so that information gathered by the central unit can be analysed in the correct sequence by examining the signalling flow, the monitoring probes generated in real time, protocol data, statistic counters, and alarms are given as per requirement. All the messages are time spaced and displayed in real time to the operator, which can choose the level of details with which the information is displayed. Besides being displayed in real time, the information can be stored for subsequent off-time analysis.

3.13 CAPACITANCE–VOLTAGE ANALYSERS

Capacitance–voltage analysers measure and give capacitance versus voltage and capacitance versus time characteristics of semiconductor devices. A high-frequency signal of 100 kHz or 1 MHz is applied to test the p–n or Schottky junction and metal insulator semiconductor devices for device characterization and process control. C–V results are highly correlated with performance parameters of functional devices such as field-effect transistors, memory cells, charge-coupled devices (CCDs), and isolation structures.

The instrument provides:

- 0.1 femto farads (0.1×10^{-18} F) typically sensitive to test small devices.
- Ranges upto 20 mF (at –100 kHz) to test large, leaky, or forward-biased devices.

- Test signal voltage of 15 mV *RMS*.
- Measurement rates of 1–1000 readings per second can be selected. Speed/resolution trade-offs can be optimized.
- Correction can be made for transmission line errors because of connections.

C–V and C–t (capacitance V and time) data are used to determine important characteristics such as semiconductor doping profiles, threshold voltage, oxide characteristics, mobile ion density, interface trap density, and minority carrier lifetime. The internal voltage source can supply upto 50 mA from –20 V to +20 V with a 5 mV resolution. The voltages can be applied in the form of a DC staircase, dual staircase, and pulse form. There is provision to apply upto 200 V using an eternal source.

1. The sensitivity for capacitance is typically 0.1 fF (0.1×10^{-15} F).
2. The sensitivity for time parameter is typically 0.1 ms.
3. The measurement range for capacitance is typically 20 nF.
4. The measurement range for time is typically 20 ms.
5. Usually a 15 mV *RMS* test signal is used for testing the devices. Testing semiconductor devices with large test signals can cause a curve shape distortion and loss of detail.

A 1 MHz test frequency specified in a C–V test procedure is considered high frequency. Testing at –1 MHz reduces errors because of cabling or device series resistance. The display and bias readings are adjusted to compensate for transmission line error.

The filter circuit in a 1-pole analog, at –37 Hz, filters both capacitance and conductance signals. Digital plotter output controls are available through IEEE-488 bias for real-time plotting of all measurements as well as results of mathematical computations.

3.14 OSCILLATORS

Signal sources are described by several names:

1. Oscillators,
2. Test oscillators, and
3. Signal generators.

These names differ depending upon the design and application. An oscillator is the basic element common to all the sources. A test oscillator implies a calibrated attenuator and an output monitor. A signal generator is usually reserved for oscillators with modulation capability.

3.14.1 Considerations in Choosing an Oscillator

1. *Frequency range*: It should be capable of producing a wide range of frequencies, from a low frequency to a very high frequency, which is from 1 Hz to 30 MHz or even higher.
2. *Output voltage*: Whether sufficient range is available.
3. *Dial resolution and accuracy*: When the dial is set to a particular value, the same frequency is obtained.
4. *Frequency stability*: It is the ability to maintain the selected frequency over a period of time. Component aging and power-supply fluctuation affect stability.
5. *Amplitude stability*: Amplitude is to remain constant when '*f*' is changed.
6. *Distortion*: It is undesirable.

Example 3.1

The Wien bridge circuit is altered by adding an inductor L in series with an RC combination and by replacing the parallel RC combination by a resistor R_P. Calculate (a) the frequency of the oscillator of this circuit, (b) the minimum gain of the amplifier for a finite value of R_p for an oscillator. E_2 and E_1 should be in phase (Figs. 3.21–3.23).

Solution

$$E_2 = E_a - E_b\beta = \frac{E_2}{E_1}$$

$$E_b = \left(\frac{Re}{R_n + R_k}\right) E_1$$

If

$$E_2 = 0, \quad E_b = \left(\frac{1}{3}\right) E_1$$

Since E_2 should not be zero for the oscillator, $\dfrac{E_b}{E_1} = \dfrac{1}{3}$.

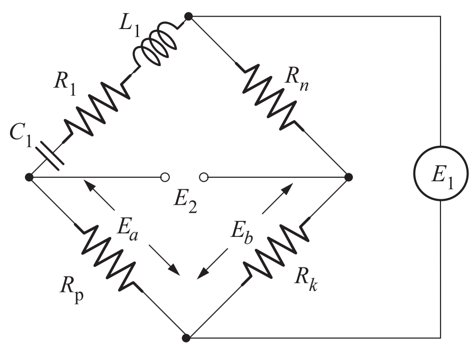


Figure 3.21 For Example 3.1

The loop phase shift must be zero and the bridge is balanced so that,

$$\omega L = \frac{1}{\omega C}$$

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

$$\beta A \geq 1$$

$$E_a = \left(\frac{R_p}{R + R_p}\right) E_1$$

$$E_2 = E_a - E_b\alpha - \beta = \frac{E_2}{E_1}$$

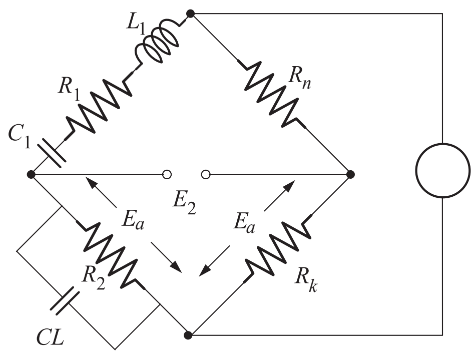


Figure 3.22 For Example 3.1

E_1 is the output of the voltage amplifier and E_2 the input of the voltage amplifier (feedback quantity):

$$E_2 = \left(\frac{R_k}{R_n + R_k} - \frac{R_p}{R + R_p}\right) E_1$$

$$\therefore \qquad -\beta = \left(\frac{R_k}{R_n + R_k} - \frac{R_p}{R + R_p}\right) - \beta A \geq 1$$

$$\therefore \qquad A \geq \frac{(R_n + R_k)(R + R_p)}{R R_k - R_n R_p}$$

In a Wien bridge, E_1 and E_2, the phase shift between them is zero only for resistors. Therefore, R_p and R_n are resistors.

$$\therefore \qquad t_1 = R - \frac{j}{wC} = (1 - j)\, R; \qquad Z_L = \frac{1}{\frac{1}{R} + jwC} = (1 - j)\,\frac{R}{2}$$

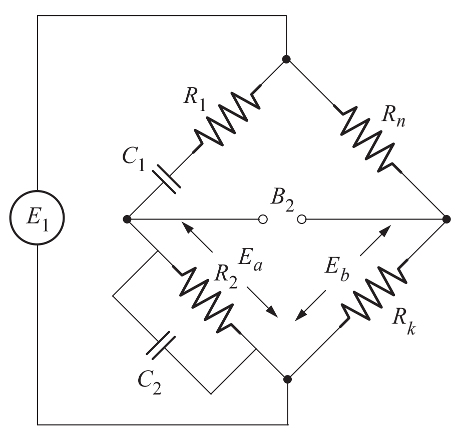


Figure 3.23 For Example 3.1

Thus, by substituting, the resonant frequency is

$$f = \frac{1}{2\pi RC}$$

∴ The voltage E_a across Z_2 equals

$$E_a = \left(\frac{Z_2}{Z_1 + Z_2}\right), \quad E_1 = \frac{1}{3} E_1$$

$$E_b = \left(\frac{R_k}{R_n + R_k}\right) E_1$$

If a null is desired,

$$E_2 = E_a - E_b = 0; \quad \text{i.e.,} \quad \frac{R_k}{R_n + R_k} = \frac{1}{3}$$

Then β will be zero. Therefore, we do not want β to be zero.

$$\frac{E_b}{E_1} = \frac{R_k}{R_n + R_k} = \frac{1}{3} - \frac{1}{\delta}$$

where δ is larger than 3.

Example 3.2

Determine the dynamic range of a spectrum analyser with a third-order intercept point of +300 dbm and a noise level of –90 dbm.

Solution

The expression for the dynamic range of a spectrum analyser is $\frac{2}{3}$ (Ip–MDS), where Ip is the power level of the third-order product (dbm) and MDS is the minimum detectable signal.

Substituting the values,

$$\begin{aligned} \text{Dynamic range} &= \frac{2}{3}(30 - (-90)) \\ &= \frac{2}{3}(30 + 90) = \frac{2}{3}(R_0) \\ &= 80 \text{ db} \end{aligned}$$

Example 3.3

Determine the minimum detectable signal of a spectrum analyser with a noise figure of 30 dB using a 1 kHz 3-db filter.

Solution

The noise level of the spectrum analyser is related to the noise figure and the IF bandwidth by the following equation:

$$\text{Minimum detectable signal (MDS)} = -114 \text{ dBm} + 10 \log\left(\frac{BW}{1 \text{ MHz}}\right) + \text{NF}$$

where BW = 3 dB bandwidth in MHz of the IF filter

NF = Noise figure in decibels

$$\text{MDS} = -114 \text{ dBm} + 10 \log\left(\frac{1\text{kHz}}{1\text{MHz}}\right) + 30$$

$$\text{MDS} = -134 \text{ dBm}$$

3.15 SUMMARY

To evaluate the quality of distortion in a waveform, stability of the output instruments is required. Harmonic distortion analyser, wave analysers, and spectrum analysers are such instruments. There are two types of wave analysers:

1. Audio frequency (AF) and
2. Radio frequency (RF) or high-frequency types.

Wave analysers are used in the reduction of sound generated by machines. However, distortion analysers measure the *RMS* value of harmonics. There are three types of HD analysers, viz.:

(i) Tunable reactive circuit type.
(ii) Heterodyne type.
(iii) Distortion measuring component type.

The heterodyne wave analyser is also called a wavemeter. Spectrum analysers are classified as LF (0–30 Hz) and HF (100 MHz to 3 GHz). High-frequency spectrum analysers are further classified as AF (upto 20 kHz) and RF (10 kHz to 3 GHz). Frequency resolution, bandwidth, sweep desensitisation, and sensitivity are the characteristic parameters of spectrum analysers. Spectrum analysers are used in the measurement of attenuation and to view the result of RF energy in radio applications; with the help of spectrum analysers, we can study bandwidth, spurious signal generation, effects of various types of modulation, etc.

Points to Remember

- The operating frequency range of an AF wave analyser is typically 10 kHz to 18 MHz.
- Wave analysers are used in industrial applications for the reduction of sound and vibrations generated by machines.
- The noise level emitted by the machines can be measured.
- A simplified block schematic of a wave analyser consists of a signal amplifier, a filter, an output amplifier, and a meter.
- Heterodyne means mixing of two signals. In super-heterodyne receiver input *AM* and *RF* are combined with a locally generated radio frequency signal to produce a signal at a new carrier frequency (intermediate frequency).
- There are three types of harmonic distortion measuring instruments, namely
  (a) Tunable selective type.
  (b) Heterodyne type.
  (c) Distortion-measuring components.
- Total harmonic distortion or distortion factor (B) is calculated as

$$B = \sqrt{D_2^2 + D_2^3 + D_2^4} + \ldots$$

where $D_2 = \frac{B_2}{B_1}$; $D_3 = \frac{B_2}{B_1}, \ldots$

B_1 = Amplitude of the fundamental.
B_2 = Amplitude of the second harmonic.

- Wave analysers are used in
  1. Sound and vibration analysis.
  2. Harmonic distortion analysis.
- Heterodyne harmonic distortion analyser is also called a wavemeter.
- Spectrum analysers are broadly classified into LF and HF.
- HF spectrum analysers are further classified into AF and RF.
- Test oscillator instruments will have a calibrated attenuator and an output monitor.

Objective-type Questions

1. The quality of waveform generated can be measured in terms of ________.
2. The instruments used to estimate the quality of waveforms are ________.
3. The attenuation provided in the wave analysers typically is about ________.
4. The attenuation at one half and twice the selected frequency in wave analysers is about ________.
5. The operating frequency of a wave analyser is typically ________.
6. The noise level emitted by a machine can be measured using ________.
7. In a super-heterodyne receiver the input AM and RF carriers are combined with a ________.
8. The disadvantage of a tunable selective harmonic distortion analyser instrument at low frequencies is ________.
9. The disadvantage of a heterodyne wave analyser or a wavemeter is ________.
10. Total harmonic distortion factor, D, is given by the expression ________.
11. The frequency range of an LF spectrum analyser is ________.
12. The frequency resolution of a spectrum analyser is its capacity to ________.

Review Questions

1. Draw the block schematic of AF wave analysers and explain the principle of working.
2. Draw the block schematic of an RF wave analyser and explain its working. What are the applications of this instrument?
3. Draw the block schematic of a tunable selective circuit-type harmonic distortion analyser and explain its working.
4. Give the schematic of a hetrodyne-type harmonic distortion analyser and explain its working.
5. Draw the block diagram of a distortion measuring component-type meter and explain its principle of working.
6. Draw the schematic of an AF spectrum analyser and explain its working.
7. Give the block diagram of an RF spectrum analyser and explain its operation.
8. What are the application of spectrum analysers? Explain.

Unsolved Problems

3.1 What is the dynamic range of a spectrum analyser if the noise level of the display is –90 dBm and two –20 dBm? Signal produces third-order intermodulation products.

3.2 Determine the dynamic range of a spectrum analyser with a third-order intercept point of +40 dBm and a noise level of –100 dBm.

3.3 Determine the resolution of a spectrum analyser using an IF filter within a 3-dB bandwidth of 20 kHz.

3.4 What is the minimum detectable signal of a spectrum analyser with a NF of 25 dB using a 1 kHz 3-dB filter?

3.5 Estimate the value of a minimum detectable signal (MDS) of a spectrum analyser with a NF of 40 dB, using a 1 kHz 3-dB filter.

4

Oscilloscopes

Introduction • Cathode ray oscilloscope • Block diagram of a CRO • Cathode ray tube (CRT) • Graticules • Electrostatic deflection sensitivity • Different controls in a CRO • Time base generators • Triggered mode • Neon time base circuit • Time base circuit for a general-purpose CRO • Lissajous figures • Types of CRO probes • High-frequency CRO considerations • Delay lines in CROs • Applications of CRO • Summary

4.1 INTRODUCTION

Oscilloscopes are also used very widely in electronic measurements and instrumentation systems. Unlike other measuring instruments, the signal being measured can be visually seen on the screen. The characteristics of the signals, like amplitude, frequency, phase, time period, duty cycle, etc. can be measured using oscilloscopes. The amplitude of the signals can vary from μV to even a few hundred volts. The frequency can range from very low (even DC) to MHzs. Because of the phosphorescence effect, the electrical signal is converted to a visible form and the shape of the signal can be seen.

In this chapter, the constructional blocks of CRO, various controls associated with CRO instruments, the principle of working, triggered sweep and dual beam CRO, and measurement of amplitude and frequency using CRO are explained. The student is expected to understand all these aspects connected with CRO.

4.2 CATHODE RAY OSCILLOSCOPE

The Cathode Ray Oscilloscope (CRO) is a very useful, general-purpose electronic instrument for testing and developing electronic circuits, systems, and instruments. Using a CRO, the shape, amplitude, and frequency of AC signals can be measured. CRO is also useful in determining the amplitude of DC signals. The versatility of CRO lies in the fact that time variation of the given electronic signal or electrical input can be studied. A part of the input signal can be expanded in the time scale, and the distortion, rise time, fall time, or any other characteristic feature of the signal can be studied. The electron beam generated in the CRO is deflected by the given electrical signal and when the deflected electron beam strikes the screen of the CRO, because of the phosphorescence effect, a visible trace is produced exactly in the same shape as the given electrical signal. Similar to the conventional graph that is plotted, the deflection of the electron beam in the *X*-direction on the screen is a measure of time, and the deflection in the conventional *Y*-direction is a measure of amplitude of the signal. The other types of recording devices such as an X–Y recorder and a strip-chart recorder are mechanical devices that are slow in response. CRO is capable of displaying events that take place over periods of microseconds and nanoseconds. Though a hard copy of the waveform seen on the screen is not obtained from CRO directly, the waveform can be photographed or stored as in the case of storage CRO. By incorporating microprocessor-based computing circuitry, CROs can directly display the characteristics of the signal on the screen such as amplitude, frequency, phase, rise time, and fall time on the screen itself along with the signal, without measurement and computation by the operator.

4.3 BLOCK DIAGRAM OF A CRO

The heart of a CRO is the cathode ray tube (CRT). Here, the electron beam is generated, accelerated, and deflected in accordance with the input signal, and a visible trace is produced on the phosphor screen. For the illumination on the screen to be bright, the velocity of the electron beam impinging on the screen and the kinetic energy (KE) must be high. Therefore, the beam must be accelerated by a high potential. Hence, a power supply circuit to generate the required high voltage also forms another part of the CRO. If the input signal amplitude is very small, it needs to be amplified. Hence, amplifier circuits also form a part of the system. In order to get a true time variation of the input signal, the electron beam must also be deflected along the X-axis linearly. Hence, a saw-tooth waveform or time base generator is to be incorporated. In addition, there will be delay lines, trigger circuits, and synchronisation circuits also in a CRO in addition to the all-important CRT. The block schematic of a CRO is shown in Fig. 4.1.

Vertical amplifier: This is also called a Y-amplifier. The electron beam deflection in the Y-direction or the vertical direction is proportional to the signal amplitude given to the Y-input or vertical plates. Hence, this is called a Y-amplifier.

The gain can be varied externally with the help of amplitude control. The bandwidth of this amplifier puts a limit to the maximum frequency of the input signal that can be measured using the CRO. If the magnitude of the external signal is large, it can also be attenuated in the potential divider attenuator section.

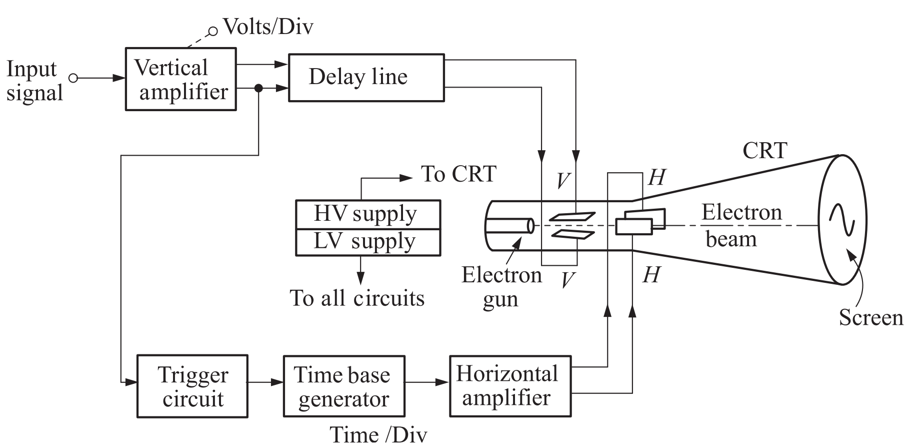


Figure 4.1 Block schematic of a CRO

Delay line: The electron beam is deflected in the *X*-direction by a saw-tooth waveform called the *time base signal*. When the set magnitude of the saw-tooth waveform is reached, it is to be reset and started again in order to get a true time variation of the given signal. The attenuators, amplifiers, pulse shapers, and circuit wiring introduce a certain amount of time delay. To allow the operator to observe the leading edge of the signal waveform, the signal drive for the vertical CRT plates must be delayed by at least the same amount. This is achieved by the vertical delay lines.

HV and LV supplies: To accelerate, deflect, and sweep the electron beam, a large voltage in kilovolts is required. This is generated by the high-voltage power supply circuits. *Vcc* and other low voltages required are generated in the low-voltage power supply circuits.

CRT: A CRT is the heart of a CRO. Here the electron beam is generated, accelerated, deflected, post-accelerated, and made to strike the fluorescent screen to give the visual display of the electrical input signal given to the vertical or the *Y*-plates.

Trigger circuit: To get a true representation of the input signal, the time base signal or the *X*-input and the *Y*-signal must be initiated at the same time. The time base signal must be initiated by the vertical signal itself for proper triggering. This is achieved by the trigger circuit.

Time base generator: This is a saw-tooth waveform generator circuit used to deflect the electron beam linearly in *X*-direction. Usually, a constant current, Miller sweep circuit is employed to generate the saw-tooth waveform.

Horizontal amplifier: The purpose of this circuit is to amplify an externally applied signal to the horizontal or *X*-plates. This also helps in adjusting the magnitude of the internal saw-tooth waveform being generated. If the internal time base waveform is not being used, an electron beam can be deflected horizontally by means of an external signal, which is applied to this circuit. This externally applied *X*-signal can be amplified by adjusting the gain.

4.4 CATHODE RAY TUBE (CRT)

A cross-sectional view of a CRT is shown in Fig. 4.2, and a simplified line sketch is shown in Fig. 4.3.

A CRT can be broadly divided into five sections:

1. Beam-generating area.
2. Beam focussing, pre-acceleration.

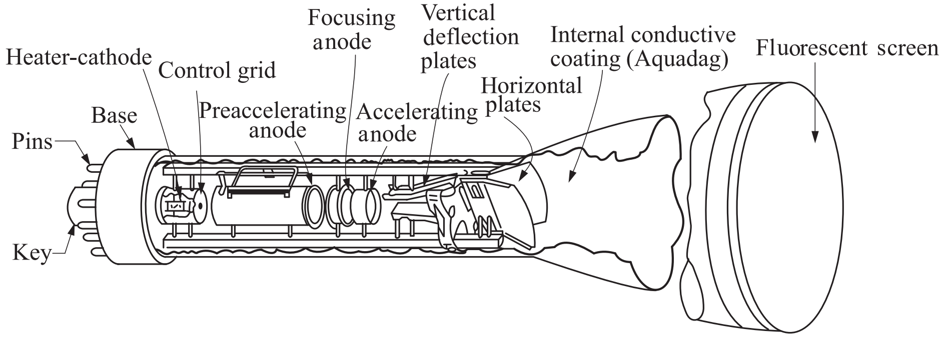


Figure 4.2 Internal structure of a CRT

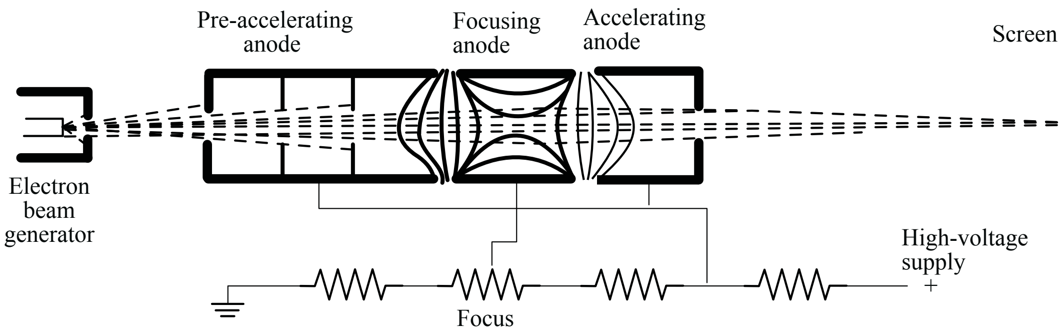


Figure 4.3 Line sketch of a CRT

3. Post-acceleration.
4. Deflecting plates.
5. Phosphor screen.

An electron beam is generated by thermionic emission. The construction is similar to that of a vacuum triode. A thermally heated oxide-coated cathode emits electrons. The intensity of the electrons can be controlled by the negative potential of the control grid. *Intensity control in CROs is also done this way*. The thermionically emitted electrons are attracted by the positive potential of the anode and emerge out of a narrow slit. The pre-accelerating anode is a hollow cylinder that is at a potential, a few hundred volts more positive than the cathodes. The electron beam will be accelerated in the electric field.

A focussing anode is mounted just ahead of the pre-accelerating anode. Its shape is in the form of a *cylinder*. Actually there will be three sections of the focussing anodes to perform electron beam focussing. An electrostatic focussing lens is formed by the electric lines of fields because of the two electrodes at different potentials. The electron lens requires three elements, with the centre element at a lower potential than the two outer elements. This is shown in Fig. 4.4.

The electron beam after emerging from the cathode and being focussed by the electrostatic focussing lens into a fine narrow beam should travel and reach the phosphor screen to give a visible display.

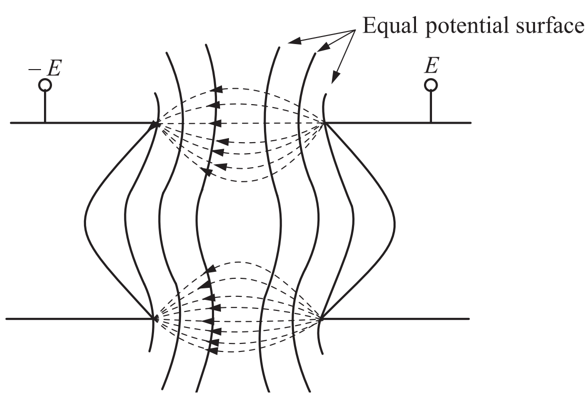


Figure 4.4 Electrostatic focussing system

Therefore, the energy of these electrons must be sufficiently high. In addition, the intensity of the beam depends on the velocity with which they strike the phosphor screen and the KE. Therefore, the electron beam must be accelerated further. This is done in the post-acceleration area. This section will have an accelerating anode at 10 kV or higher. Usually, post-acceleration is done after deflection, because deflecting a beam with high accelerating energy needs large deflecting voltages. Therefore, in practice, after the deflection plates, post-acceleration is done.

In the beam-deflecting area, the *Y-plates* or vertical plates cause deflection in the *Y* or the vertical direction. These plates are *mounted horizontally* to the plane of the CRO. *X-deflecting plates* or horizontal-deflecting plates cause the beam to deflect in the *X*-direction. These plates are *mounted vertically*. The time base signal or saw-tooth waveform is applied to these plates when the external signal applied to the *Y*-plates is being triggered internally.

When the electron beam strikes the screen, a visible trace is produced because of *phosphorescence effect*. The screen material on the inner surface of the CRT that produces this effect is called *phosphor*. This phosphor absorbs the KE of the bombarding electrons and re-emits energy at a lower frequency in the visible spectrum. The property of some crystalline materials that emit light when stimulated by radiation is called *fluorescence*. Fluorescent materials have a second characteristic called *phosphorescence*. This is the property of the material to continue emitting light even after the source of energy is extinguished. The intensity of the light emitted from the CRT screen depends on the number of electrons striking the screen, velocity of the electrons, the time period the beam strikes a given area of the phosphor, and the characteristic of the phosphor materials. Phosphor is the commercial name used for the chemical substances exhibiting this characteristic.

A coating of graphite powder known as *Aquadag* coating is applied to the walls of the CRT, which absorb secondary electrons and provide return paths for the current flow.

4.5 GRATICULES

Graticules are the scale marking provided on the front of the CRT screen. These are vertical and horizontal lines in centimeters sub-divided into millimeters. Using these scales, the amplitude and frequency of the input signal are determined. There are two types of graticules.

1. External graticules.
2. Internal graticules.

External graticules are the scale markings fixed outside the CRT screen glass shell. If the spacing between the phosphor screen and graticules is large, there will be errors in measurements because of parallax. This is an old method, and recent CROs use only internal scales.

Internal graticules are scales engraved on the CRT screen glass envelope itself. Therefore, parallax errors will not be there. However, once the CRT is assembled, if there is a misalignment between deflecting plates and internal graticules, correction has to be made electronically. Physical realignment is not possible. The second difficulty with the internal graticules is that it is more difficult to illuminate the internal scales for photographic purposes. Therefore, special arrangement is to be made to illuminate these lines. Separate electron guns have to be used to illuminate the entire phosphor screen to increase the brightness of the graticules. A CRT line diagram with different electrode voltages and adjustments is shown in Fig. 4.5.

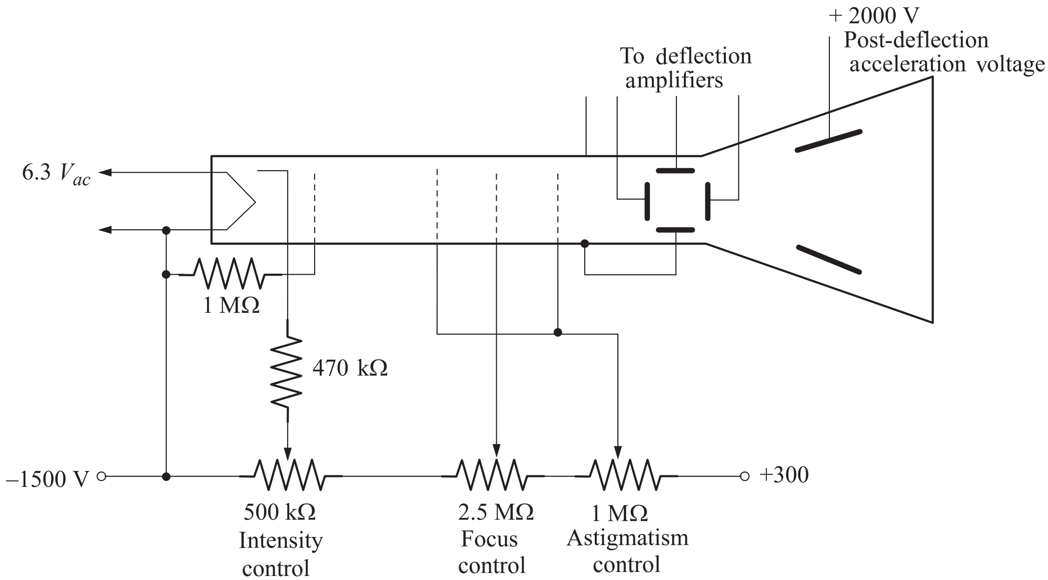


Figure 4.5 CRT line diagram with various controls

4.6 ELECTROSTATIC DEFLECTION SENSITIVITY

This is defined as the deflection produced on the CRT screen in millimeters or centimeters when the deflecting voltage is 1 V:

$$S_E = \frac{Y}{V_d}$$

where

Y = deflection produced on the screen in mm or cm

V_d = deflecting voltage

The expression for S_E can be derived as given below (Fig. 4.6).

Let

V = The velocity of the electron beam entering the deflecting plates,

V_d = The deflecting voltage applied to the deflecting plates,

l = The length of the deflecting plates,

$L = \left(\frac{l}{2} + l'\right)$ the distance between the screen and the centre of the plates,

d = The distance between the plates,

y = The deflection produced on the screen,

l' = The distance between the outer edge of the deflecting plates and the CRT screen, and

V_a = The accelerating anode voltage.

The electron enters the deflecting plates with a velocity $v = \sqrt{\dfrac{2eVa}{m}}$.

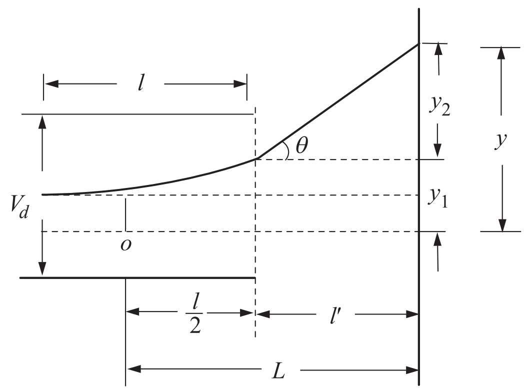


Figure 4.6 Trajectory of the electron upto the screen

$$f = e\varepsilon = ma$$

$$\varepsilon = \frac{V_d}{d}$$

Acceleration in the Y-direction,

$$a_y = \frac{e\varepsilon}{m}$$

$$= \frac{eV_d}{dm}$$

$$v_y = \text{velocity in the } Y\text{-direction} = ay\ t$$

$$t = \frac{l}{v}$$

$$v_y = \left(\frac{eV_d}{dm}\right)\left(\frac{l}{v}\right)$$

$$\frac{v_y}{v} = \left(\frac{eV_d}{dm}\right)\left(\frac{l}{v^2}\right)$$

$$\tan\phi = \frac{y_2}{l'} = \frac{v_y}{v}$$

$$\frac{y^2}{l'} = \left(\frac{eV_d}{dm}\right)\frac{l}{v^2}$$

$$\text{or } y_2 = \left(\frac{eV_d}{d\,m}\right)\left(\frac{l}{v^2}\right) l'$$

$$y = ?$$

$$s = ut + \frac{1}{2}at^2; \qquad u = 0 \quad \because v_y = 0$$

$$t = \frac{l}{v}$$

$$y_1 = \frac{1}{2}at^2$$

$$a = \frac{eV_d}{dm}$$

$$y_1 = \left(\frac{1}{2}\right)\left(\frac{eV_d}{dm}\right)\left(\frac{l^2}{v^2}\right)$$

$$y = y_1 + y_2$$

$$\left(\frac{1}{2}\right)\left(\frac{eV_d}{dm}\right)\left(\frac{l^2}{v^2}\right) + \left(\frac{eV_d}{dm}\right)\left(\frac{l}{v^2}\right)l^1 = \left(\frac{eV_d}{dm}\right)\left(\frac{l}{v^2}\right)\left(\frac{l}{2}+l'\right)$$

$$\left(\frac{l}{2}+l'\right) = L$$

$$\therefore y = \left(\frac{eV_d}{dm}\right)\left(\frac{l}{v^2}\right)L$$

Deflection sensitivity

$$S_E = \frac{y}{V_d} = \frac{e}{dm}\frac{l}{v^2}L$$

$$v^2 = 2\left(\frac{eV_d}{dm}\right)$$

$$\therefore y = \left(\frac{\cancel{e}V_a}{d\cancel{m}}\right)\left(\frac{lL\cancel{m}}{2\cancel{e}V_d}\right)$$

$$\therefore y = \left(\frac{lL}{2d}, \frac{V_a}{V_d}\right)$$

$$\text{or deflection sensitivity } S_E = \frac{y}{V_d} = \frac{lL}{2dV_a}$$

$$\therefore \boxed{S_E = \frac{lL}{2dV_a}}$$

4.6.1 Design Criteria

A CRT with a high deflection sensitivity can be designed by having large l and L, small d and V_d. However, each parameter has a limiting value. Therefore, optimum values are to be chosen. If L is large, S_E can be large, but the size of the CRO becomes very large. If l is large, the electron beam after deflection may hit the deflecting plates and may not be able to come out. If d is made very small to increase S_E, the electron beam after deflection in the Y-direction will hit the plates. If V_d is made small to increase S_E, the deflection produced will be less. Therefore, optimum values of these parameters must be chosen in designing a CRT with a reasonable value of the deflection sensitivity.

4.7 DIFFERENT CONTROLS IN A CRO

AC	When AC signals are given as *Y*-input, the CRO is kept in this mode. DC is blocked.
DC	When DC signals are given as *Y*-input, the CRO is kept in this mode. Both AC and DC are seen.
GND	Zero *V* or ground potential is applied to the *Y*-plates. Therefore, no deflection is seen on the CRO screen. The electron beam will be at the zero axis, along the *x*-axis.
Intensity	The brightness of the electron beam can be varied.
X-shift	The signal can be moved or shifted in the *X*-direction on the screen.
Y-shift	The signal can be moved or shifted in the *Y*-direction on the screen.
Focus	Electron beam focussing can be adjusted.
Trigger	Auto mode: The time base signal is initiated by the *Y*-input in the auto mode. Normal mode: In the normal mode, this is adjusted manually. The sweep signal is not in a free running mode.
Slope ±	The sweep signal (time base signal) can be triggered by the positive-going or negative-going portion of the triggered signal.
Level	DC offset voltage adjustment.
Time base	ms/div, Ms/div: The frequency of the internal saw-tooth waveform (time base) can be varied.
Y-input	mV/cm, V/cm: The gain of the *Y*-amplifier can be varied.
INT/EXT	In the internal mode, time base signal, i.e., internally generated saw-tooth waveform, is applied to the *X*-plates. In the external mode, the internal time base signal is cut off. An external signal must be given to sweep the electron beam in the *X*-direction.
Y-input vertical input	The signal to be observed on the screen is applied to the vertical deflecting plates.
X-input horizontal	Saw-tooth waveform is applied in the INT mode to deflect the beam in the *X*-input direction.
CAL	Calibration can be done with the standard signal generated internally, which can be applied to *Y*-plates.

4.7.1 How to Operate a CRO

1. Switch ON the power.
2. Keep the selector switch AC/DC/GND in the ground (GND) position.
3. Keep the time base selector switch in the INT mode (internal mode).
4. Adjust the intensity knob to be in a mid-position.
5. Keep the time base selector in the AUTO mode.
6. Adjust the *X*-shift if needed.
7. Adjust the *Y*-shift if needed.
8. Now keep the *Y*-input selector switch in the AC/DC position.
9. Give the input signal to be observed on a CRO to the *Y*-input or to the *A* or *B*-input knobs.

10. Adjust the trigger level to get a stable input.
11. Adjust the time base knobs (cm/sec) so that the frequency of the time base is varied to get a suitable number of cycles on the screen.
12. Adjust the vertical amplifier (V/sec) suitably to get the required amplification and to vary the magnitude of the input signal.

4.8 TIME BASE GENERATORS

CRO application involves measurement of a waveform that varies with time. Therefore, the electron beam spot should move across the screen with constant velocity. The deflection of the electron spot should increase linearly upto the other end of the screen and must retrace back to the starting point after reaching the edge of the screen or after the completion of one cycle. Since the deflection is proportional to the deflecting voltage, the voltage applied to the *X*-plates of the CRT should increase linearly and suddenly drop to zero. Hence, the time base waveform being given to the *X*-plates must be a saw-tooth waveform.

The horizontal amplifier accepts either a run-up or run-down ramp voltage and supplies both waveforms to its push pull/output circuit which in turn gives it to the horizontal deflection plates. The circuit that produces the saw-tooth waveform is called a *time base generator* or a *sweep generator*. The output of a sweep generator is called the *sweep voltage*. The waveforms are shown in Fig. 4.7.

T_s is called the sweep time. Typically, it is about 100 μsec. T_r is called the retrace time or fly-back time. Its typical value is about 1 μsec. This type of waveform can be generated by a capacitor getting charged through a resistance *R*. Hence, the voltage across the capacitor increases exponentially. When the capacitor is shorted by a switch, it discharges quickly. Therefore, the waveform across the capacitor is approximately a saw-tooth waveform. The basic circuit can be as shown in Fig. 4.8.

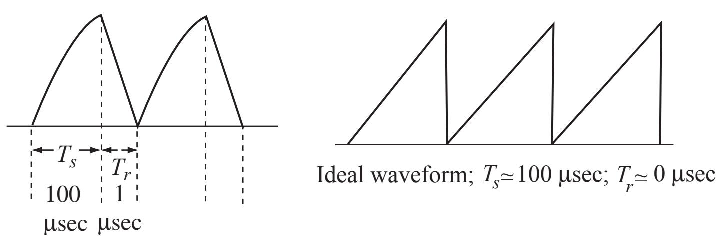


Figure 4.7 Time base waveforms

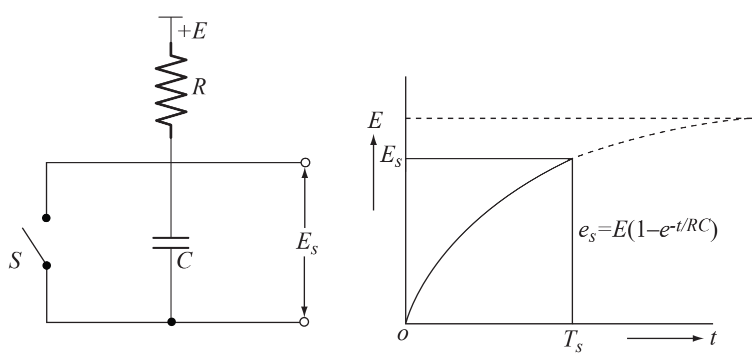


Figure 4.8 Time base RC circuit

4.8.1 Time Base Circuits

To deflect a beam from left to right on the screen,

1. The right-hand horizontal deflection plate is supplied with a free-running ramp voltage.
2. The left-hand horizontal deflection plate is grounded.

A typical saw-tooth waveform is as shown in Fig. 4.7. T_s is called the *sweep time*, T_r the *retrace* or *fly back time*.

A basic RC circuit that is used to obtain a saw-tooth waveform is as shown in Fig. 4.8. Switch S is initially closed and the voltage across C is O. At $t = 0$, the switch is opened and the voltage across the capacitor is $e_S = t\ (1-e^{-t/RC})$. If the capacitor charging process is stopped early in its exponential rise, the capacitor voltage rise is slight and approaches a linear rise in voltage. The slope error is about 10% when the supply voltage E is 200 V. The switch performing this function is a vacuum tube or gas tube or thyratron tube or SCR, or a transistor.

A thyratron can replace the switch. A thyratron 'fires' when the anode reaches its breakdown potential corresponding to the critical grid voltage. At this point, the tube loses control over the plate because of variation in the grid voltage. Once the thyratron is fired, the drop across the tube remains constant, and the current through the tube is determined by the number of electrons (Fig. 4.9).

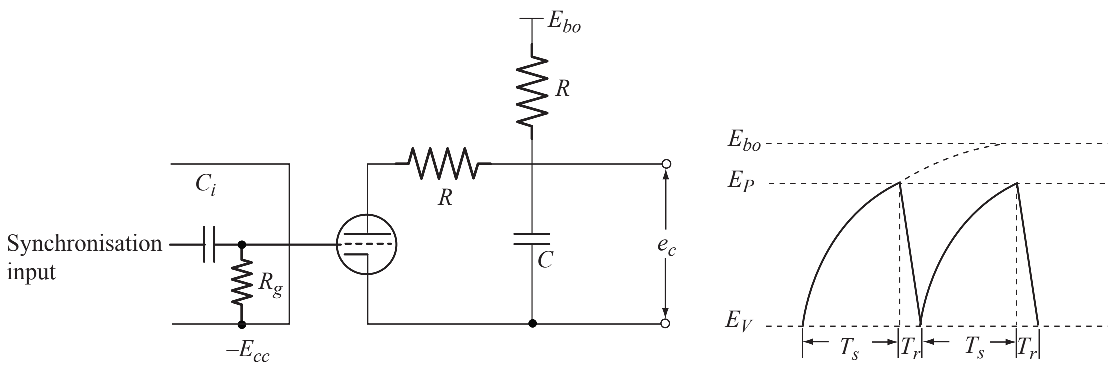


Figure 4.9 Time base and waveform circuits

Capacitor C charges exponentially through R approaching supply voltage E_{bb}. When the plate voltage reaches a value E_p, the thyratron ignites and conducts heavily. Capacitor C will discharge rapidly through the tube and series resistor R. The tube will turn off by itself when the discharge current is less than the minimum required for ignition and conduction ceases. At the inlet and the outlet, the tube presents resistance, and the capacitor starts discharging again through resistor R. The process repeats itself and a saw-tooth is obtained.

The maximum frequency to which thyratron can be used is 100 kHz.

The deionisation potential of a thyratron is a function of the nature of gas and the gas can be Argon. The pressure is about 1 bar or 1 mm of Hg. The waveform for time base is as shown in Fig. 4.10(a). It is a saw-tooth waveform.

$$T_s = \text{Sweep time}$$
$$T_r = \text{Retrace time}$$

T_r must be as small as possible. The sweep waveform, though exponential in nature, can be approximated to be linear, as in Fig. 4.10(b).

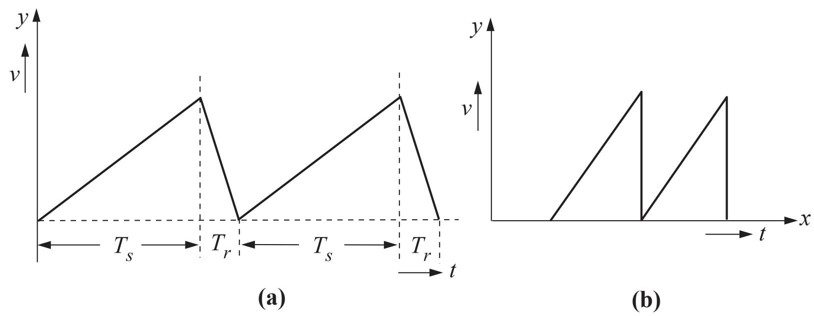


Figure 4.10 Sweep waveforms: (a) practical waveform and (b) ideal waveform

The circuit diagram for generating saw-tooth waveform for a CRO time base is as shown in Fig. 4.11(a). Due to capacitor charging and discharging, a saw-tooth waveform will be generated (Fig. 4.11(b)).

The voltage across the capacitor increases, following the equation

$$V_s = V(1 - e^{-t/RC})$$

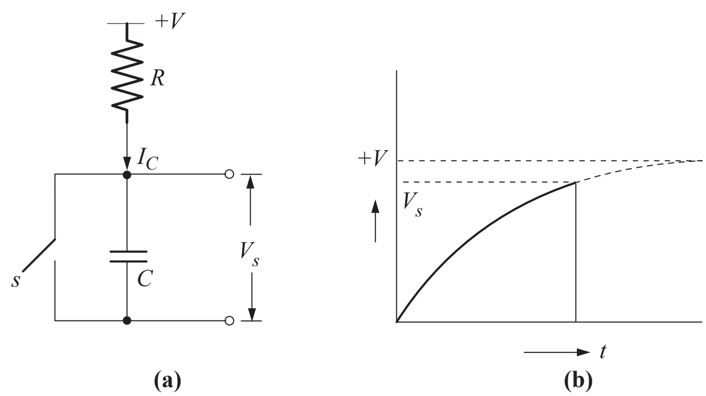


Figure 4.11 (a) Basic circuit schematic and (b) sweep waveform

This is not exactly a linear equation as such. Therefore, there will be a slope error to the extent of about 10%. The circuit must be designed so that this error is as minimum as possible. In addition, the charging current I_c of the capacitor will not be constant as the voltage across the capacitor increases. Hence, non-linearity creeps in. The device functioning as the switch S, which is to remain open when the capacitor is charging and is to be closed after the voltage across the capacitor reaches a preset value V_s, can be an electronic device such as a thyristor, neon tube, thyratron, unijunction transistor (UJT), or a bipolar junction transistor. In order to ensure that the charging current remains almost constant and that the slope error is minimum, different types of circuits are developed. Some of the electronic circuits that generate saw-tooth waveforms and that can be used for time base circuits in CROs are as follows:

1. Thyratron RC time base circuit.
2. Neon RC time base circuit.
3. Miller current sweep circuit (Miller integrator).

4. 555 timer circuit, taking the output voltage across the capacitor *C*.
5. A stable multivibrator.
6. Bootstrap sweep circuit.
7. UJT relaxation oscillator circuit.

These circuits and their operations are explained in previous electronic courses. Hence, it is not repeated here. **In practice, a constant current Miller integrator circuit is used for time base in CROs.**

4.9 TRIGGERED MODE

A waveform that is to be observed on the CRO may not be periodic but may perhaps occur at irregular intervals. In that case, it is desirable if the sweep is in operation and that the sweep signal (time base signal) is initiated by the waveform itself. Even if the waveform is periodic, it is only for a short duration, that is, a waveform may consist of a number of pulses of 1msec duration or less with a time interval of 100 msec between the pulses. If the CRO is in the free-running mode, then the pulses or the part of the signal carrying information may occupy a very small portion of the screen. Therefore, all the details of the pulse would be lost. If the sweep speed is 100 msec and if the time base is spread over 10 cm of the screen, then the pulse of 1 cm 'ON Time' and 100 cm 'OFF Time' would occupy 0.1 cm on the screen as shown in Fig. 4.12 (because 100 msec corresponds to 10 cm. 1 msec corresponds to 10/100 × 1 = 0.1 cm). Hence, all details of the pulse would be lost.

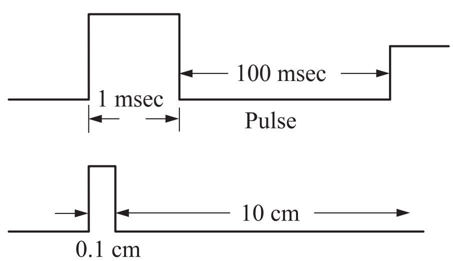


Figure 4.12 Pulse waveform as seen on CRO screen in the free-running mode

In order to spread the pulse over the entire CRO screen, a sweep period of less than 1.5 msec initiated by the pulse itself is required. A circuit that accomplishes this task is called a *triggered sweep circuit*. The required information in the input signal can be expected and details like rise time, fall time, distortion in the input signal can be analysed.

4.9.1 Free-Running Mode

If the saw-tooth waveform is repetitive, a new sweep is started immediately after the previous sweep is terminated. This mode of operation is called the **free-running**, **stable**, or **recurrent mode**. When a periodic signal of frequency f_v is applied to the vertical input terminals of the CRO, while a sweep frequency f_s is applied to the horizontal axis, a stationary pattern of n cycles will appear on the screen if f_v is an integral even multiple of $f_s = nf_s$, where n is an integral even multiple. In Fig. 4.13, if $n = 2$, $f_v = 2f_s$. Therefore, two stationary cycles are observed on the screen. T_s is the sweep time and T_r is the retrace time. On the CRT screen, the beam retraces very rapidly from right to left if the retrace time T_r is very small, which is usually so. Because of this, the '*retrace path*' or the '*fly back path*' of the electron beam is not seen or is not visible on the CRO screen. To get a stationary pattern on the CRO, the sweep generator must be synchronised to the vertical signal. If it is not so, the waveform will drift across the screen. Synchronisation is achieved by applying the vertical deflection signal to the gate of the thyristor-switching device and thus controlling the point at which the device is triggered.

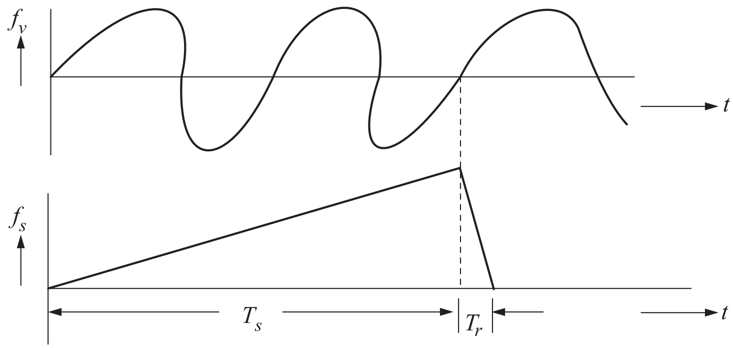


Figure 4.13 Free-running mode of operation

4.9.2 Synchronisation of the Sweep Circuit

The time base signal or the saw-tooth waveform is to be synchronised to an external signal to get the full details of the input signal on the screen. The external signal is applied to the synchronisation input or synchronisation input terminal of the time base circuit in such a way that the capacitor of the time base circuit discharges earlier. In order for synchronisation to occur, the synchronising pulse must occur at time intervals causing the charging process of capacitor '*C*' to stop prematurely. Therefore, $T_p < T_o$ as shown in Fig. 4.14. If $T_p > T_o$, synchronisation will not occur. It is not only the time of occurrence of the synch-pulse that is important but also the magnitude of the synch-pulse that is equally important. If the magnitude is very small, the switching device being used in the synchronisation circuit will not be able to change state. Hence, the capacitor '*C*' in the time base circuit will not discharge to initiate the next sweep. If the amplitude of the synch-pulse is very large, two sweeps will be initiated. The time base circuit for a general-purpose CRO is as shown in Fig. 4.15. For the convenience of readers, the block schematic of a CRO is again given in Fig. 4.16.

If an external signal is not being applied to the *X*-plates, the Time base is applied to the *X*-amplifier. Brightness is varied by varying the voltage applied to the accelerating anodes. Thereby, the velocity of the electron varies.

Focussing adjustment is done by varying the voltage applied to the focussing anodes. Therefore, the electron beam can be made to focus as a fine dot on to the screen.

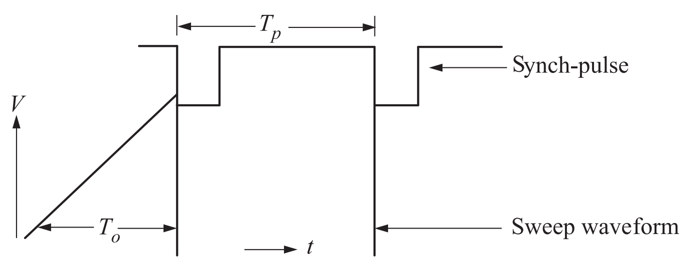


Figure 4.14 Synchronisation waveform if $T_p > T_o$ synchronisation will not occur

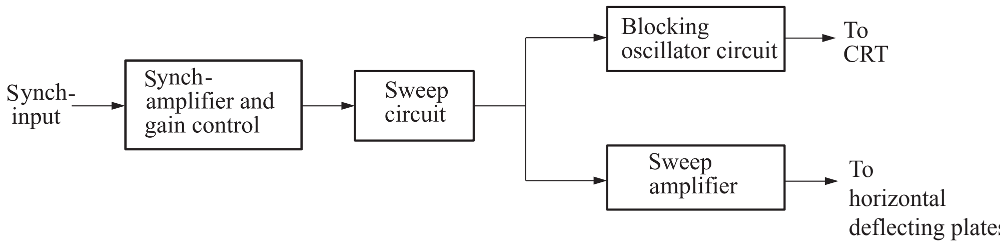


Figure 4.15 Time base circuit connection in a CRO

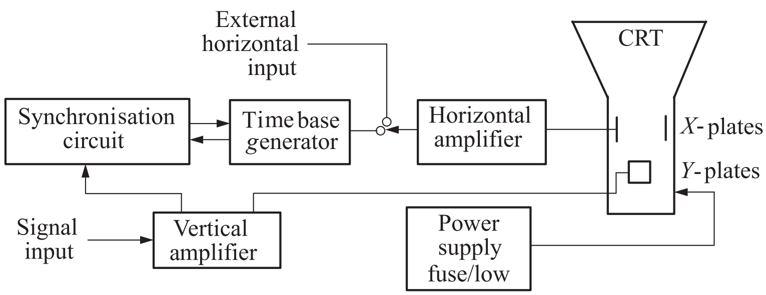


Figure 4.16 Block schematic of a CRO

4.9.3 Types of CROS

1. Single beam.
2. Double beam.
3. Dual race.
4. Storage.
5. Sampling CRO.

4.9.4 Sections of CRTs

The various sections of a CRT are as shown in Fig. 4.17:

(i) Beam-generating area.
(ii) Beam focus section.
(iii) Beam deflection region.
(iv) Post-acceleration.
(v) Target screen.

4.9.5 Deflection Sensitivity Equation

Electrostatic Deflection Sensitivity:

$$S_E = \frac{lL}{2dV_a}$$

l = Length of deflecting plates

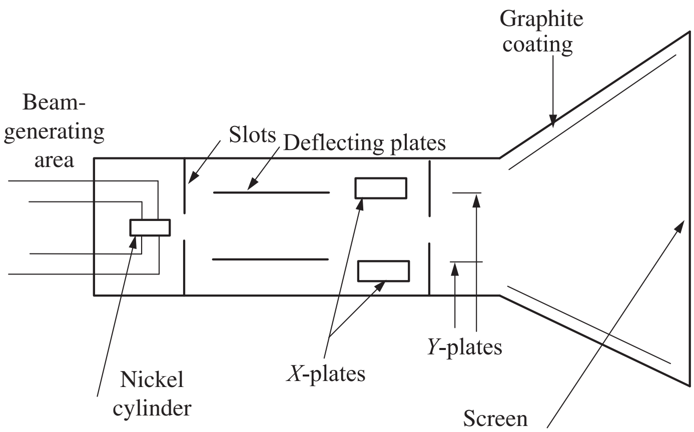


Figure 4.17 Cathode ray tube

L = Distance from the screen to the centre of the plates
d = Distance between the plates
S_E = Electrostatic deflection sensitivity
S_M = Electro-magnetic deflection sensitivity
V_a = Accelerating voltage

$$S_M = \sqrt{\frac{e}{m}}\,\frac{lL}{\sqrt{2V_a}}$$

4.10 NEON TIME BASE CIRCUIT

Neon tubes start conducting at a definite voltage called the *striking voltage*. Once the conduction has started, even if the voltage across the lamp is reduced slightly, conduction does not cease, since ions are still present in large numbers. If the voltage is continuously reduced, the conduction because of ionisation ceases. The voltage thus reached is called the *extinguishing voltage*. This property of the Neon tube has been utilised in 'Neon tube time base' circuits (Fig. 4.18).

In the initial state, the neon tube is not conducting. When E_{dc} is applied, current flows through high-resistance R and condenser C, charging the condenser. Voltage e across the condenser exponentially increases.

$$e' = E_{dc}\,(1 - e^{-f/RC})$$

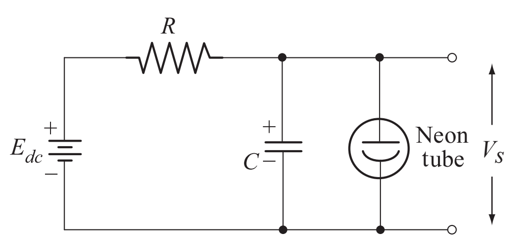


Figure 4.18 Neon tube time base circuit

Typical values

$$E_{dc} = 200 \text{ V}$$
$$V_s = 170 \text{ V}$$
$$R = 1 \text{ M}\Omega$$
$$V_e = 140 \text{ V}$$
$$C = 0.02 \ \mu\text{F}$$

The rate of increase in voltage depends upon the time constant RC. As soon as the striking voltage is reached, the neon tube starts conduction because its resistance '*r*' is very small. The condenser C then discharges through the neon tube and its voltage drops exponentially with time constant RC. The value of '*r*' is very small and hence the discharge is rapid. When the voltage across *C* reduces to the extinguishing voltage, the neon tube ceases to conduct and it acts as an open circuit. Charging of the condenser starts again and the cycle of operation is continued. So a saw-tooth waveform will be produced.

4.10.1 Frequency of Neon Time Base

$$V_s = E_{dc}\,(1 - e^{-t_s/RC}) \tag{4.1}$$

This is because when

$$t_s = \infty, \quad V_s = E_{oc}$$

At t_s the striking voltage is given by the above exponential equation.
At t_1 the extinguishing voltage V_e is given by,

$$V_e = E_{dc}\left(1 - e^{\frac{-t}{RC}}\right) \tag{4.2}$$

At t_1 during the exponential rise of voltage (Fig. 4.19),

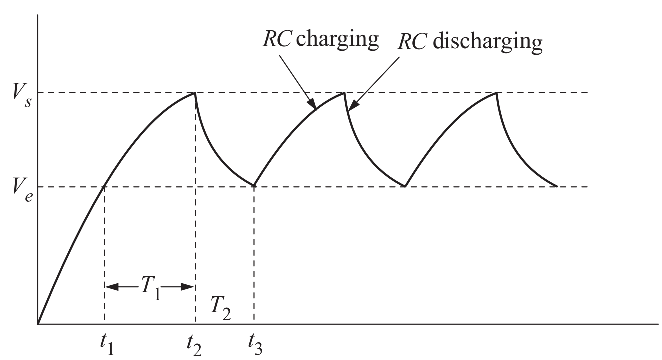


Figure 4.19 Time base waveforms

$$e^{-t_2/RC} = 1 - \frac{V_s}{E_{dc}} = \frac{E_{dc} - V_s}{E_{dc}} \tag{4.3}$$

$$= \frac{E_{dc} - V_c}{E_{dc}} \tag{4.4}$$

$$e^{(t_2 - t_1)/RC} = \frac{E_{dc} - V_c}{E_{dc} - V_s} \tag{4.5}$$

$$(t_2 - t_1) = T_1 \tag{4.6}$$

$$\frac{T_1}{CR} = \ln\left(\frac{E_{dc} - V_e}{E_{dc} - V_s}\right) \tag{4.7}$$

T_2 is very small compared to T_1.

Total time period: $T = T_1 + T_2 \simeq T_1$

$$T = CR \ln\left(\frac{E_{dc} - V_e}{E_{dc} - V_s}\right)$$

$$\therefore \qquad f = \frac{1}{T} = \frac{1}{CR \ln\left(\frac{E_{dc} - V_e}{E_{dc} - V_s}\right)}$$

For a given neon tube, V_e and V_s are fixed. Therefore, R, C, and E_{dc} are the only variables. To make the curve linear by increasing E_{dc} in spite of all care in the design, the sweep voltage curve departs markedly from the linear curve. Hence, it is not widely used.

4.10.2 Neon Lamp

It consists of two metal electrodes in a gas envelope—a low pressure of Neon gas. As the applied voltage is increased, the effective field strength between the electrodes increase. Because of the shape of the electrodes, the field is non-uniform. At a certain voltage, and the field strength is sufficient to ionise the neon gas atoms. Thus, a conducting pressure of neon ions, electron, and neon atoms occurs. The electrons excited from the cathode are given high energy as they are accelerated by the steep gradient in the cathode region.

The collusion of these electrons with Neon atoms produces highly excited atoms that lose some of their energy by emitting light. The Neon tube is used to generate saw-tooth waveforms.

4.10.3 Free-Running Mode of CRO

The frequency of an oscillator is a function of the supply voltage E_{bb}, critical grid voltage E_{cc}, and time constant $T = RC$. The frequency of the oscillator is varied by varying R and C. R is called a timing resistor and C is called a timing capacitor.

T_S = Sweep time

T_R = Retrace time

To get a stationary pattern on CRO, the sweep generation must be synchronised to the vertical signal. Otherwise the waveform will drift across the screen. Synchronisation is achieved by applying the vertical deflection signal to the grid of the thyratron, thus controlling the grid voltage and hence the tube's variation potential (Fig. 4.20).

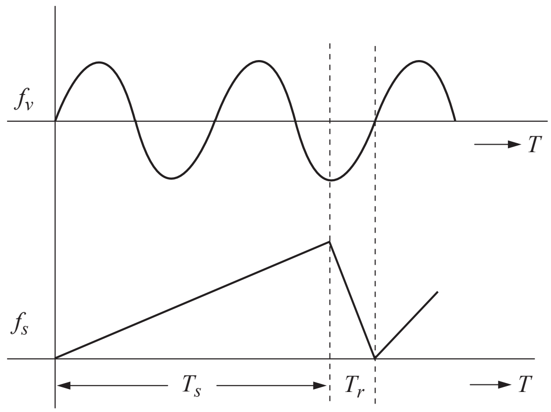


Figure 4.20 Sweep circuit

4.10.4 Using CRO in Triggered Mode

If the CRO is in the free-running mode, and if the waveform is synchronised to the sweep frequency, a stationary pattern would appear on the CRT screen (Figs. 4.21, 4.22).

Figure 4.23 shows a sweep circuit using a thyratron. The values of R_2 and R_3 are chosen such that the voltage at their junction is less than E_p, the thyratron ionisation potential. When the thyratron is not conducting, the timing capacitor C charges towards the thyratron-firing voltage, but a point is reached when the diode conducts, and prevents a further rise in capacitor voltage. Therefore, the capacitor voltage then reaches the thyratron-firing voltage and the capacitor cannot discharge. It is chopped at a voltage E_r as determined by R_2 and R_3. If a positive voltage is now applied to the grid of the thyratron, its ionisation potential is below the value of E_r and the tube fires. The capacitor discharges quickly through R_1 and again charges exponentially through R towards E_{bb} but is clamped when it reaches E_v.

The pulse train is used to trigger the sweep circuit. At the leading edge of the pulse, the thyratron is triggered and the capacitor discharges first, after which the charging cycle starts again. The sweep speed can be adjusted such that the entire pulse faces within the sweep time T_s. The sweep circuit responds to the trigger signal only after it has been applied, so that a small part of the initial waveform is lost because of time base retrace. Therefore, this is used for low-frequency waveforms. The triggering circuit can be improved by reducing the retrace time. Retrace time is determined by the ionisation time of the gas and it is of the order of 10 msec. A circuit using UJT has a smaller retrace time. Most modern high-frequency CROs use vacuum tubes or transistor switches instead of a thyratron or UJTs. By means of a switch, the trigger circuit can be changed to the free-running mode. When S is closed, R_2 is shorted and diode D is connected directly to E_{bb}. Under these circumstances, the diode D conducts since its cathode is at a higher potential than the ionisation voltage of the thyratron.

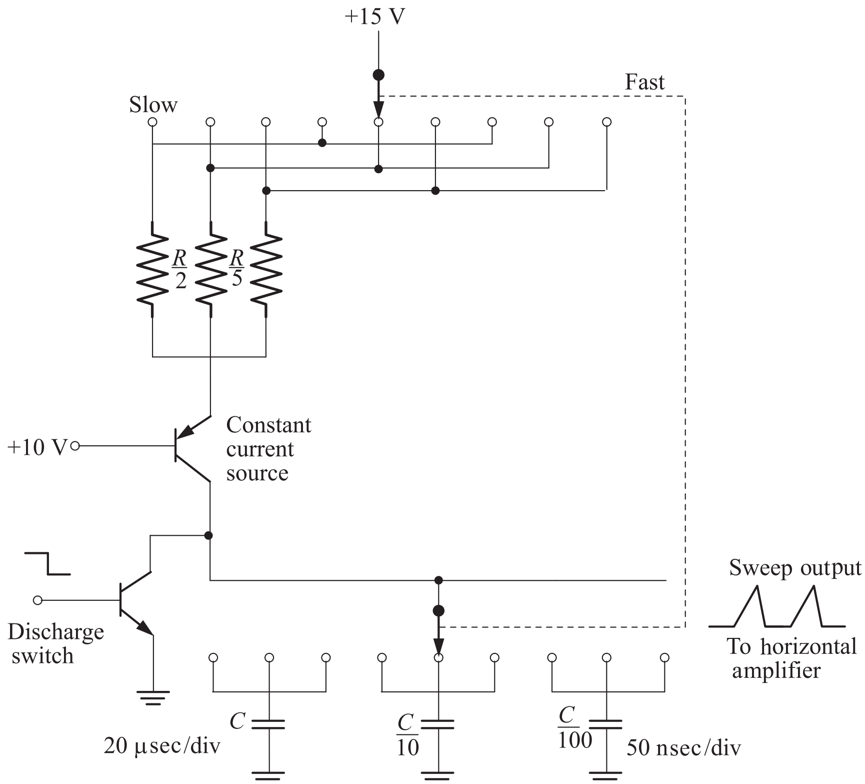


Figure 4.21 Simplified schematic of a triggered time base for an oscilloscope

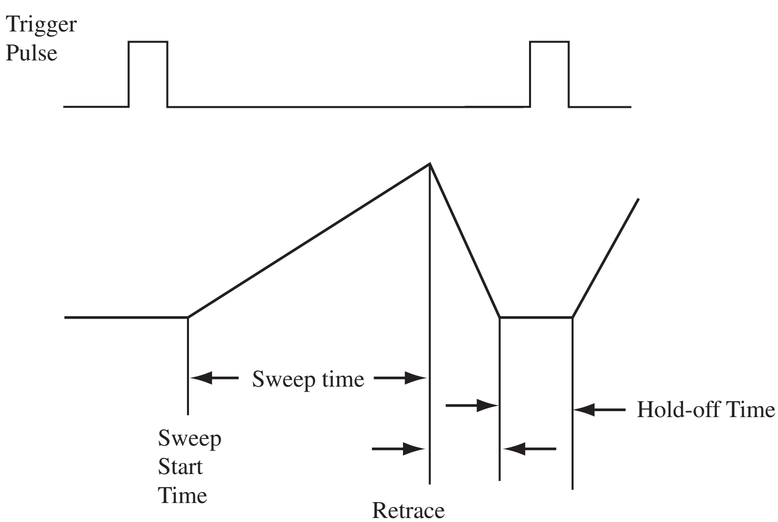


Figure 4.22 Relationship between the trigger pulse and the sweep in an oscilloscope

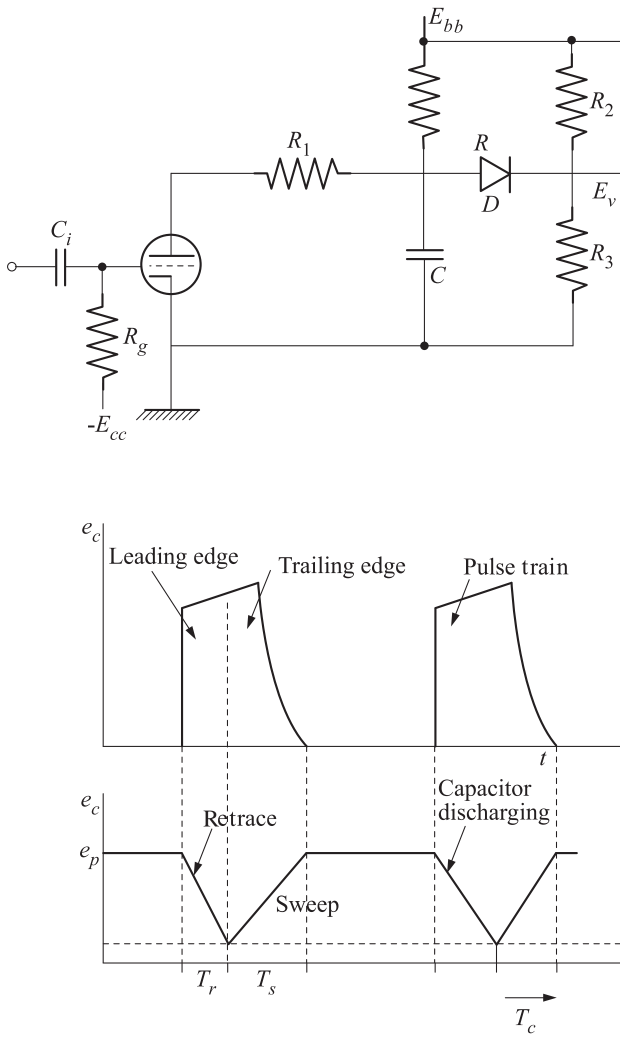


Figure 4.23 Triggering circuit and waveforms

4.10.5 Automode of Sweep

In the auto mode, the sweep can be locked to trigger a signal by the signal itself. This is done when the input signal frequency is > 50 Hz.

4.10.6 Normal Mode

When the input signal frequency is < 50 Hz, the sweep can be stopped and started by the trigger signal itself. In the normal mode, the sweep signal is not free running. It gets locked to trigger a signal.

4.11 TIME BASE CIRCUIT FOR A GENERAL-PURPOSE CRO

The synchronising amplifier is of a conventional design and is preceded by a gain control. The thyratron sweep circuit is essentially the same as that of a diode clamp. The blocking circuit consists of a high-pass RC differentiating network followed by an amplifier (Fig. 4.24).

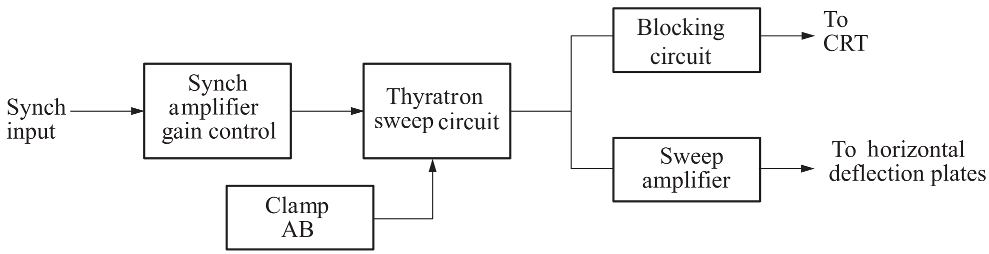


Figure 4.24 Block schematic of a time base circuit

Example 4.1

A time base of an oscilloscope has a resistance 400 kΩ and a capacitance of 0.025 μF. Determine the percentage of non-linearity in a saw-tooth output waveform having a period of 0.4 msec (Fig. 4.25).

Solution

This exponential is for the charge of C because at $t = \infty$, $V = E$.

$$V = E[1 - e^{-t/RC}] = E\left[1 - \left(1 - \frac{t}{RC} + \frac{t^2}{2R^2C^2}\right)\right]$$

At $t/2$, a linear line curve can be made followed by an exponential line:

$$V = E\left[\frac{t}{RC} - \frac{t^2}{2R^2C^2}\right]$$

At $t = \dfrac{t}{2}$,

$$V = V_{2s}$$

$$V_2 = E\left[\frac{(t/2)}{RC} - \frac{(t/2)^2}{2R^2C^2}\right] = \frac{Et}{2RC}\left[1 - \frac{1}{4RC}\right]$$

For a linear rise V_3 at $(t/2)$ should be

$$\frac{V_1}{2} \quad \text{and} \quad V_1 = \left[\frac{t}{RC} - \frac{t^2}{2R^2C}\right]$$

$$\therefore\ V_3 = \frac{V_1}{2} = \frac{E}{2}\left[\frac{t}{RC} - \frac{t^2}{2R^2C^2}\right] = \frac{tE}{2RC}\left[1 - \frac{t}{2RC}\right]$$

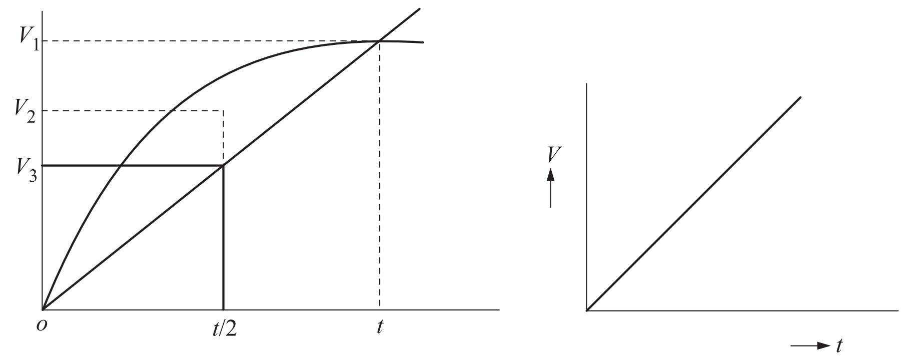


Figure 4.25 Waveforms for Problem 4.1

$$\% \text{ deviation} = \frac{V_2 - V_3}{V_2} \times 100 = \frac{\left(1 - \dfrac{t}{4RC}\right) - \left(1 - \dfrac{t}{2RC}\right)}{1 - t/4RC} \times 100$$

$$= \frac{\dfrac{t}{2RC} - \dfrac{t}{4RC}}{1 - \dfrac{t}{4RC}} \times 100 = \frac{\dfrac{t}{4RC}}{1 - \dfrac{t}{4RC}} \times 100$$

$$= \frac{t}{4RC} \quad \because \quad \frac{t}{4RC} << \text{and} \quad t = 4 \text{ msec}$$

$$\% \text{ deviation} = \frac{4}{10 \times 10^3} \times \frac{1}{4 \times 400 \times 10^3} \times \frac{100}{0.025 \times 10^{-6}} = 1\%$$

4.11.1 Synchronisation Issues

To make C discharge prematurely, the capacitor C charges till the breakdown voltage of the thyratron E_p is reached. When E_p is reached the thryatron conducts and hence the capacitor discharges rapidly through the thyratron. To supply waveform, assume that the capacitor discharges through zero time. The saw-tooth waveform of the thyratron can be synchronised to an external signal. Then the external signal is applied to the synchronised input terminal of the thyratron in such a way that the breakdown voltage (or firing voltage) of the thyratron decreases. This is done when a positive voltage is applied to the given thyratron (Fig. 4.26).

The first several pulses have no effect on the sweep generator. Now if a pulse occurs exactly at the time when the capacitor is discharging (that is, the amplitude of the ramp suddenly becomes 0), the breakdown voltage of thyratron decreases.

For synchronisation to occur, the synchronising pulses must take place at time intervals causing the charging process to stop prematurely (since if a positive voltage is applied to the grid of the thyratron, its breakdown voltage is reduced and therefore the capacitor starts discharging). Therefore, $T_p < T_o$. When $T_p > T_o$, synchronisation does not occur because when $T_p > T_o$, before a pulse is given to the

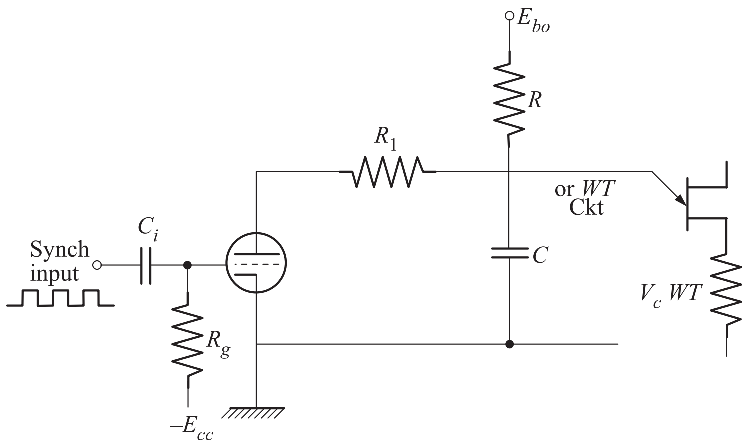


Figure 4.26 Synchronising circuit if $T_p > T_o$; synchronisation will not occur

grid of the thyratron, the capacitor gets charged fully. Another point to note is that if the amplitude of the synchronisation pulses is very small, then synchronisation also does not occur. The amplitude of synchronisation pulses should be such that it can bridge the gap between the quiescent breakdown voltage of the thyratron and the quiescent breakdown voltage minimised by amplitude pulse in starting the ramp (Fig. 4.27).

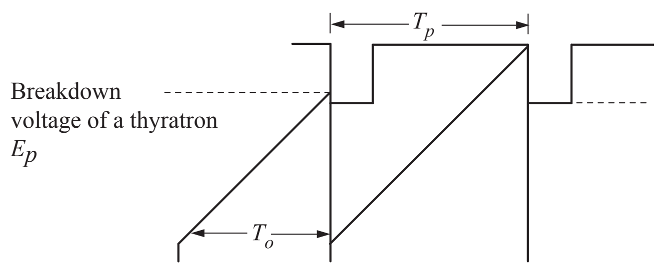


Figure 4.27 Synchronization waveforms

4.11.2 Line Synchronisation

In this mode, the frequency of the screen voltage is equal to the line frequency (supply frequency) or its harmonics. The sweep frequency is preferably locked to the line frequency in this mode (Figs. 4.28, 4.29).

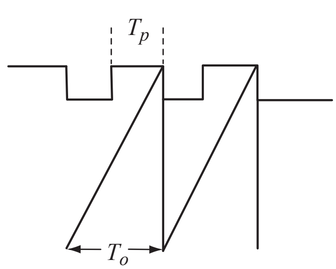


Figure 4.28 Synchronisation

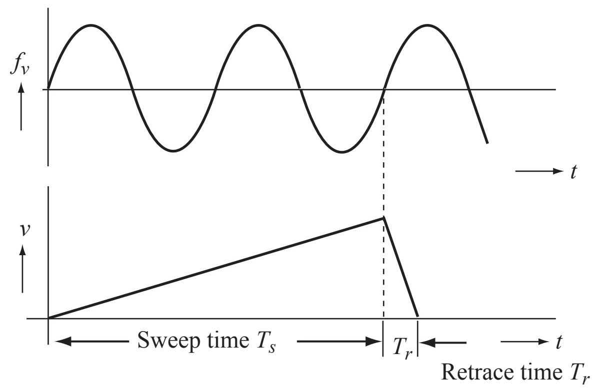


Figure 4.29 Time base synchronisation with input waveforms

If the amplitude of the synchronisation signal is very large, two sweeps will be unitiated. A simplified schematic of a triggered time base circuit is shown in Fig. 4.30.

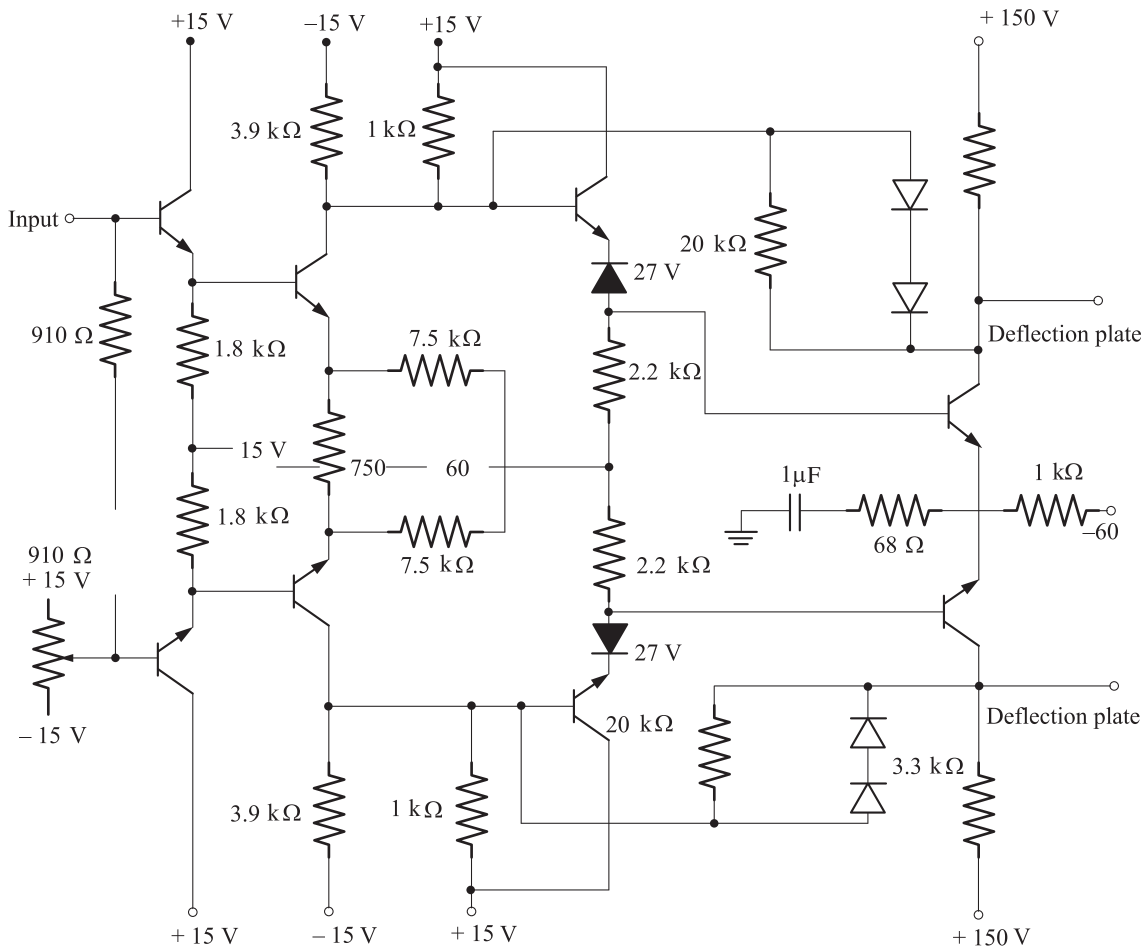


Figure 4.30 Differential deflection amplifier

Example 4.2

A sinusoidal voltage of 83.3 kHz from a standard signal generator gave nine free waves on the screen starting from the *X*-axis when connected to *y* terminals of a CRO, while the tenth wave was slightly short of being a full wave, the end of the trace being at a position that was half the amplitude away from *X*-axis.

(a) If the time base is internally synchronised, determine the rise and decay time of the saw-tooth time base voltage.

(b) Assume that the decay time is the same. When the time base frequency increases to give about four full waves of the same input signal, the time base remaining internally synchronised, what is the rise time of the second saw-tooth time base voltage?

Solution

Being synchronised, the frequency of the saw-tooth wave will be a submultiple of the signal.

$$\therefore \text{ Frequency of saw-tooth curve } = \frac{83.3}{10} = 8.33 \text{ kHz}$$

$$\text{Period of the saw-tooth curve} = \frac{1}{8.33} \times 10^3 \text{ sec} = 120 \text{ μsec}$$

$$\text{Sine wave } y = A \sin\theta$$

$$\frac{y}{A} = \frac{1}{2}, \qquad \because \sin\theta = \frac{1}{2}; \quad \theta = 30°$$

Hence the 10th wave is in short of a complete sine wave by 30°.

$\therefore$ No. of full waves of sine waveform seen on the screen are = $9\frac{330}{360} = 9\frac{11}{12}$ waveforms

Rise time + Decay time = Period of wave = 120 μsec

Rise time/Decay time = $9\frac{11}{12}/(10-9\frac{11}{12})$; $\therefore$ Decay time = $120 \times \frac{\frac{1}{12}}{10} = 1$μsec

$\therefore$ Rise time = (120 –1) = 119 μsec

Increase time base frequency = $\frac{10}{4}$ times the final value

Length of trace blanked in degrees due to flyback time = $\frac{10}{4} \times 30° = 75°$

Period of new time base = $120 \times \frac{4}{10} = 48$ μsec. $\therefore$ Rise Time = 48 – 1= 47 μsec

4.12 LISSAJOUS FIGURES

One of the applications of a CRO is to determine the phase, frequency, amplitude, and other characteristics of a given waveform. These characteristics can be determined in terms of known signal characteristics by the Lissajous patterns or Lissajous figures method. Lissajous patterns named after a scientist called Lissajous result when sine waves are applied simultaneously to both the horizontal and vertical deflection plates of the CRO. The waveform seen on the CRO screen will be the same as that given as the *Y*-input if the internal time base circuit is used. If two different signals are given to the *X* and *Y*-plates, the pattern or figure seen on the screen is the resultant of the two inputs. From the shape or pattern of the resultant figure, the characteristics of the unknown input can be determined in terms of the characteristics of the known input waveform. These are as shown below:

Case 1:

Let

$$Y = A \sin \omega t$$
$$X = B \sin \omega t$$

Two sine wave inputs are given to the X and Y-plates. The CRO is kept in the EXT (External) mode. An internally generated saw-tooth waveform is not applied to the X-plates. The frequency of the two sine wave inputs is the same (ω). However, the peak amplitudes A and B are different. Then the resultant deflection of the electron beam on the CRO screen is

$$\frac{Y}{X} = \frac{A \sin \omega t}{B \sin \omega t} = \left(\frac{A}{B}\right) = m$$

or

$$Y = mX$$

where

$$m = \frac{A}{B}$$

This is the equation of a straight line passing through the origin with slope 'm'. So, the Lissajous pattern is a straight line passing through the origin (Fig. 4.31). Thus, if the Lissajous pattern seen on the screen is a straight line passing through the origin, the frequency of the two signals is the same. The phase shift is 0°, but the peak amplitudes are different. The ratio of the peak amplitudes is equal to the slope of the straight line.

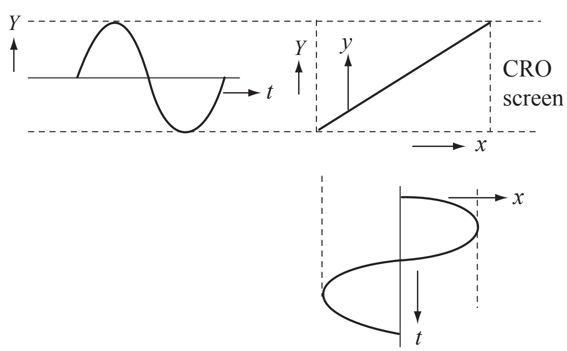


Figure 4.31 Straight-line Lissajous figure

Thus, If frequency of the signal is same
Phase shift between them is zero
Peak amplitudes are different
Lissajous figure is: *A straight line passing through the origin*

Case 2:

Let
$$Y = A \cos \omega t$$
$$X = A \sin \omega t$$

$$X^2 = A^2 \sin^2 \omega t$$
$$Y^2 = A^2 \cos^2 \omega t$$
$$\therefore \quad X^2 + Y^2 = A^2$$

If *the Lissajous pattern is a circle with radius A*, the peak amplitude of the two signals is the same and is equal to the radius of the circle. The frequency of the two signals is the same, but the phase shift between them is 90°. If the phase shift is also 270°, the circle will be formed (Fig. 4.32).

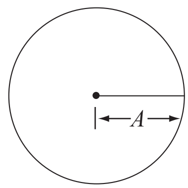


Figure 4.32 Lissajous figure is a circle with radius A when $Y = A \cos \omega t$ and $X = A \sin \omega t$

Case 3:

Let $$Y = A \sin \omega t; \quad X = B \cos \omega t$$

then

$$\frac{Y^2}{A^2} + \frac{X^2}{B^2} = 1$$

This is the equation of an ellipse (Fig. 4.33). If the amplitudes of the two signals are different but the frequencies are the same with a phase shift of 90°, the Lissajous pattern is an ellipse. If the vertical signal has a larger amplitude, that is, $A > B$, then the ellipse will have a vertical major axis. When $B > A$, the main axis of the ellipse curve will be along the horizontal axis. With the help of an ellipse, we can find the phase difference between two signals using a single-beam CRO. If the major axis lies in the first quadrant, the phase angle is between 0° and 90° and when the major axes pass through the third quadrant, it is between 270° and 360°.

Thus, If the peak amplitude of the two signals are different
Frequencies are same
Phase shift between them is 90°
The Lissajous pattern is ellipse

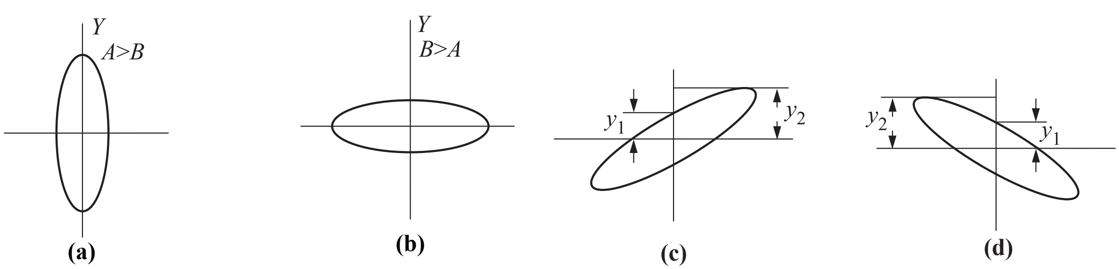


Figure 4.33 Lissajous figure is an ellipse when $Y = A \sin \omega t$ and $X = B \cos \omega t$

When the higher frequency signal ($2\omega t$) is ahead of a low-frequency signal (ωt) and 90°, phase shift is there, a parabola will result (Fig. 4.34). A pattern of this type is called a double image because the electron beam after recasting the direction traces out exactly the same path.

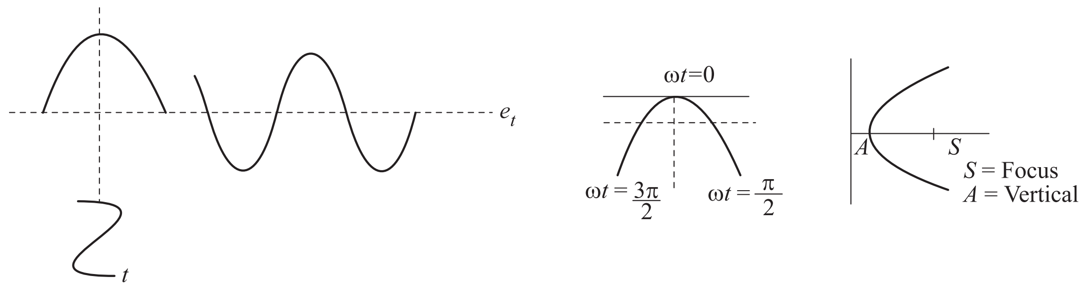


Figure 4.34 Lissajous figure—parabola

Case 4:
Let

$$y = A\cos 2\omega t \quad (\because\ y = A\sin(2\,\omega t + 90°))$$
$$x = A\sin\omega t$$
$$y = A\cos 2\omega t = \cos^2\omega t - \sin^2\omega t$$
$$\therefore\ y = A(\cos^2\omega t - \sin^2\omega t)$$
$$y = A(1 - 2\sin^2\omega t) \quad (\because\ \cos^2\omega t = 1 - \sin^2\omega t)$$
$$\therefore\ y = A\left(1 - \frac{2x^2}{A^2}\right)$$

This Lissajous figure is a parabola.

If the Y signal frequency is twice that of an X signal and there is no phase shift, then a figure of 8 will form along the X-axis. (Fig. 4.35)

$$X = A\sin\omega t; \quad Y = A\sin 2\,\omega t$$

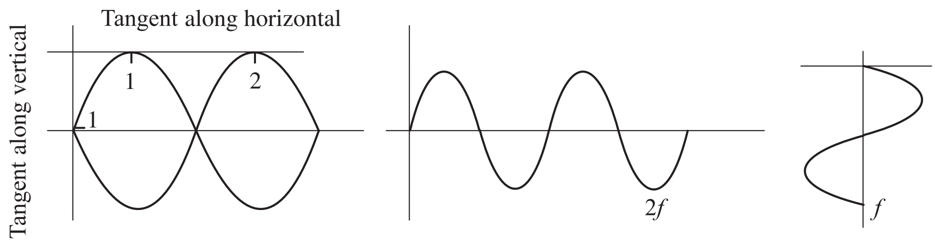


Figure 4.35 Lissajous figure: figure of eight

If the X signal frequency is double the Y signal, then a figure of 8 will form along the Y-axis.

$$Y = A\sin\omega t \quad \text{and} \quad X = \mathrm{A}\sin 2\,\omega t$$

Thus, from the shape of the Lissajous pattern, the characteristics of the unknown signal in terms of a known signal can be established.

The ratio of the frequencies of the X and Y signals must be integers ($\frac{1}{1}$, $\frac{1}{2}$, $\frac{1}{5}$, and so on) in order for the pattern to be stationary. A tangent drawn against the top edge of the pattern would make contact at two places. A tangent drawn against a vertical side would touch at one place.

The horizontal tangent corresponds to the vertical deflection voltage and the vertical tangent corresponds to the horizontal deflection.

Hence, the ratio of frequencies of the $\dfrac{\text{Vertical side}}{\text{Horizontal side}} = \dfrac{2}{1}$

$$\frac{f_Y}{f_X} = \frac{\text{Meeting points of the horizontal tangent}}{\text{Meeting points of the vertical tangent}}$$

When the Lissajous pattern formed on CRO is a parabola, a tangent drawn against an open end of the pattern is counted as a half tangency. For example, a tangent drawn against the parabola along the horizontal axis makes two contacts, which are open ended and, therefore, the contact on tangency is one ($\frac{1}{2} + \frac{1}{2} = 1$). If we adjust the vertical side, there is only one contact; therefore $\frac{1}{2}$ (Fig. 4.36).

$$\therefore \quad \text{The ratio of signals is} \quad \frac{f_Y}{f_X} = \frac{1}{1/2} = \frac{2}{1}$$

$$\frac{f_Y}{f_X} = \frac{1+\frac{1}{2}}{\frac{1}{2}+\frac{1}{2}} = \frac{3}{2}$$

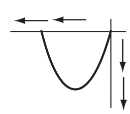


Figure 4.36 Parabola

A figure of a parabola will be formed when the ratio of frequency is 3:2 and the amplitude of the signals is same and there is no phase shift.

1. If $w_x = w_y$ and $\alpha = \dfrac{\pi}{2}$ (Fig. 4.37), an ellipse is formed.

$$X = V_x \sin w_x t$$

$$Y = V_y \cos w_y t$$

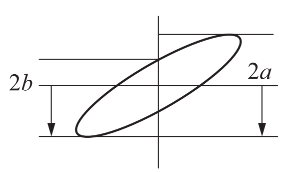


Figure 4.37 Ellipse

$$\frac{X^2}{V_x^2} + \frac{Y^2}{V_y^2} = 1$$

$$X = V_x \sin \omega_x t$$

$$Y = V_y \cos \omega t$$

This gives an ellipse (Fig. 4.38).
We can find the phase difference between two sinusoidal signals:

$$\tan \theta = \frac{A}{B}$$

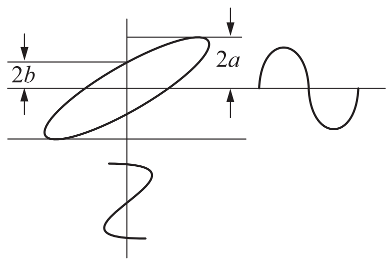


Figure 4.38 Ellipse

$$Y = A \sin(\omega t + \theta), \quad X = A \sin \theta, \quad a = A, \quad \theta = \sin^{-1}\left(\frac{X}{A}\right)$$

2. Let $w_x = w_y$, $\theta = \frac{\pi}{2}$, $V_x = V_y = V$

$$\therefore \qquad X = V \sin \omega t$$

$$Y = V \cos \omega t$$

$$X^2 + Y^2 = V^2$$

This Lissajous figure is a circle (Fig. 4.39).

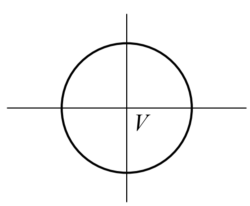


Figure 4.39 Circle

3. If $w_x = 2w_y$,

$$X = V \sin 2\omega t$$

$$Y = V \sin \omega t$$

This gives a figure of 8 (Fig. 4.40). We get a closed figure with multiple loops.

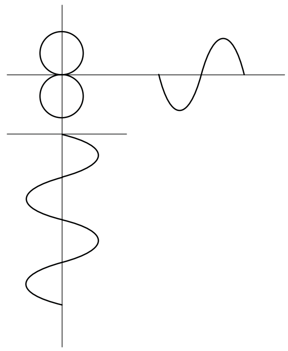


Figure 4.40 Figure of 8

4. If $w_y = 2w_x$, then (Fig. 4.41).

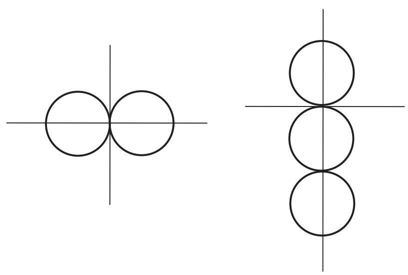


Figure 4.41 Multiple circles

4.13 TYPES OF CRO PROBES

The different types of CRO probes are as follows:

1. Direct probe.
2. High-impedance probe.
3. Detector probe.
4. High-voltage probe.

4.13.1 Direct Probe

A direct probe consists of a coaxial cable with a probe tip. There is a direct connection from CRO to the test point. The cable has a capacitance of 50 pF. This will add up to the shunt capacitance of the vertical amplifier. Therefore, this cable should be used for low frequencies only.

4.13.2 High-Impedance Probe

The high-impedance probe reduces the loading effect by increasing the input resistance and by reducing the input capacitance. The equivalent circuit is shown in Fig. 4.42. The probe consists of a voltage divider network R_1 and R_2, the values of which are taken to get an attenuation of 10:1 over the entire frequency range by adjusting the variable capacitor C_1. Before using the high-impedance probe, it should be made sure that it is properly compensated. This can be done by giving a square wave input signal rich in high-frequency harmonics to the probe tip and by adjusting the variable capacitor C, until an optimum display response is obtained. Therefore, a probe that is properly adjusted for one CRO should not be used for another.

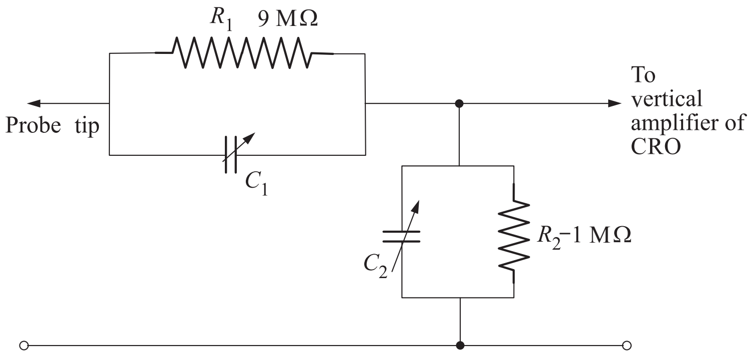


Figure 4.42 Equivalent circuit of a high Z probe

4.13.3 Detector Probe

A detector probe is used in the response analysis of tuned circuits when the shape of the response curve and not the frequency is important. The probe consists of a diode detector. The diode provides rectification of the modulated input signal. The shunt capacitor C_2 bypasses the RF component. Therefore, the low-frequency signal is fed to the CRO. The circuit is shown in Fig. 4.43.

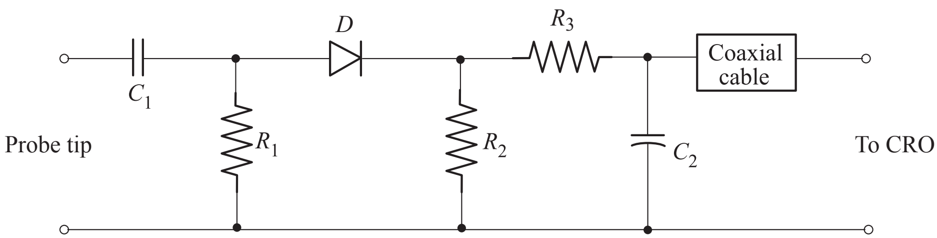


Figure 4.43 Detector probe circuit

While testing circuits operating at frequencies beyond the capability of a vertical amplifier, a demodulator probe is used. If the input is a modulated one, the envelope of the waveform is the output at the probe (Fig. 4.44).

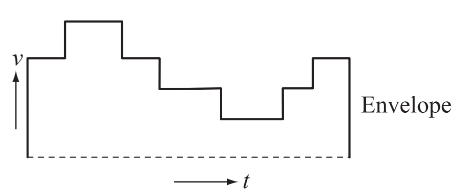


Figure 4.44 Envelope waveform, as the output of the probe

4.13.4 High-Voltage Probe

A high-voltage probe is a cathode follower type. It provides high input impedances and unity gain. Therefore, attenuation will not be there (Fig. 4.45).

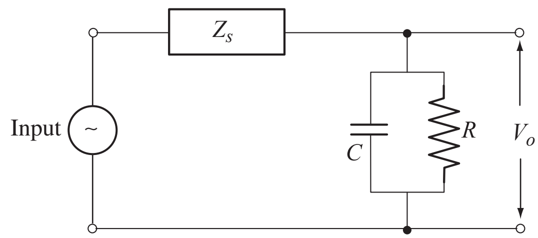


Figure 4.45 Input impedance elements of a CRO probe

Although the resistance *'R'* of the vertical amplifier of an oscilloscope is high, its capacitance C is also large. This necessitates the use of a shielded cable. If shielded cables are not used, the stray fields in the vicinity of the leads would produce intolerable interference in the pattern on the CRO. Therefore, the shielded cable provides high-input impedance with a low capacitance. It is a compromise between high impedance and low attenuation. Because of the high impedance of the cable, the input signal will be attenuated. In most cases the attenuation will be 10:1.

If the time constant R_1C_1 of the probe is made equal to R_2C_2 of the CRO, the complex waveform will be passed through the probe without distortion. In other words, the circuit responds independent of frequency (Fig. 4.46).

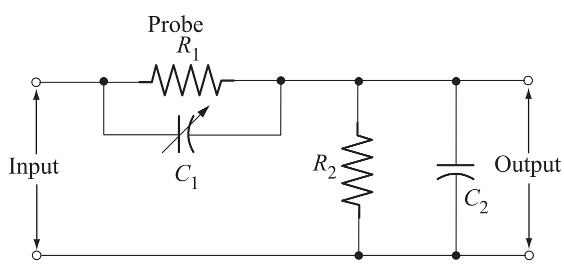


Figure 4.46 CRO probe circuit

If $R_1C_1 = R_2C_2$, the low capacitance probe acts as a simple resistance voltage divider, independent of the frequency:

$$\frac{e_{out}}{e_{on}} = \frac{Z_2}{Z_1 + Z_2}$$

$$= \left.\frac{R_2X_{C2}}{R_2 + C_2}\right| \frac{R_1X_{C1}}{R_1 + X_{C_2}} + \frac{R_2X_{C2}}{R_2 + X_{C_2}}$$

$$= \frac{\dfrac{R_2 \,|\, J\omega C_2}{R_2 + \dfrac{1}{J\omega C_2}}}{\dfrac{R_1 \,|\, J\omega C_1}{R_1 + \dfrac{1}{J\omega C_1}} + \dfrac{R_2 \,|\, J\omega C_2}{R_2 + \dfrac{1}{J\omega C_2}}}$$

$$= \frac{\dfrac{R_2 / J\omega C_2}{(J\omega R_2 C_2 + 1)/ J\omega C_2}}{\dfrac{R_1 / J\omega C_1}{(J\omega R_1 C_1 + 1)/ J\omega C_1} + \dfrac{R_2 / J\omega C_2}{(J\omega R_2 C_2 + 1)/ J\omega C_2}}$$

$$= \frac{\dfrac{R_2}{J\omega R_2 C_2 + 1}}{\dfrac{R_1}{1 + J\omega R_1 C_1} + \dfrac{R_2}{1 + J\omega R_2 C_2}}$$

If $R_1 C_1 = R_2 C_2$,

$$\frac{R_2}{R_1 + R_2} = \frac{e_{at}}{e_{in}}$$

Therefore, if $R_1 C_1 = R_2 C_2$, the low capacitance probe acts as a simple resistance voltage divider, independent of frequency. If a CRO has 100 pF and 1 MΩ impedance for a low-capacitance probe that provides 10:1 attenuation,

$$R_1 = 9M; \quad C_1 = \frac{100}{9} = 11.1 \text{ pF}$$

$$\because \quad R_1 C_1 = R_2 C_2; \quad \frac{R_2}{R_1 + R_2} = \frac{1}{10}$$

$$R_2 = 1 \text{ M}\Omega; \quad C_2 = 100 \text{ pF}$$

$$\therefore \quad R_1 = 9 \text{ M}\Omega; \quad C_1 = 11.1 \text{ pF}$$

4.14 HIGH-FREQUENCY CRO CONSIDERATIONS

If a high-frequency signal is applied to the vertical amplifiers, the rate of change of deflecting voltage increases, since 'f' is large. Because of this, the intensity on the CRT screen decreases. Therefore, to get a pattern of reasonable intensity, the electron beam must be accelerated to a higher velocity so that more interactive energy is available for transfer to the screen, and normal image brightness is obtained. A higher electron beam velocity is achieved by increasing the voltage on accelerating anodes.

Deflective sensitivity

$$S = \frac{D}{E_D}$$

where D = Deflection produced in mm or cm on the screen

E_D = Deflecting voltage (applied to Y or vertical plates)

Deflection

$$D = \frac{Ll_d E_d}{2dE_a}$$

$$D = \frac{E_d}{E_a}$$

E_a = Accelerating voltage in the range 2000 V – 10 kV

d = Distance between deflection plates

L = Distance between screen and deflecting plates

l_d = Length of deflecting plates

E_d = Deflection voltage

If E_a is more, to get the same deflection, E_d should also be increased (to get the same deflection sensitivity). Therefore, the gain in the vertical amplifiers should be large.

A travelling-area-type CRT, which responds to higher frequencies, consists of a series of deflection plates. These deflection plates are so shaped and spaced such that an electron travelling between them will recourse from each plate and produce an additional deflecting force in the proper time sequence. The vertical deflection signal is applied to each plate through a delay line that is so designed that the time delays correspond exactly to the transit time of the electrons travelling down the CRT towards the screen. New fluorescent materials have been developed to increase the image brightness at higher frequencies.

In a vertical amplifier, Gain–bandwidth product is constant for higher frequencies. If the bandwidth (BW) ceases to be large, the gain will reduce.

The (Gain–BW) product is also referred to as the Figure of merit of the amplifier

For an RC-coupled vacuum tube amplifier the voltage gain in a single stage may be expressed as

$$A_v - \frac{E_{\text{out}}}{E_{\text{in}}} = -G_m R_{sh}$$

where G_m is the transconductance and R_{sh} the total shunt resistance. A negative sign indicates a phase shift of 180°.

Total gain of n stages = $A_n = (-J_m R_{sh})^n$. The upper half power frequency of an amplifier is

$$f_n = \frac{1}{2\pi R_{sh} C_{sh}}$$

When n stages are connected in cascade, the bandwidth changes by a focus $\sqrt{n}$; f_{2n} is the upper half power frequency

$$\frac{1}{\sqrt{2}} = \frac{1}{\left[1+\left(\frac{f_{2n}}{f_2}\right)^2\right]^{n/2}}$$

$$1 + \left(\frac{f_{2n}}{f_2}\right) = 2^{1/n}$$

$$\frac{f_{2n}}{f_2} = \sqrt{2^{1/n} - 1}$$

$$\therefore \quad f_n(\text{cascaded}) = \frac{1}{2\pi R_{sh} C_{sh}} \sqrt{n}$$

For the RC-coupled transistor amplifier

$$A_i = \frac{I_{\text{out}}}{I_{\text{in}}} = \frac{h_{ie} R_{sh}}{R_i}$$

Cascading n stages gives $\left(\frac{h_{ie} R_{sh}}{R_i}\right)$. The upper half power frequency = $\frac{1}{2\pi R_{sh} C_{sh}}$.

The gain in an amplifier can be increased substantially by connecting n stages in a cascade. However, the bandwidth reduces. If C_{sh} is less, the bandwidth can be high. C_{sh} can be minimized by choosing a device with low junction capacitances.

1. In high-frequency CROs, because f is large, 'T' of the signal is small. The energy transferred to a CRO plane screen decreases. Hence, the illumination of the waveform seen on the screen decreases.
2. Another problem is that, the bandwidth of the vertical amplifier must be large. Since f of the input signal to the vertical amplifier is large, the gain–bandwidth product for a given amplifier circuit is constant. Hence, if the bandwidth is high, the gain is low.
3. To increase the brightness of an image, the electrons are to be accelerated high. So the accelerating voltage E_a has to be high since the velocity of electrons is directly proportional to E_a.

(Since f is large, T is small. Hence energy transferred to the screen is less.)

Limitations

1. Low intensity of the waveform. (Since T is small (f is large), the phosphorescence effect is limited and the intensity is small.)
2. The bandwidth of vertical amplifiers should be large.
3. Since the gain–bandwidth product is constant for a given amplifier, multistage coupling is to be done.
4. Deflection sensitivity is less, as S_E is inversly proportional to the accelerating voltage.

To overcome these limitations and to increase the intensity of the beam on the screen, the accelerating voltage E_a must be increased. Then the deflection sensitivity D decreases:

$$D = \frac{L l_d E_d}{2 d E_a}$$

If E_a is increased, D decreases. To increase D, E_d should be decreased. Therefore, gain of the vertical amplifiers must be large. However, the gain-bandwidth product for a given amplifier is constant. Therefore, if the gain increases, the bandwidth decreases. To obtain a large voltage gain, cascaded amplifier stages must be employed:

$$f_n = \text{Upper half power frequency} = \frac{1}{2\pi R_{sh} C_{sh}}$$

For n stages,

$$f_n = \frac{\sqrt{n}}{2\pi R_{sh} C_{sh}}$$

To increase f_n, C_{sn} of the amplifier is to be reduced. Therefore, for the given amplifier circuit, R_{sh} and C_{sh} should be less.

4.15 DELAY LINES IN CROs

All electronic circuits in the CROs such as attenuators, amplifiers, pulse shapers in trigger circuits are used to generate trigger pulses, and the lengthy wires used for connectors cause a certain amount of delay in the transmission of the signal voltage to the deflecting plates. In a CRO, a part of the signal being applied to the vertical plates is used to initiate the time base signal. This is called *triggering*.

Now the trigger circuit will cause a delay because of the inherent inductances and capacitances of the circuit, elements, and wires. This delay will be of the order of 80 – 200 nsec. Hence, the horizontal signal and the vertical signal are not synchronised. Owing to this, the details of the leading edge and the trailing edge of a pulse are lost. Hence, a delay line will be incorporated in the circuit. However, this should be after the trigger pick-off. A delay of the vertical signal allows the horizontal sweep to start before vertical deflection. The schematic, after the inclusion of the delay line is shown in Fig. 4.47. The waveform is shown in Fig. 4.48.

There are two types of delay lines:

1. Lumped parameter delay lines.
2. Distributed parameter delay lines.

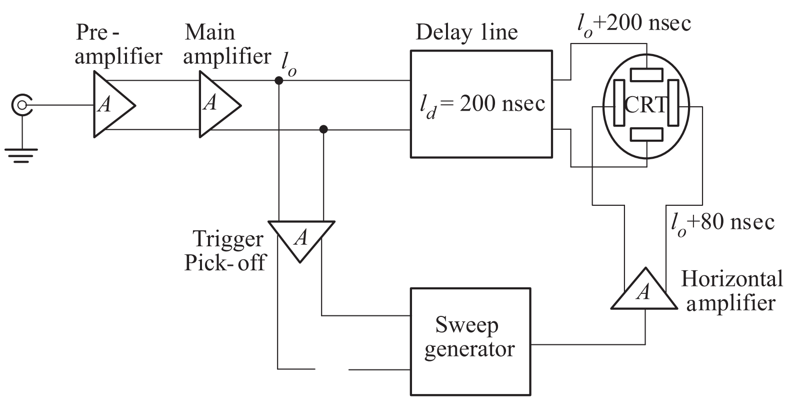


Figure 4.47 Delay line incorporation in CRO

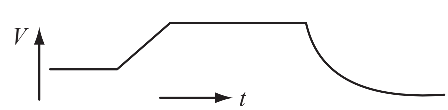


Figure 4.48 Waveform

4.15.1 Lumped Parameter Delay Line

The lumped parameter delay line consists of a number of cascaded symmetrical LC networks, that is, *T*-sections. This is shown in Fig. 4.49.

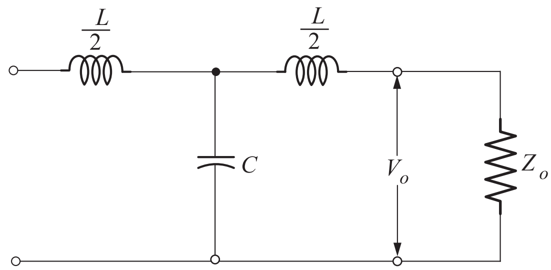


Figure 4.49 *T*-section of lumped parameter delay line

The upper limit of the pass band, the cut-off frequency of the filter, is given by

$$f_c = \frac{1}{\pi\sqrt{LC}} \tag{4.8}$$

If the spectrum of the input signal V_i consists of frequencies much less than the cut-off frequency, the output signal V_o will be a faithful reproduction of V_i, but delayed by a time,

$$t_s = \frac{1}{\pi fc} = \sqrt{LC} \tag{4.9}$$

where t_s is the time delay for a single *T*-section. Number of *T*-sections cascaded into a lumped parameter delay line increases the total time delay to, $t_d = nt_s$, where n is the number of cascaded *T*-sections. Because of the sharp cut-off frequency of the lumped parameter delay line, amplitude, and phase, distortion becomes a problem when the frequency of the input signal increases. It is important to match the delay line as closely as possible to its characteristic impedance Z_o at both the input and the output ends.

The distributed parameter delay line consists of a specially manufactured coaxial cable with a high value of inductance per unit length. For this type of delay line, the straight central conductor of the normal coaxial cable is replaced by a continuous coil of wire, wound in the form of a helix on a flexible inner core. To reduce eddy currents, the outer conductor is usually made of banded insulated wire, electrically connected at the ends of the cable. The constructional details are shown in Fig. 4.51.

The inductance of the delay line is produced by the inner coil and it equals that of a solenoid with n turns per meter. The inductance can be increased by winding the helical inner conductor on a ferromagnetic core, which has the effect of increasing the delay time and the characteristic impedance Z_o. The capacitance of the delay line is that of two coaxial cylinders separated by a polyethylene dielectric. The capacitance can be increased by using a thinner dielectric spacing between the inner and outer conductors. Typical parameters or a helical high- impedance delay line Z_o = 1 kΩ, t_d = 180 ns/m.

The cut-off frequency of the filter is

$$f_c = \frac{1}{\pi\sqrt{LC}}$$

The delay time of a single T-section

$$t_s \frac{1}{\pi\sqrt{LC}} = \sqrt{CL}$$

The total delay time $t_d = nt_s$.

4.15.2 Distributed Parameter Delay Line

For this, a continuous coil of wire wound in the form of a helix on a flexible inner core is used and inductance is produced (Fig. 4.50) by the line.

The capacitance of the delay line is that of two coaxial cylinders, separated by a polyethylene dielectric. Typical values (Fig. 4.51) are:

$$Z_o = 1000\ \Omega$$
$$t_d = 180\ \text{nsec/m}$$

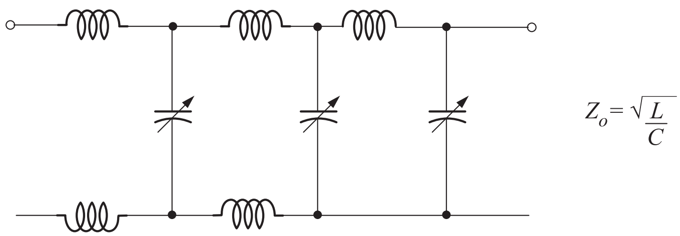


Figure 4.50 Distributed parameter delay line

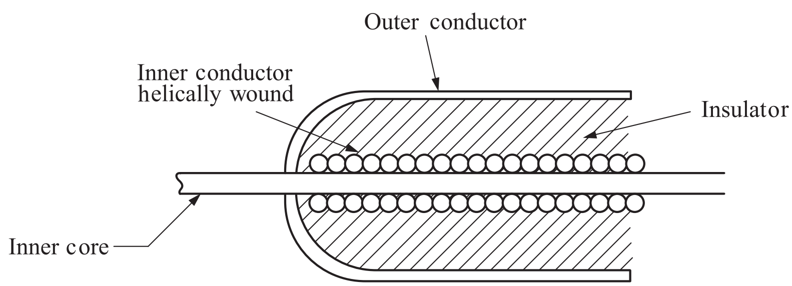


Figure 4.51 CRO cable structure

4.16 APPLICATIONS OF CRO

1. Television: A CRT is used along with sweep circuits.
2. Radar: A CRT screen is used to provide visual displays.
3. To measure voltage or current, the drop across a resistor is measured.
4. To study waveforms: A visual display is seen.
5. Measurement of frequency and phase angle.

Specification of CRO (with typical values)

Model No.	xxxxx
Type	100 MHz, 2 channel, 6 trace CRO
Vertical deflection channels 1 and 2	2 mV/div to 5V/ div in 11 steps in 1–2–5 sequence
Accuracy	±3%
Bandwidth	DC-100 MHz
Rise time	3.5 msec or less
Input impedance	1 MΩ in parallel with 22 pF
Maximum input voltage	400 V (DC + AC peak)
Sensitivity	0.5 V/div
Chop frequency	500 kHz per channel
Time base (A and B)	20 nsec/div. to 0.5 sec/div
Time accuracy	±3%
Sweep magnification	×10

Example 4.3

Determine the velocity of the electron beam of an oscilloscope when the voltage applied is 2500 V.

Solution

$$V_a = 2500 \text{ V}$$

$$e = 1.602 \times 10^{-19} \text{ C}$$

$$m = 9.107 \times 10^{-31} \text{ kg}$$

$$\text{Velocity } V = \sqrt{\frac{2eV_a}{m}} = 29.65 \times 10^6 \text{ m/sec}$$

Example 4.4

The deflection sensitivity of a CRT is 0.05 mm/V and an unknown voltage is applied to the horizontal deflection plate, which shifts the spot by 5 mm towards the right. Determine the unknown applied voltage.

Solution

$$\text{Deflection sensitivity} = 0.05 \text{ mm/V}$$

$$\text{Spot deflection} = 5 \text{ mm}$$

$$\text{Applied voltage} = \frac{5}{0.05} = 100 \text{ V}$$

Example 4.5

The *x*-deflection plates of a CRT are 20 mm long and 5 mm apart. The centre of the plate from the screen is 25 cm away. The accelerating voltage is 3000 V. Determine the deflection sensitivity and the deflection factor.

Solution

$$l = 20 \text{ mm}; \qquad s = 0.25 \text{ mm}$$
$$d = 5 \text{ mm}; \qquad V_a = 3000 \text{ V}$$
$$D = \frac{I_S}{2dV_a}$$
$$Df = \frac{1}{p}$$
$$\text{Deflection sensitivity} = \frac{0.02 \times 0.25}{2 \times 0.005 \times 3000}$$
$$= 0.1667 \text{ mm}$$
$$\text{Deflection factor} = \frac{1}{\text{Deflection sensitivity}} = \frac{1}{0.1667}$$
$$= 6 \text{ V/mm}$$

Example 4.6

The x-deflection plates in the CRT are 1 mm apart and 25 mm long. The centre of the plate is 20 cm from the screen. The accelerating voltage is 3000 V. Find the V_{rms} of the sinusoidal voltage applied to x-deflection plates if the length of the trace is 10 cm. Find the electrostatic deflection sensitivity.

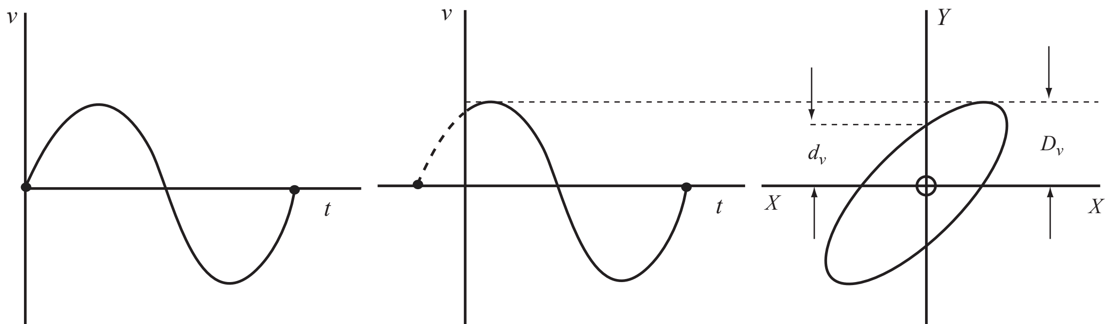
Solution

$$l = 0.025 \text{ m}; \quad d = 0.005 \text{ m}; \quad L = 20 \text{ cm} = 0.2 \text{ m}$$
$$s = 0.2 \text{ m}; \quad V_a = 3000 \text{ V} = E_a$$
$$\text{Trace length} = 2y = 10 \text{ cm}$$
$$\text{Deflection produced, } s = \frac{y}{V_d} = \frac{Ll_d}{2dV_a}; \qquad \therefore y = \frac{Ll_d V_d}{2dV_a}; \qquad V_d = \frac{2dV_a y}{Ll_a}$$
$$V_d = \frac{2dV_a y}{Ll_d} = \frac{2 \times 0.005 \times 0.05 \times 3000}{0.2 \times 0.025}$$
$$V_{rms} = \frac{300}{\sqrt{2}}$$
$$= 212 \text{ V}$$
$$s = \text{Deflection sensitivity} = \frac{Ll_d}{2dV_a} = \frac{0.025 \times 0.2}{2 \times 0.005 \times 3000}$$
$$= 0.167 \text{ mm/V}$$

Example 4.7

Two sinusoidal voltage signals of equal frequency are applied to vertical and horizontal deflection plates of a CRO. Find the phase angle in each trace.

$$\theta = \sin^{-1} \frac{dvo}{DV}$$



(a) (b) (c)

Solution

(a)

$$\theta = \sin^{-1} \frac{dvo}{DV} = 0°$$

(b)

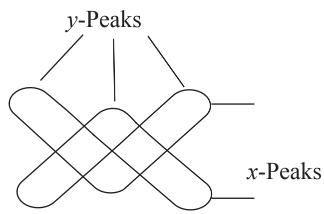
$$\theta = \sin^{-1} \frac{3}{6} = 30°$$

(c)

$$\theta = \sin^{-1} \frac{1}{1} = 90°$$

Example 4.8

A closed Lissajous pattern is shown in the figure. Find the ratio of frequencies of vertical and horizontal signals.

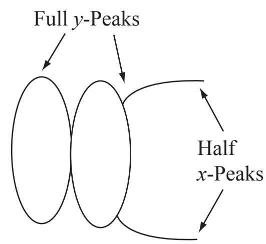


Solution

$$\frac{fy}{fx} = \frac{wy}{wx} = \frac{\text{positive } y\text{-peak pattern}}{\text{positive } x\text{-peak pattern}} = \frac{3}{2} = 1.5$$

Example 4.9

For a Lissajous pattern shown in the figure, find out the frequency of the vertical signal if the frequency of a horizontal signal is 3 kHz.



Solution

$$y\text{-Peak pattern} = 2 + \left(\frac{1}{2}\right) = 2.5 = 2\tfrac{1}{2}$$

$$x\text{-Peak pattern} = \frac{1}{2} + \frac{1}{2} = 1$$

$$\frac{fy}{fx} = 2\tfrac{1}{2}$$

$$\text{From vertical voltage signal} = 2\tfrac{1}{2} \times 3 = 75 \text{ kHz}$$

4.17 SUMMARY

Cathode ray oscilloscopes (CROs) are extensively used in electronic circuit testing for the visualisation of electronic signals and the measurement of signal parameters. An electron beam generated in the CRO is deflected by the given electrical signal and the beam so deflected strikes the screen in a CRO. Owing to the phosphorescence effect, a visible representation of the electrical signal is seen. A cathode ray tube (CRT) is the heart of a CRO. Vertical amplifiers, horizontal amplifiers, time base generators, and trigger circuits are the other blocks in a CRO. Graticules are scale markings provided on the CRT screen. The signal to be observed is applied to a vertical amplifier or *Y*-plates. A time base signal that is a saw-tooth waveform is applied to a horizontal amplifier or *X*-plates in the conventional mode. The time base signal must be a sawtooth waveform because the electron beam must be deflected in the horizontal direction linearly to obtain a time representation with respect to time. When the electron beam reaches the edge of the screen, the beam must return to the starting point. Hence, this is achieved by a saw-tooth wave shape only. If different signals are applied to *X* and *Y*-inputs of a CRO based on their phase, frequency, and amplitude relationship, Lissajous figures are seen on the CRO screen.

Points to Remember

- CROs are used to determine the amplitude, frequency, and phase of a given signal.
- CRT is the heart of CRO. Horizontal amplifier, vertical amplifier, time base generator, trigger circuits are the other blocks in a CRO.
- Beam intensity can be varied by changing the accelerating anode potential.
- Beam focus adjustment can be made by varying the focusing anode potential.
- Graticules are the scales provided on the CRO screen.
- Deflection sensitivity of a CRO/CRT is expressed as the deflection produced for a unit-deflecting voltage. It is expressed in cm/V.

- In the triggered mode, the sweep is initiated by the signal waveform (to be observed) itself.
- If the time base is repetitive, and a new sweep is started immediately after the previous sweep is terminated, it is known as a free-running stable or recurrent mode.
- Phosphor materials have persistence, i.e., even if the signal is removed, the trace is seen on the screen for some time. This period varies from μsec to a few seconds.
- Based on the phase, frequency, and amplitude relationship between the two signals applied to the *X* and *Y*-plates, various shapes or Lissajous figures are seen on the CRO screen. If $Y = A \sin wt$ and $X = B \cos wt$, an ellipse is seen.

Objective-type Questions

1. Vertical amplifier in CRO is also known as _______.
2. The shape of a waveform used as the time base is _______.
3. The order of voltage required for accelerating the electron beam deflect and sweep is _______.
4. Triggering of the signal in CRO is done to get _______.
5. The type of electron emission that takes place in a CRT is _______.
6. Scale markings provided on the front side of the CRT screen are called _______.
7. The control knob LEVEL in a CRO is used to _______.
8. If the saw-tooth waveform is repetitive and a new sweep is started immediately after the previous sweep is terminated, the mode of operation is known as _____.
9. If the frequency of the two signals *X* and *Y* are the same, but the peak amplitudes are different, the Lissajous pattern seen on the CRO screen is _______.
10. The two types of storage techniques used in analog CROs are _______.
11. In a cathode ray tube, the focusing anode is located between _______ and _______.
12. The source of an electron emission in a cathode ray tube is _______.
13. Phosphor coating for a CRT screen is given _______ the surface.
14. The purpose of Acquadag coating in CRT is _______.
15. The focus adjustment of a CRO varies the _______ voltage.
16. A steady waveform on a CRO screen is obtained within _______.
17. Frequency of saw-tooth waveform is varied by _______ in a CRO.
18. The integrator circuit used in a CRO time base is _______.
19. The patterns used to measure phase and frequency within a CRO are called _______.
20. The Lissajous pattern that is seen on a CRO screen if the two signals are equal in magnitude and frequency but with a phase shift of 90° is _______.

Review Questions

1. What are the considerations to be made in designing high-frequency CROs?
2. With the help of a block schematic, explain the operation of a sampling CRO.
3. Compare analog and digital storage CROs.
4. Give the block schematic of a digital storage CRO used for storing transients.
5. Draw the neat sketch of an analog storage CRT and explain its working.
6. Describe the various applications of CROs.

Unsolved Problems

4.1 Determine the velocity of an electron beam in an oscilloscope if the accelerating voltage is 3000 V.

4.2 For a given CRO, the electrostatic deflection sensitivity S_E is 0.08 mm/v. The applied voltage to the *Y*-plates causes a deflection of the spot by 3 mm towards the right. Determine the value of the deflecting voltage.

4.3 In a CRT, the length of the deflecting plates is 3 cm. The distance between them is 6 mm. The centre of the deflecting plates is 20 cm away from the screen. If the value of the applied accelerating voltage is 2 kV, determine the deflection sensitivity S_E.

4.4 In the case of a CRT, the deflecting plates are 30 mm long and 6 mm apart. The centre of the plate is 22 cm away from the screen. The accelerating voltage applied V_a = 3.1 kV. Determine the value of the sinusoidal voltage applied to *X*-deflecting plates if the length of the trace is 10 cm. Determine the value of S_E.

4.5 Two sinusoidal voltage signals of the same frequency but with different peak amplitudes and with a phase difference are applied to the *X* and *Y*-plates of a CRO. What is the shape of the Lissajous figure seen on the screen ? If the *Y*-deflection trace maximum and minimum values are 5 mm and 2.5 mm, determine the phase difference between the two signals.

4.6 A Lissajous pattern has three *Y*-peaks and two *X*-peaks Find the ratio of frequencies of vertical and horizontal signals.

4.7 A Lissajous pattern is in the shape of figure 8 in the horizontal direction. If the frequency of the horizontal signal is 4 kHz, determine the frequency of the vertical signal.

4.8 The horizontal deflection plates of a CRT are 5.5 mm apart and 18 mm long. The centre of the plate is 18 cm from the screen. The accelerating voltage is 2600 V. Determine the volume of electrostatic deflection sensitivity.

5

Special Types of CROs

Special types of oscilloscopes • Dual beam CRO • Dual trace CRO • Sampling oscilloscope • Storage oscilloscopes • Digital storage CRO • Frequency/Period-timer/Counter circuits • Frequency measurement • Period measurement • Errors in frequency/Period measurements • Universal counters • Extending the range of frequency counters • Glossary • The ABC's of oscilloscopes • Summary

5.1 SPECIAL TYPES OF OSCILLOSCOPES

In some applications, the input signal given to the circuit or system and the output signal are to be observed simultaneously and analysed. Hence, the cathode ray oscilloscope (CRO) must have provision to give two signals to two different *Y*-plates and separate time bases. Such CROs are called *double beam oscilloscopes*. The bandwidths of the vertical and horizontal amplifiers of the CRO limit the maximum frequency of the input signal that can be given to the *Y*-input of the CROs. Therefore, CROs with provision for maximum frequency are to be constructed. These are called *high-frequency CROs*. In the conventional cathode ray tube (CRT), the *persistence* of the phosphor ranges from a few milliseconds to several seconds. Hence, an event that occurs only once will disappear from the screen after a relatively short period of time. To study such signals, storage CROs are to be constructed. All such CROs are categorised as special-purpose or special-type CROs. In this unit, such CROs and associated circuits are described. The different types of phosphors, the colour given by them, relative luminescence, and other details are given in Table 5.1.

Table 5.1 Types of phosphors used for CRO screens

Phosphor type	Fluorescence	Phosphore-scence	Relative luminescence (%)	Decay to 0.1% (ms)	Comments
P_1	Yellow-green	Yellow-green	50	95	General-purpose replaced by P_{31} in most applications
P_2	Blue-green	Yellow-green	55	120	Good composition for high and low-speed applications

P_4	White	White	50	20	Television displays
P_7	Blue	Yellow-green	35	1,500	Long decay observation of low-speed phenomena
P_{11}	Purple-blue	Purple-blue	15	20	Photographic applications
P_{31}	Yellow-green	Yellow-green	100	32	General-purpose brightest available phosphor

5.2 DUAL BEAM CRO

The dual beam CRO is built around a special CRT, which displays two completely independent beams. Thus, the CRT has two electron guns and two independent sets of vertical and horizontal deflection plates. Unusual versatility in displaying is achieved through many different combinations of vertical amplifiers and time bases. For example, one channel may be used to show the phase shift between two sine wave signals while a related function is displayed with time variation on the other channel. Slow and fast signals can be observed simultaneously. The same signal can also be studied by giving the signal to Y_1 and Y_2-inputs of the double beam CRO by sweeping the signals at different sweep rates.

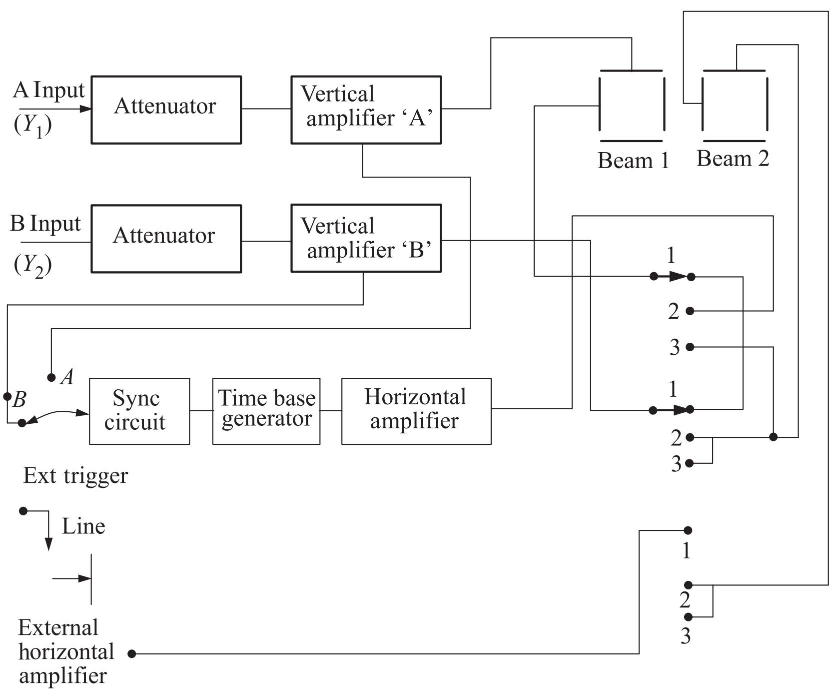


Figure 5.1 Line schematic of a double beam CRO

A double beam CRO uses two identical vertical amplifiers with a sensitivity of 100 div/cm or better and a bandwidth of typically 500 kHz or above. A double beam CRO can also be used as a surge beam CRO by keeping the used Y_2 beam of the screen and giving the input to the other vertical input. The CRO can use one time base in the range typically from 1µsec/cm to 5 sec/cm. However, in such a case, both the signals Y_1 and Y_2 can be swept at the same rate. Some CROs may have two independent time base signals. Here, the flexibility will be more but the cost of the CRO will also be high. The block schematic of a dual beam or a double beam oscilloscope is shown in Fig. 5.1.

When the display control switch is in position 1, the beam is used for a single beam operation and produces *y–t* plots. When the playswitch is in position 2, both beams are used, one with an external horizontal amplifier to produce *x–y* plots and the other with the internal sweep. In the third position of the switch, dual plots can be made by using an external horizontal amplifier to drive both beams simultaneously. A typical value of the sensitivity of the vertical amplifiers is 100 µV/cm. The bandwidth of the CRO is typically 1 MHz. The cost of the CRO increases as the flexibility increases.

It is built in the CRO but is not connected to the time base.

Switch position 1: Single beam CRO, $X - Y$ plots.
Switch position 2: Dual beam, $X - Y$ plot, X versus t plot.
Switch position 3: Dual beam, $X - t$ and $Y - t$.

A digital readout CRO is to introduce the concept of providing digital readout of signal information such as voltage or time, in addition to the conventional CRT display. The circuits of both are connected by means of a display logic control. The digital readout CRO gives a readout of the rise time, amplitude, and time difference.

5.3 DUAL TRACE CRO

In this type of CRO, two separate vertical input signals can be displayed simultaneously. The CRO consists of a single beam CRT, single time base generator, and two identical vertical amplifiers with an electronic switch. The output of the vertical amplifiers is connected to the electronic switch via a mode control switch. It allows two modes of operations:

1. The alternate mode.
2. The chopped mode.

In the alternate mode, the electronic *switch alternately connects each vertical amplifier to the CRT*. The switching rate of the CRT is controlled by the sweep rate of the time base generator. Hence, the CRO alternately displays the two vertical signals. Each vertical amplifier has its own calibrated input alternators and position controls so that the amplitude of each signal can be separately adjusted.

In the chopped mode, the electronic switch *connects the vertical amplifiers irrespective of the time base frequency*. The chopping rate is approximately 100 kHz or 0.01 msec. This is of greater use for simultaneous observation of the low-frequency signals, whose frequency is much lower than the chopping frequency.

5.4 SAMPLING OSCILLOSCOPE

If a very high-frequency signal is to be observed on the CRO, the writing speed of the electron beam increases. Therefore, the image intensity reduces. If the same intensity is to be maintained, then the electrons must be accelerated to a higher speed. Hence, the accelerating potential is to be increased.

However, the deflection sensitivity decreases. To maintain the same sensitivity, the deflecting potential must be increased. Therefore, the vertical amplifier needs to be redesigned and many modifications are to be made. In the sampling oscilloscope, a different technique is used. Here the waveform is reconstructed from many samples taken during recurrent cycles of the input waveform. Hence, the bandwidth of the amplifier can be low and a low deflection sensitivity will not be a problem. The technique is as shown in Fig. 5.2.

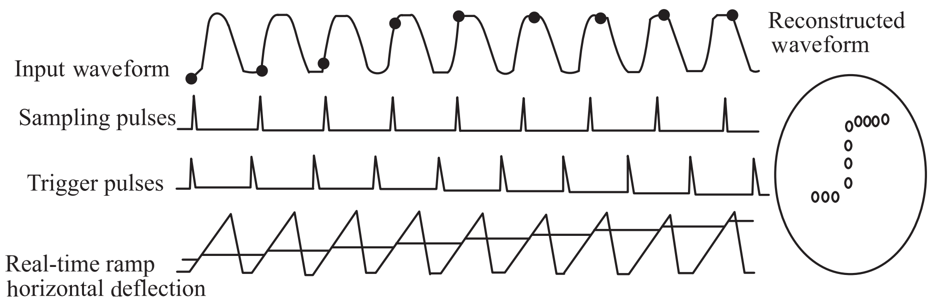


Figure 5.2 Reconstructing a high-frequency waveform in a sampling CRO

In reconstructing the high-frequency waveform, the sampling pulse turns the sampling circuit ON for a very short time interval. The waveform voltage at that instant is measured. The CRT spot is then positioned vertically to the corresponding voltage input. The next sample is taken during a subsequent cycle of the input waveform at a slightly later position. The CRT spot is moved horizontally over a very short distance and is repositioned vertically to the new value of the input voltage. In this manner, the oscilloscope plots the waveform point by point using as many as 1000 samples to reconstruct the original waveform. The sample frequency may be as low as one-hundredth of the input signal frequency. If the input signal has a frequency of 100 MHz, the required bandwidth of the amplifier would be only 10 MHz, a very reasonable figure.

The block schematic of the sampling circuit is shown in Fig. 5.3. The input waveform that must be repetitive is applied to the sampling gate. When a trigger pulse is received, the available blocking oscillator starts a linear ramp voltage, which is applied to the voltage comparator.

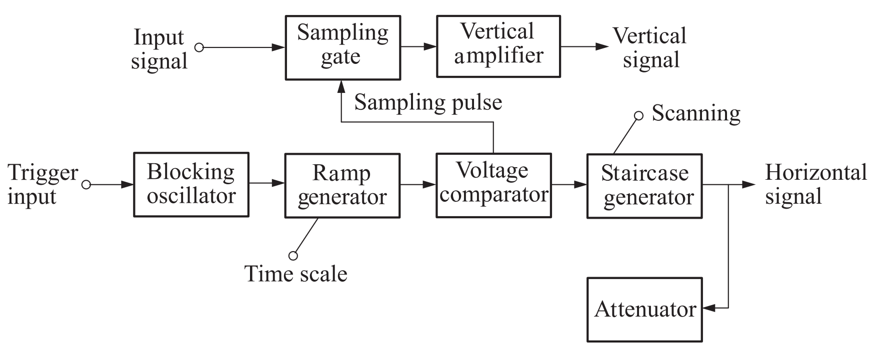


Figure 5.3 Simplified block diagram of the sampling circuitry

The ramp voltage is compared with a staircase voltage. When the two voltages are the same, the staircase generator is allowed to advance one step further and at the same time a sampling pulse is given to the sampling gate. Therefore, a sample of the output is taken, amplified, and applied to vertical deflecting plates.

5.4.1 Sampling Oscilloscopes — Vertical and Time Base

High-frequency CROs though designed even upon 1 MHz suffer from the following:

1. Poor sensitivity.
2. Low brightness.
3. A small CRT display area.

Sampling oscilloscopes avoid these problems of the real-time CROs by translating high-frequency signals to a low-frequency domain. Instead of continuously monitoring the signal under test, the sampling device of the oscilloscope samples the signal amplitude at regular intervals.

5.4.2 Sampling Vertical

A basic sampling circuit is shown in Fig. 5.4.

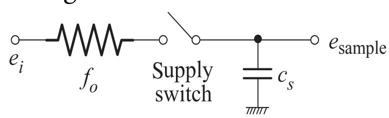


Figure 5.4 Sampling circuit

The sampling switch is a diode. When the sample is applied, it is forward biased and checked (switch closed). When no input is given, the diode is reverse biased. A four-diode network works better than a single diode. As the frequency of the input signal is high, the diode will be forward biased for a start period.

The system to be sampled is represented by a voltage source e_i with z, z_o. The sampler consists of a gate l_a and sampling capacitor C_s. When a sample is to be taken, the switch or the gate is closed for a short period, which allows a sampling capacitor C_s to charge to some fraction of e_i. The switch is then opened with the sample of the input left stored on C_s. If the voltage on C_s is assumed to be reset to zero before each new sample, the efficiency (η) of the circuit called sampling efficiency is given by

$$\eta = \frac{\text{Voltage on } C_s \text{ after each sample } (e_{\text{sample}})}{\text{Input voltage}}$$

The bandwidth of the sampler circuit is the frequency at which

$$\eta = f_L \text{ (low frequency use) } e_{samp} \cong 0.05 e_i$$

Sampling is accomplished by momentarily closing the sampling switch. The sampled voltage is transferred to the capacitor C_i. This voltage is amplified and sent to the stretcher or the memory switch. This switch is closed at the same time as the sampling switch, but remains closed for a much longer time than the former. Therefore, C_s will get charged to the full voltage of the output of the AC amplifier. This voltage is applied to the vertical amplifier where it is amplified sufficiently to drive vertical deflection plates of the CRT. This new level is also fed back through a feedback attenuator RFB to the input capacitor. The gain in an AC amplifier and feedback circuit is so designed that the voltage feedback to the input capacitor will be 100% of the sampled signal level. When the next sample is taken, only changes from the previous level will be detected (Fig. 5.5).

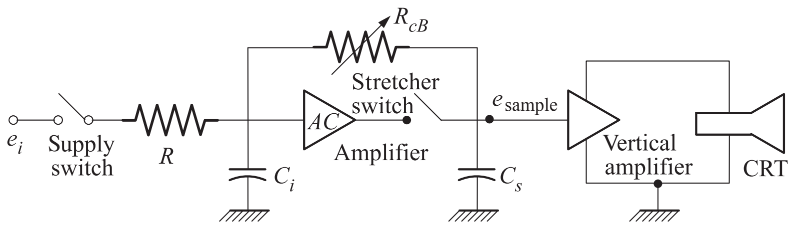


Figure 5.5 Sampling of input signal for vertical amplifier of CRO

5.4.3 Sampling Time Base

Figure 5.6 shows an entire sampling system. The *X*-axis system consists of a synchronised circuit, time base, and horizontal amplifier. The synchronised circuit determines the sampling rate and establishes a reference point in time with respect to the signal. The time base generates both a timing ramp and a staircase waveform, which advances one step per sample. A coincidence of timing of the ramp and the staircase level provides a sampling command to the sampler and stretcher switches. The horizontal amplifier builds the time base signal to a sufficient amplitude to drive the horizontal deflection plates.

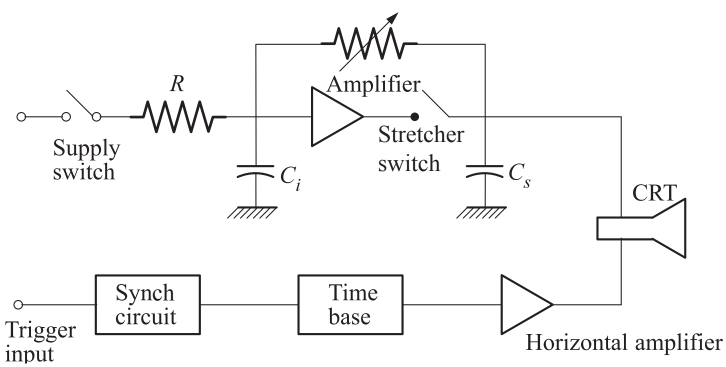


Figure 5.6 Sampling system

The sampling time base moves the beam across the screen but not as a continuous movement. It positions the beam horizontally after a sample is taken and holds the beam at this location until the next sample is taken.

The CRO plots the waveform point by point using as many as 1000 samples or more to reconstruct the original waveform. The sample frequency may be as low as one-hundredth of the input signal frequency. If the input signal has a frequency of 1000 MHz, the bandwidth of the amplifier would be only 10 MHz (Fig. 5.7). The block schematic of a sampling oscilloscope is shown in Fig. 5.7.

The input waveform that must be repetitive is given to the sampling gate. This forward biases the diodes, thereby momentarily connecting the gate input *C* to the test front. Therefore, the capacitance gets charged and this is amplified and given to vertical deflection plates. The sampling has to be synchronised with the input signal frequency; hence, the signal is delayed in the vertical amplifier, allowing sweep triggering to be done by the input signal. When the triggered pulse is received, the oscillator starts a linear ramp voltage, which is applied to the voltage comparator. This comparator compares the ramp voltage to the output of a staircase generator. When these two are equal, the staircase generator is allowed to advance one step ahead in the horizontal device, and simultaneously a sampling pulse is applied to the sampling gate. At this point, a sample of the input voltage is taken, amplified, and applied to the vertical deflection plates (Fig. 5.8).

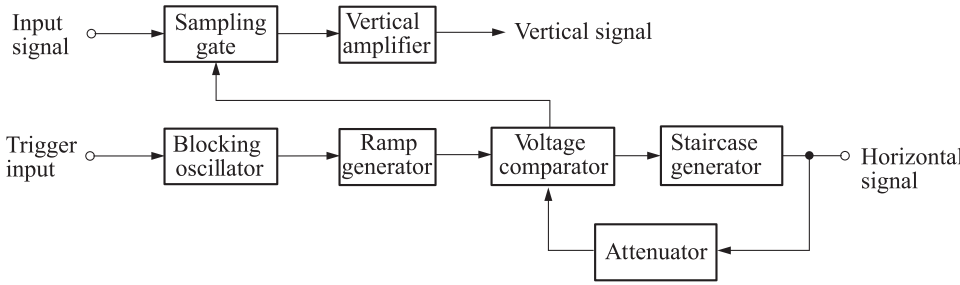


Figure 5.7 Block schematic of a sampling oscilloscope

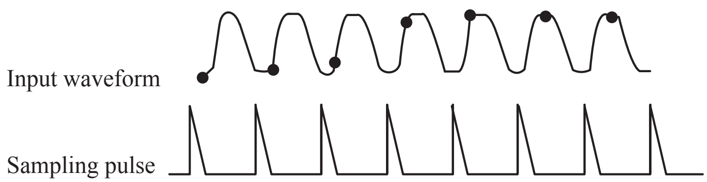


Figure 5.8 Sampling CRO waveforms

Example 5.1

In the process of photography of CRO image and grid lines, an exposure of (1/100) sec is set for grid line impression, which is equivalent to 1 candle power (CP) intensity. In the second exposure of the film for 10 sec the transient is to be photographed and the CRO screen has 10^{-4} sec persistence. If the trace intensity is 4 CP for normal setting of CRO voltages, estimate the boost required for electron gun-accelerating voltage while displaying transient for photography.

Solution

The emission of light that would be received by photographic paper in both exposures must be the same, and the product of time and light intensity is to be the same.

$$\text{Hence, the image intensity required} = 1 \times \frac{1/100}{10^{-4}} = 1000$$

$$\therefore \quad \text{The intensity boost required} = \frac{100}{4} = 25 \text{ times}$$

The light emitted is proportional to the kinetic energy (KE) of the electron while it strikes the screen, which is equal to $\sqrt{V}$ times, where V is the velocity. While striking,

$$\text{Boost in velocity} = \sqrt{25} = 5 \text{ times}$$

The velocity acquired in the electron gun, because of accelerating voltage, is proportional to V_{acc} and hence the boost required in the accelerating voltage is:

Case (i)

$$\text{Time} = \frac{1}{100} \text{sec}$$

$$\text{Light intensity} = 1 \text{ CP}$$
$$\text{Resistance} = 10^{-4} \text{ sec}$$

Therefore, the product of CP and time, taking into account the persistence of the screen is

$$\frac{1 \times (1/100)}{10^{-4}} = 100 \text{ CP}$$

For good photography, 100 CP light is required.

However, for the normal setting of a CRO (if the intensity is known), the intensity that we can obtain is 4 CP. Therefore, this should be: 100/4 = 25 times of intensity boost is required. Therefore, the light emitted is the KE of electrons while it hits the screen straight, which is equal to $\frac{1}{2}$ mV^2 (where V is the velocity while striking), boost in velocity = $\sqrt{25}$ = 5 times. Therefore, the accelerating voltage of the accelerating anodes should be increased by 5 times.

Example 5.2

A bullet passes through two magnetic coils D distance apart at time intervals of t_1 and t_2. How do you determine the velocity of the bullet using an electronic counter?

Solution

When the bullet passes though the first coil, a pulse is generated. This is applied to an AND gate. The other input to the AND gate is from a crystal oscillator. When the first enable pulse from the first coil is applied, the counter starts counting. When the bullet passes through the second coil, another pulse is generated, which is used to stop counting; this is the disable pulse. Therefore, the number of counts accelerated during this interval is a measure of the time taken by the bullet to traverse a distance d between the two coils. Therefore, velocity = d/t.

5.5 STORAGE OSCILLOSCOPES

Storage CRTs continue to display a waveform after the input signal ceases. The period of image retention runs from a few seconds to several hours. Three common kinds of display measurements, very difficult to view on a standard oscilloscope CRT, are easily displayed on a storage oscilloscope tube.

Single-shot events (such as the waveform of an explosion) are transient in nature and are quickly lost to the observer unless the waveform can either be photographed or stored on the CRT itself.

Signals with very low repetition rates are also difficult to view on standard oscilloscopes because of phosphor decay. This is most annoying in medical displays, mechanical motions, and radar displays—three prominent areas of application for CRT indicators. A comparison of two waveforms is often difficult on a standard oscilloscope as CRT "refresh characteristic" requires simultaneous testing of the two circuits.

In contrast, a storage oscilloscope being able to store a signal trace permits comparison with a second signal obtained at some other point.

The storage oscilloscope amply justifies its additional cost and complexity for many users because of these important applications.

Storage targets can be distinguished from standard phosphor targets because they have the ability to store or retain waveform patterns for some time, independent of phosphor persistence.

The storage CRO uses a special CRT called the storage tube. This special CRT contains not only all the elements of a conventional CRT such as the electron gun, deflection plates, and a phosphor screen but also a number of special electrodes.

Two storage techniques are used in oscilloscope CRTs:

1. Mesh storage.
2. Phosphor storage.

5.5.1 Mesh Storage

A mesh-storage CRT contains dielectric material deposited on a storage mesh, a collector mesh, a flood gun, and a collimator, in addition to all the elements of a standard CRT (Fig. 5.9).

The storage target, a thin deposition of a dielectric material such as magnesium fluoride on the storage mesh, makes use of a property known as secondary emission.

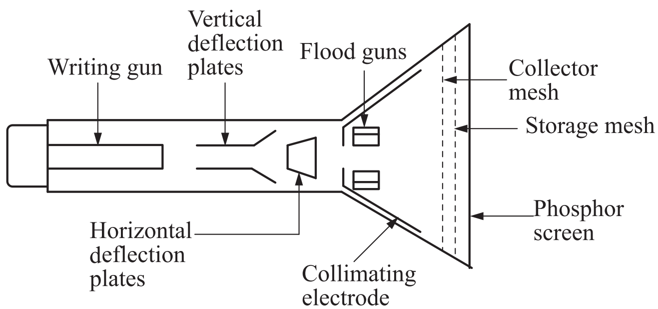


Figure 5.9 Mesh-storage CRT structure

This characteristic, illustrated in Fig. 5.10, is common to most materials when bombarded by electrons of sufficient energy. Between the first and the second crossover, more electrons are emitted than that are absorbed by the material, and a net positive charge results. Below the first crossover, a net negative charge results, as impinging electrons do not have sufficient energy to force an equal number of electrons to be emitted. In the presence of a flood gun biased at ground and a collector mesh biased at V_c (Volts), two stable points of operation may be found for the storage surface (Fig. 5.11). Note that these points bracket the first crossover region, which is an unstable point.

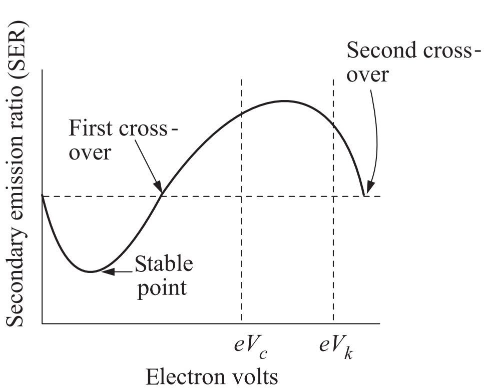


Figure 5.10 Secondary emission

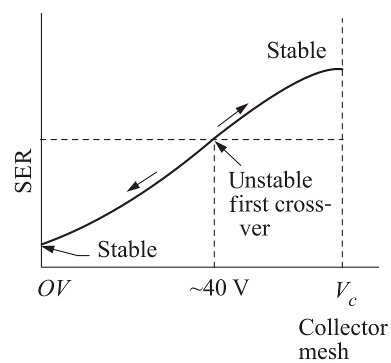


Figure 5.11 Storage-mesh potential

In order to store a trace, assume that the storage surface is uniformly charged to the lower stable point, and the write gun (the beam-emission gun of the standard CRT) is biased well beyond the first crossover with respect to the storage-mesh potential. Thus, writing-beam electrons will hit the storage target with energy eV as in Fig. 5.12, where the secondary emission ratio is much greater than unity. Those areas of the storage surface hit by the deflecting beam will lose electrons, which are collected either by the collector mesh or the display phosphor target. Thus, the write-beam deflection pattern is traced on the storage surface as a positive charge pattern. Since the insulation of the dielectric material is adequate to prevent charge migration for a considerable length of time, the pattern is effectively stored. Secondary emission will take place from the storage mesh. The pattern is stored as a positive charge.

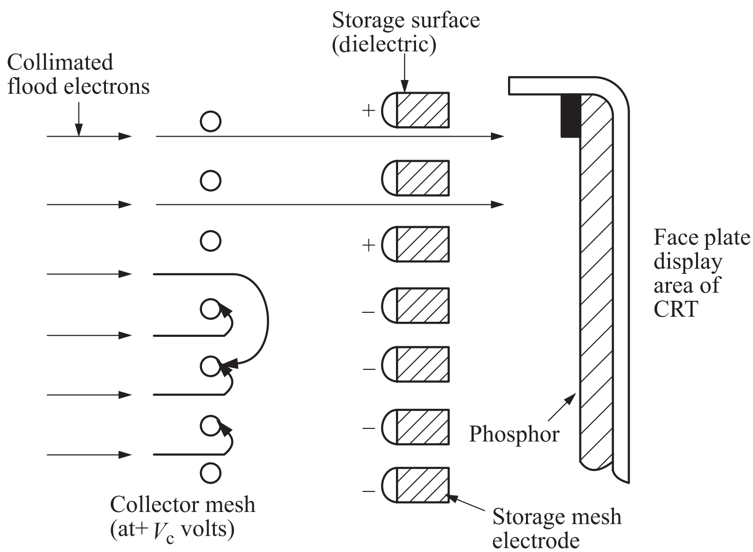


Figure 5.12 Display of the stored-charge pattern on a mesh-storage CRT

To view the stored trace, a flood gun is used when the write gun is turned off. The flood gun, biased very near to the storage-mesh potential, emits a great flood of electrons, which migrate towards the collector mesh since it is biased slightly more positive than the deflection region. The collimator, a conductive coating on the CRT envelope with an applied potential, helps to align the flood electrons so that they approach the storage-target perpendicularly. When the electrons penetrate beyond the collector mesh, they encounter either a positively charged region on the storage surface or the negatively charged regions where one trace has been stored. The positively charged areas allow the electrons to pass through to the post-accelerator region and the display targets phosphor. The negatively charged regions repel the flood electrons back to the collector mesh. Thus, the charge pattern on the storage surface appears reproduced on the CRT display phosphor just as though it were traced with a deflected beam.

The stored pattern eventually degrades, primarily because ion-generated flood-gun electrons charge other regions of the storage surface and the entire display consequently appears to be written. This is called fading positive. A typical mesh-storage CRT will store a trace for an hour or more, and the trace may be displayed at bright intensity for at least a minute.

5.5.2 Variable Persistence

The mesh-storage technique is capable of using a complex erasure pulse to vary the effective persistence seen on the screen. This is really a storage mode with continuous electrical control of the duration of storage. A phosphor-storage tube is capable of erasure electronically at the end of every sweep, but continuous persistence control is denied to it because of the bistable nature.

Variable persistence allows continuous control of the persistence from about 200 μsec to several minutes and thus allows the trace persistence to be set for a small trace-decay tail to indicate the past history of a slow-moving dot (radar use) to show five or more sequential traces (to observe medical anomalies) or to have the previous trace fade just as a new one replaces it.

5.5.3 Phosphor Storage

Although work on storage targets with secondary emission properties has been conducted since the start of the century, a storage-target CRT design based on the dielectric mesh target was first constructed in 1947 by Dr. Andrew Haeff. Some years later, Robert Anderson developed the *bistable phosphor storage* by using similar principles of secondary emission. These two approaches nearly comprise the commercial storage CRT construction today. Significant improvements in both types are regularly forthcoming.

In the bistable storage tube, the same material is used for both the storage target and the display. The material used is a P_1 phosphor doped for good secondary emission characteristics. A necessary condition for such a target is that boundary migration of stored charge must be eliminated; scattered phosphor particles achieve this condition provided that the deposition is shallow enough so that the surface is not electrically continuous. The layer may be more than one particle thick, which allows a viewing phosphor of continuous nature, but there is a thickness threshold beyond which no storage is possible.

The controlling electrode for a bistable phosphor is the conductive back plate, a transparent metal film deposited on the inside surface of the face plate before the phosphor is deposited. A wide range of operating voltage on their electrode (about 100–200 V) gives stable storage characteristics. Voltage below 100 V will uniformly erase the target and voltage above 200 V will uniformly write the target.

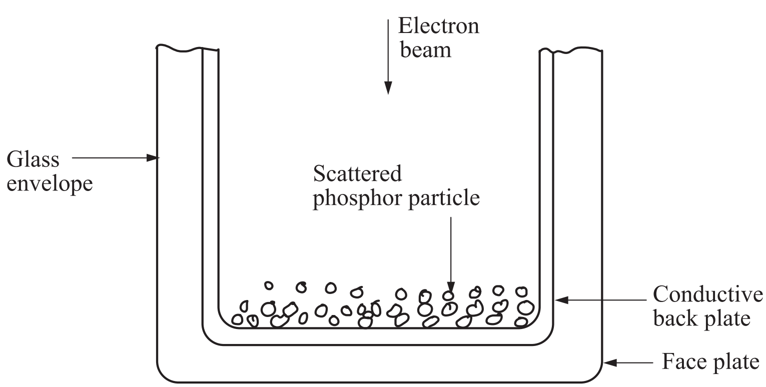


Figure 5.13 Target structure of a phosphor-storage CRT

The bistable nature of the storage on the phosphor means that a trace is either stored or it is not, and brightness is thus *on* or *off*. Neither halftones nor variable persistence is possible, although simulated effects may be produced. The bistable tube has been manufactured in a split-screen version by depositing two independent conductive back plates, one covering the top half of the RT viewing area and the other covering the bottom half. Thus, by operating the top back plate at 150 V for the flood guns, and the bottom half at 50 V, the CRT is a storage tube on the upper half and a standard refreshed phosphor display on the bottom half. The structure of a phosphor-storage CRT is shown in Fig. 5.13.

5.5.4 Phosphor Characteristics

Phosphor is the commercial name for the material used for CRO screens. It has the capability of converting electrical energy to light energy. These are classified as follows:

1. Short persistence decay of intensity is less than 1 μsec.
2. Medium persistence decay of intensity is less than 2 sec.
3. Long persistence decay of intensity is a few minutes.

The colour of the light depends upon the chemical compositor of the material.

P_{15}: Phosphor gives a blue-green display.
P_{19}: Gives an orange display.
P_{31}: Gives a yellow green display. This is good for the eye as our eyes are more sensitive in the yellow-green region.
P_1: Zinc silicate: Mu yellow-green CRO radar.
P_2: Zinc sulphide: Cu blue-green Mono.
P_4: Zinc sulphide: Zinc cadmium Ag white Chromium/Tungsten.
P_{31}: Zinc sulphide: Cu green CRO
P_{26}: Potassium Magnesium Flouride Mu orange radar

5.5.5 Persistence of Phosphor Materials

Based on decay time, i.e. the time taken by the visible light on the CRO screens to fade out, phosphor materials are classified as:

With decay time of 1 sec or longer: very long persistence.
100 msec–1 sec: Long persistence.
1 msec–100 msec: Medium persistence.
10 μsec–1 msec: Medium short persistence.
1 μsec–10 μsec: Short persistence.
< 1 μsec: Very short persistence.

The bistable nature of the storage on the phosphor means that a trace is either stored or it is not, and brightness is thus *ON* or *OFF*. Halftones are therefore not possible, more so is variable persistence, although simulated effects may be produced. The bistable CRO tube is manufactured in a split-screen version by depositing two independent conductive backplates, one covering the top half of the CRT viewing area and the other covering the bottom half. Thus, by operating the top backplate at 150 V for the flood guns and the bottom half at 50 V, the CRT is a storage tube on the top half and a standard refreshed phosphor display on the bottom half. This arrangement is advantageous because it does not require the careful adjustment of a lumped parameter line and it occupies much less space.

5.5.6 CRO Subsystems

To understand the digital storage CROs, conventional analog CRO subsystems are explained here. The primary subsystems of a CRO are shown in Fig. 5.14.

1. Vertical deflection system (attenuator, amplifier).
2. Horizontal deflection system (including time base generator, synchronisation circuit).
3. CRT.
4. HV and LV power supplies.

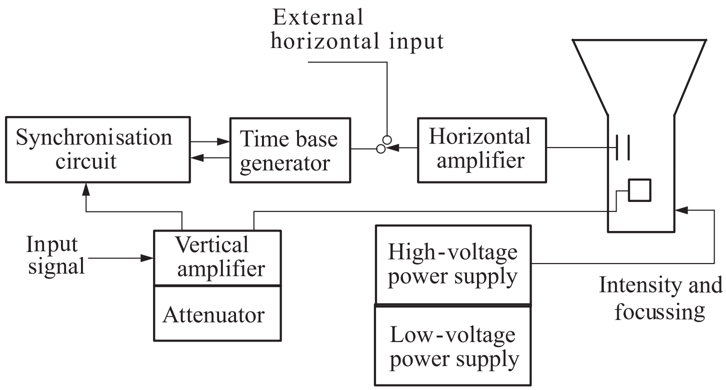


Figure 5.14 CRO block schematic

A *vertical deflection system* consists of an attenuator and a number of amplifier stages so that a wide range of input voltages are accepted.

The *horizontal deflection system* provides the device's voltage for moving the beam horizontally. Synchronisation circuits enable the sweep to be started at a specific instant with respect to the waveform under observation.

Synchronisation: To synchronise the vertical signal and the time base signal. To start both vertical and horizontal signals at the same time.

Trigger circuit: To initiate the time base by the vertical signal.

Storage CRO: In the conventional CRT, the persistence of the phosphor ranges from a few μsec to a few seconds. The storage CRO uses a special CRT called the storage tube. This special CRT has a number of special electrodes compared to the conventional CRT.

The storage mesh, mounted just behind the phosphor screen, is a conductive mesh covered with a highly negative coating of magnesium fluoride. The write gun is a high-energy electron gun similar to the conventional gun giving a narrow focussed beam, which can be deflected and used to write the information to be stored. The write gun writes a positively charged pattern on the storage target by knocking off secondary emission electrons. Because of the excellent insulating properties of MgF_2, the positively charged pattern remains exactly in the position on the storage target where it was first deposited.

The stored pattern may be made available for viewing at a later time by the use of two special electron guns called *flood guns*. These are placed between the deflection plates and the storage target. They emit low-velocity electrons over a large area towards the entire screen. The electron trajectories are adjusted by the collimating electrodes that constitute a low-voltage electrostatic lens system, so that flood electrons cover the entire screen area. Most of the flood electrons are collected by the collector mesh and therefore never reach the positive charge on the storage target. The positive field pulls some of the flood electrons through the stored mesh and these electrons continue to hit the phosphor. The CRT display (Fig. 5.15) will therefore be an exact copy of the pattern, which was initially stored on the target and the display will be visible as long as the flood guns continue an emission of low-energy electrons. To erase the pattern that is etched on the storage target, neutralisation of the stored positive charge is done.

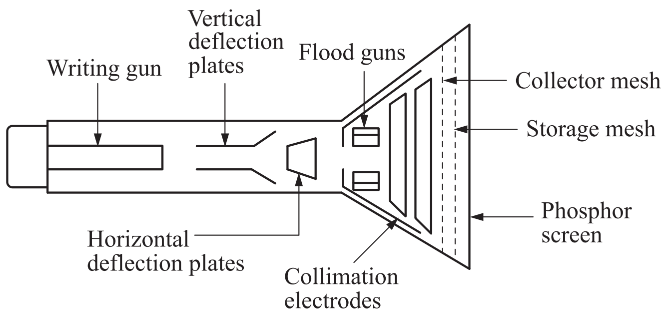


Figure 5.15 A CRT

While reproducing the image, when the electrons from the flood guns hit the storage mesh, the positive charge will not be neutralised by the electrons because the screen is at a positive potential, so the electrons will be accelerated by the positive potential and the positive charge and they continue to reach the screen.

5.6 DIGITAL STORAGE CRO

A digital storage oscilloscope uses an ADC (A/D converter) to digitise the input waveform. That means, it will sample the waveform at many points and then converts the instantaneous amplitude at each point to a binary number value proportional to the amplitude. These binary numbers are then stored in the memory. A D/A converter at the output of the memory circuit reconverts the binary words to an analog voltage capable of driving the CRT vertical deflection system. The memory is scanned many times within a second; hence, the CRT screen is constantly refreshed by the data stored in memory before it can fade out (Fig. 5.16).

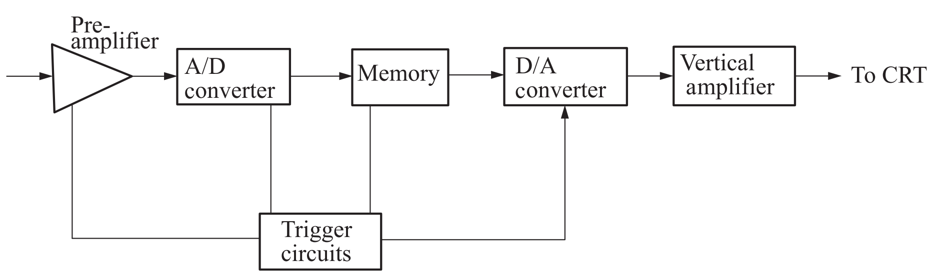


Figure 5.16 Block schematic of a digital storage CRO

The digital storage technique finds extensive use in medical displays.

A detailed block schematic is shown in Fig. 5.17. The input is amplified and attenuated as in any other CRO. The input circuit and CRO probes are the same as for any other conventional CRO. Digital CROs may also be used as normal CROs by passing the storage facility. The output of the input signal amplifier feeds an analog to digital conversion A/D converter (ADC). Thus, an ADC must be of a high-speed type since the input signals can be of a high frequency. If the CRO is to display 100 μsec for a free trace for one division, a time of 10 μsec will be required assuming that there are 10 divisions on the CRO screen. For an 8-bit A/D converter, 256 conversions would be required for 100 μsec of time. For conversion, the time is (100 μsec/256) = 390 nsec. For an 8-bit A/D converter, 8(N) or 8 + 1 = 9 (N + 1) clocks are required for each conversion and so the clock must be of 21 or 23 MHz.

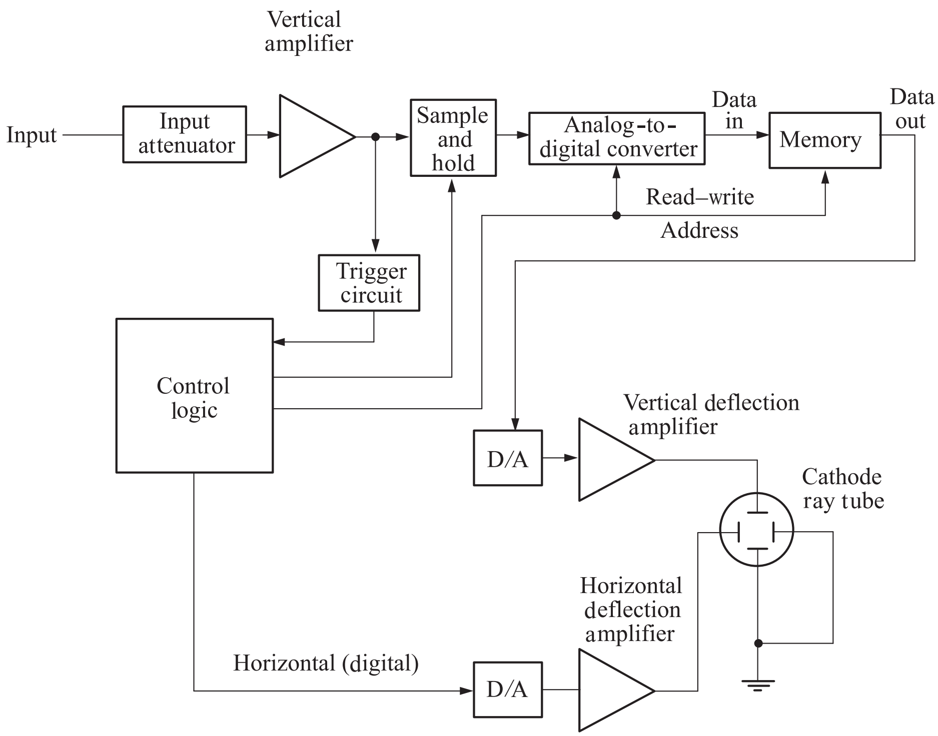


Figure 5.17 Block diagram of a digital CRO

For very fast digital storage oscilloscopes, particularly for strong one-time transient events, a method of analog storage is often used. The block schematic for the same is shown in Fig. 5.18. The input waveform is stored in an analog shaft register. The principle is storing a charge on a capacitor, where the charge is proportional to the input voltage. The charge is transferred to another capacitor storage element so that the first capacitor is free to accept a new charge. When the charge is transferred, the analog signal is transferred from one capacitor to the other. A buffer amplifier is used in between the capacitors to prevent loss of charge. The output of an A/D converter is stored in a RAM for retrieval or display. For display, each bit stored in the RAM is taken, given to a D/A converter, and is displayed as an analog deflection on the oscilloscope trace. A display rate of once per 10 msec is sufficient to present a flicker. This means that 256 samples could be retrieved and displayed every 39 μsec.

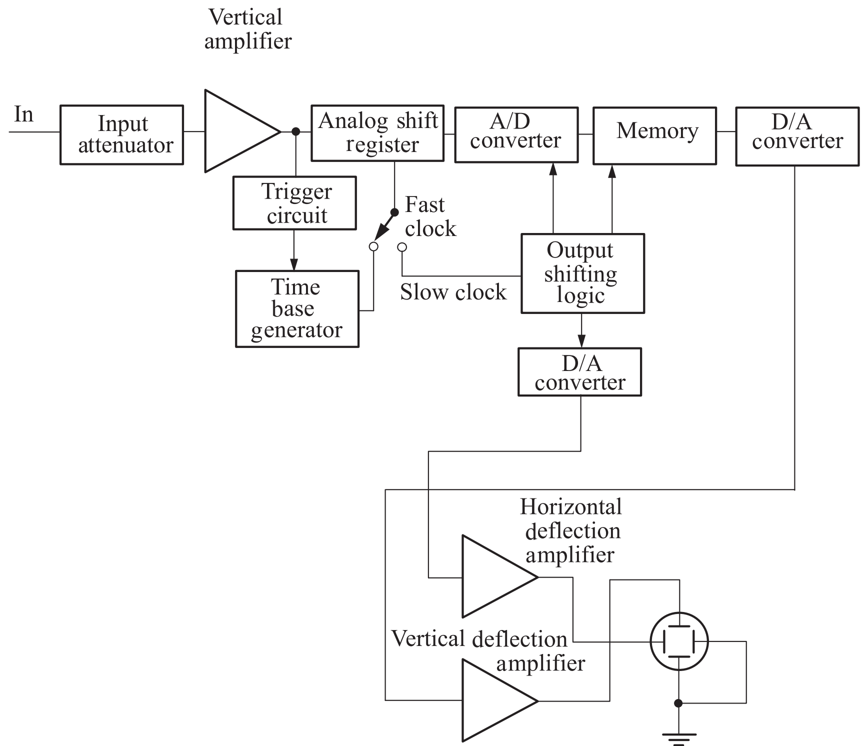


Figure 5.18 Digital storage oscilloscope for transient events

5.6.1 CRO Probes

A typical vertical amplifier presents an input impedance of 1 MΩ shunted by a capacitance of 12 pF to 47 pF. This has to be taken into account to make an accurate measurement without the loading effect. Figure 5.19 shows the equivalent circuit of a signal source connected to the input terminals of the vertical amplifier. If the source impedance Z_s is high, the small current drawn by the CRO input circuit may cause

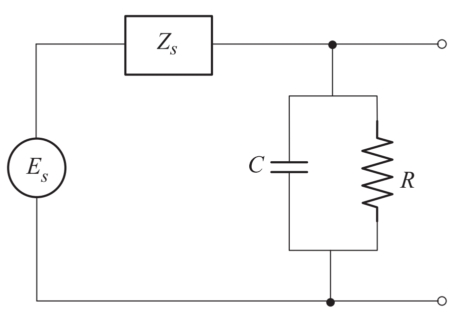


Figure 5.19 Equivalent circuit of a signal source and a CRO

a voltage drop across the source impedance itself. Therefore, the CRT display is not a faithful reproduction of the original signal amplitude. At high frequencies, the shunt capacitance plays a role in the reduction of the signal amplitude and the phase shift. The cable being used to connect the signal to the CRO also acts as a combination of R and C and can affect the signal reproduction on the CRO screen. Hence, different types of CRO probes are designed for different applications.

5.7 FREQUENCY/PERIOD-TIMER/COUNTER CIRCUIT

Countable signals must be in a form suitable for digital operation. They must have a fast rise time and should be free of noise. These factors are taken care of in the input wave shape by the Schmitt Trigger. To eliminate flicker in the display and for a constantly charging display value, there is a sample and hold circuit. The display may be a neon tube, an incandescent lamp, or a light-emitting diode. The signal can be differentiated and then clipped and inverted to form a pulse at its leading edge. If the reset pulse is blocked, the count in the electronic counter will be held indefinitely. This is the *hold function*. A field-effect transistor in the input circuit provides high-input Z for calibration purpose. The series trimmer capacitor of the crystal oscillator is adjusted to the frequency displayed with the known frequency of the standard input signal for calibration purpose.

Example 5.3

A 10 kHz signal modulated with 200 kHz is applied to an oscillograph having an input resistance of 10 Ω. A double L-section filter is to be used to limit the lowest frequencies to 1.3%. An impedance ratio of 10:1 is acceptable as sufficiently good to avoid an interaction effect between the section of filter. Determine the RC values of each section.

Solution

In a low-pass L-filter section with R and C, we have

$$\frac{e_2}{e_1}(j\omega) = \frac{1}{j\omega R_e + 1}$$

To achieve 10:1 attenuation, R_f in the second section is 10 times more than in the first section. The RC value should be the same (Fig. 5.20). Therefore C_f in the second section is one-tenth of that in the first section. For a double-section filter

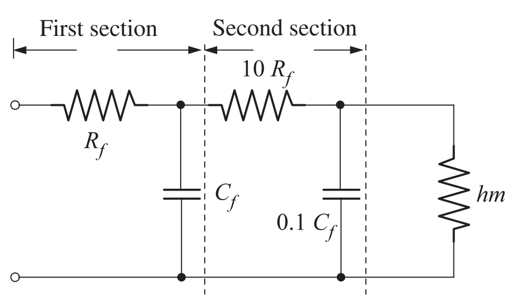


Figure 5.20 For Example 5.3

$$\frac{e_2}{e_1}(j\omega) = \frac{1}{j\omega R_f C_f + 1} \times \frac{1}{j\omega(10R_f)\,(0.1C_f) + 1} = \frac{1}{\sqrt{1+(\omega R_f C_f)^2}}$$

The lowest frequency in a modulator of 10 kHz with 200 kHz is

$$200 \text{ kHz} \pm 10 \text{ kHz} = 190 \text{ kHz}$$

$$\therefore \frac{1.3}{100} = \frac{1}{\sqrt{(2\pi \times 19{,}000 R_f C_f)^2 + 1}} \quad (R_f C_f = 0.00074 \text{ s})$$

$$\therefore \text{Run is } 10^7\ \Omega,\ 10R_f = \frac{10^7}{10} \quad (\because 10{:}1 \text{ attenuation is to be true})$$

$$R_f = 10^5\ \Omega$$

$$C_f = \frac{0.00074}{10^5} \times 10^6\ \mu\text{F} = 0.007\ \mu\text{F}$$

Example 5.4

Voltage V_1 is applied to the vertical input and voltage V_2 to the horizontal input of a CRO. The Lissajous pattern is symmetrical about the vertical and horizontal axes, with V_1 and V_2 having the same frequency. The slope of the major axis is positive with a maximum value of 2.5 div. The point where the figure crosses the vertical axis is 1.2 div high. Calculate the phase shift between V_2 and V_1.

Solution

$$\theta = \frac{1.2}{2.5} = 0.48 = \frac{Y_1}{Y_2}$$

$$\theta = 28.7° \quad \text{or} \quad 331.3 \quad (\because\ 360 - \theta)$$

Since the ellipse is lying in the I and the III quadrants, the angle is 0° to 90° or 360° –θ (Fig. 5.21). If the ellipse is in the II and the IV quadrants, the angle is 180°–θ.

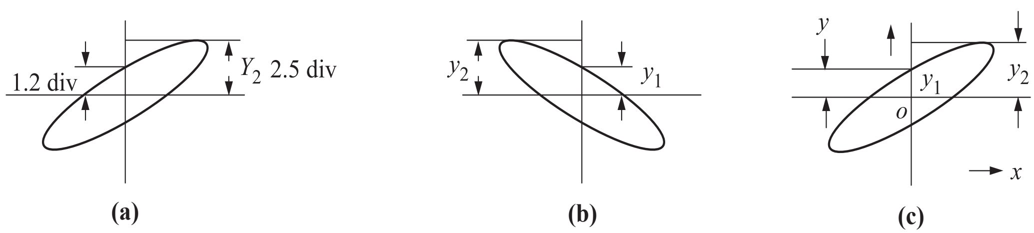


Figure 5.21 For Example 5.4

When

$$y = A \sin(\omega t + \theta)$$

$$x = A \sin \omega t$$

$$x = 0$$

when

$$\omega t = 0$$

Then

$$y = A \sin \theta$$

Let the value of y be Y_1. y is maximum when $(\omega \ + \theta) = \pi/2$ and it is equal to $A{:}Y = A$. Let the value of y be Y_2:

$$\therefore \quad y_1 = A \sin \theta \quad \text{and} \quad A = Y_2$$

$$\therefore \quad y_1 = Y_2 \sin \theta \quad \text{or} \quad \sin \theta = \frac{Y_1}{Y_2}$$

Example 5.5

In determining the loss angle of a condenser by a CRO, one-two hundredth of a condenser voltage is applied to X-plates. Y-plates are impressed by voltage 100,000 times the magnitude of a condenser current but lagging $\pi/2$ behind I. The circuit is set to give a line pattern if C is less free and the oscillograph sensitivity is 0.6 V per mm on both axes. The elliptic pattern has an area of 2 cm^2, the x-deflection is ± 4 cm and the y-deflection ± 3 cm. Determine the power loss for a capacitor.

Solution

If I_c leads V_c by 90°, C will be a lossless ideal capacitor, and it will have infinite resistance R. Therefore, I_c is leading V_c by < 90°. θ is the loss angle of capacitor (Fig. 5.22).

$$\text{Power factor (pf)} = \cos \theta = 1 \quad \text{when} \quad \theta = 0$$

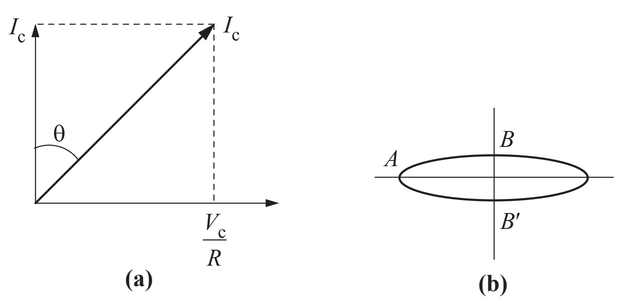


Figure 5.22 For Example 5.5: (a) Phasor digram and (b) Lissajous figure

$$\text{Condenser voltage} = \frac{V_{max}}{\sqrt{2}}$$

$$= \frac{1}{\sqrt{2}}(\text{CRO sensitivity}) \times (x\text{-deflection})$$

Therefore, a condenser voltage is applied to the x-axis. The deflection sensitivity = 0.6 V/mm = 0.6 × 10 V/cm. One-two thousandth of C voltage is applied to the x-plates.

$$\therefore \text{ Condenser voltage} = \frac{1}{\sqrt{2}} \times 0.6 \times 10 \times 4 \times 200 = 4800\sqrt{2} \text{ V}$$

$$\text{Condenser current} = \frac{1}{\sqrt{2}} \times 0.6 \times 10 \times (3) \times \frac{1}{100{,}000} \text{A} = \frac{18}{\sqrt{2} \times 10^5} \text{ A}$$

$$\text{Volt amperes} = \frac{4800}{\sqrt{2}} \times \frac{18}{\sqrt{2} \times 10^5} = 0.432 \text{ VA}$$

If p.f. = 1, the ellipse could have a major axis of 8 cm (x-axis) and a minimum axis of 6 cm (y-axis).

$$\text{With area} = \sqrt{1}(ab) = \frac{\pi}{4} \times (2a)(2b) = \frac{\pi}{4} \times 8 \times 6 = 12\pi$$

the corresponding power is 0.432 W.

$$\text{Power loss} \times \text{Total angle} = 0.432 \times \frac{2}{12\pi} = 0.023 \text{ W}$$

Example 5.6

A Lissajous pattern on a CRO is stationary and has six vertical maximum values and five horizontal maximum values. The frequency of the horizontal input is 1500 Hz. Determine the frequency of the vertical input.

Solution

$$\frac{f_y}{f_x} = \frac{\text{No. of points the target meets per bottom } (x\text{-axis})}{\text{No. of points the target meets per bottom } (y\text{-axis})} = \frac{6}{5}$$

$$\therefore\ f_y = \frac{6}{5} f_x = \frac{6}{5} \times 1500 = 1800 \text{ Hz}$$

Example 5.7

A voltage V_1 is applied to X-plates and V_2 is applied to Y-plates. V_1 and V_2 have the same frequency. Trace is a ellipse. The slope of major axis is positive. The maximum vertical value is 2.5 divisions and the point where the ellipse crosses the vertical axis is 1.25 divisions. The ellipse is symmetrical about the horizontal and vertical axes. Determine the possible phase angles of V_2 with respect to V_1.

Solution

$$\sin\theta = \frac{b_1}{b_2} = \frac{a_1}{a_2}$$

$$\frac{b_1}{b_2} = \frac{1.25}{2.5} = \frac{1}{2};\ \theta = \sin^{-1} = 30°$$

Therefore, the possible phase angles are 30° and 330° (Fig. 5.23).

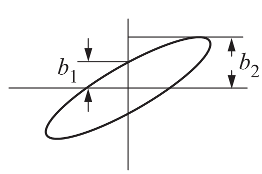


Figure 5.23 For Example 5.7

Example 5.8

In a particular CRO, the anode voltage is 2000 V. The voltage across the deflecting plates is 100 V. The axial length of the deflecting plate is 5 cm and the distance between then is 1.5 cm. The screen is located at 30 cm away from the edge of the deflecting plates. Calculate the velocity of the electrons when they enter the elective field of the deflecting plates and the deflection produced on the screen.

Solution

$$\frac{e}{m} = 1.76 \times 10^{11} \text{ c/kg}$$

$$L = 30 + 2.5 \text{ cm} = 32.5 \times 10^{-2} \text{ m}$$

$$a = 1.5 \text{ cm}$$

$$l = 5 \text{ cm} = 5 \times 10^{-2}; \quad V_o = 2000 \text{ V}$$

$$V_d = 100 \text{ V}; \quad \frac{e}{m} = 1.76 \times 10^{21}$$

$$\text{Deflection} = \frac{lLV_d}{2dV_o}$$

$$= \frac{162.5}{6.0} \times 10^{-2} \text{ m} = 2708 \text{ cm}$$

$$\text{Velocity } V_o = \sqrt{\frac{2eV_o}{m}}$$

$$= \sqrt{2 \times 200 \times 1.76 \times 10^{11}}$$

$$= 2 \times 10^2 \times 1.296$$

$$= 2.592 \times 10^7 \text{ m/kg}$$

Example 5.9

A saw-tooth waveform is applied to an average reading diode voltmeter, calibrated in terms of the average value of a sine wave. Find the error in meter indication.

Solution

For a saw-tooth waveform, the *rms* value (Fig. 5.24)

$$V_{rms} = \sqrt{\frac{1}{T}\int_T^{2T}\left(\frac{V_m}{V_2}\right)^2 dt} = \frac{V}{T_m}$$

$$\text{Average value } V_a = \frac{1}{2\pi}\int_0^T\left(\frac{V_m}{4}\right)dt = \frac{V_m}{4}; \ V_{av} = 0.433V_{rms}$$

$$I_{av} = \frac{0.433V_{rms}}{R}$$

$$\% \text{ error in meter indication} = \frac{0.433V_{rms} - 0.45\left(\frac{V_{rms}}{R}\right)}{0.45\frac{V_{rms}}{R}} \times 100 = -3.8\% \text{ (Less)}$$

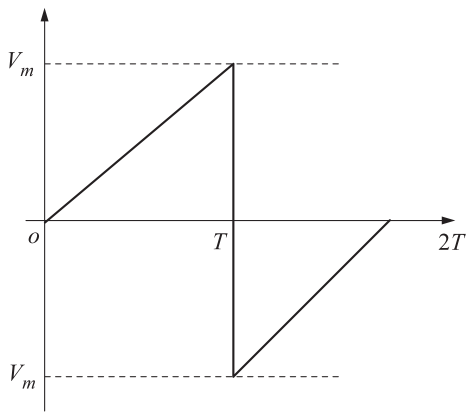


Figure 5.24 For Example 5.9

Specification of a Dual Trace Oscilloscope

Vertical Deflection	Horizontal Deflection
Two CH_1, CH_2	Time base: 0.1 msec/cm
BW: DC to 20 MHz	Accuracy: ±3%
Sensitivity = 2 mV/cm	Variable
Accuracy = ±3%	Sweep > 2.5
I/p imp = 1 m	Trigger
I/p coupling = AC/DC	Source, slope, sensitivity, coupling
Display modes	Supply — 220/240 V± 10%, 65 Hz
Single trace	Operating temp range — 0 to 50° C
Dual trace	Dimensions —140 × 305 × 650 m
Inert CH_2	Weight—6 kg
Inert CH_1	

5.8 FREQUENCY MEASUREMENT

Frequency can be accurately measured by counting the number of cycles of the unknown signal for a precisely controlled time interval. The logic block diagram for a counter in the frequency mode of operation is shown in Fig. 5.25.

When the output of the AND gate is high, it is counted as 1, irrespective of the shape. The wave shape is not important for counting.

There are two signals that need to be traced:

1. The input signal (whose frequency is to be measured).
2. The gating signal that determines the length of time during which the decimal counting assemblies (DCA) are allowed to accumulate pulses.

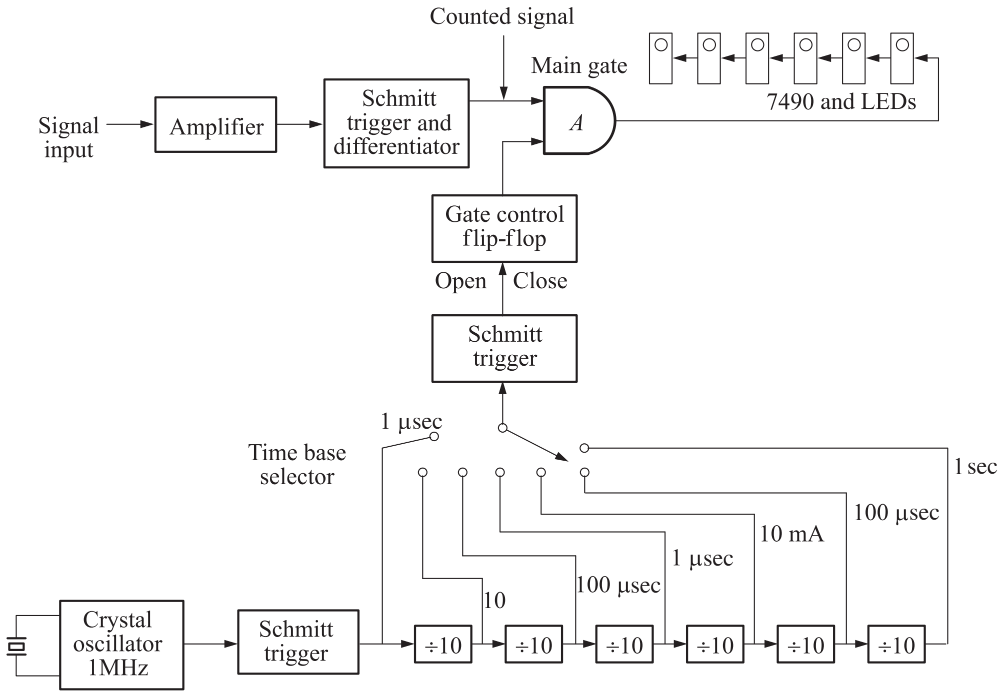


Figure 5.25 Block schematic for frequency measurement

The input signal is amplified and passed to a Schmitt trigger. Here it is converted to a square wave with very fast rise and fall times; they are then differentiated and clipped. As a result, the signal that arrives (the Schmitt trigger that acts as a signal shaper cuts to get uniform pulses. DCAs are bistable multivibrational connectors) at the input to the main gate consists of a series of pulses separated by the period of the original input signal. The internal oscillator frequency is 1 MHz. The time base output is shaped by a Schmitt trigger so that positive spikes 1 μsec apart are applied to a number of decade dividers. In Fig.5.23 six decade divider assemblies (DDA) are used, the outputs of which are connected to a time base selection switch. This allows the time interval to be selected from 1 μsec to 1 sec. The first output pulse from the time base selection switch passes through a Schmitt trigger to the gate control flip-flops. The gate flip-flop assumes a state such that an *enable signal* is applied to the main gate. Since this is an AND gate, the input signal pulses are allowed to enter the DCAs, and they are totalised and displayed. This continues until the second pulse from the DCAs arrives at the control flip-flop. The gate control assumes the other state that removes the *enable signal* from the main gate. The main gate closes and no further pulses are admitted to the DCAs. The display of DCA is now in a state that corresponds to the number of input pulses received during a precise time interval, which was determined by the time base. Usually, the time base selection switch moves the decimal point in the display area allowing the frequency to be read directly in Hz, kHz, or MHz.

5.9 PERIOD MEASUREMENT

The gating signal is derived from the unknown input signal, which now controls the opening and closing of the main gate. The precisely spaced pulses from the crystal oscillator are counted for one period of the unknown frequency. The time base of 1 MHz is set to 10 μsec (100 kHz) and the number of pulses that occur during one period of the unknown signal is counted and displayed by the DCAs.

The accuracy of the period measurement is increased by using the multiple period average mode of operation. The main gate is held open for more than one period of the unknown signal.

In Fig. 5.26, the multiple period average mode of operation is shown. The 1 MHz crystal frequency is divided by 1 DDA to a frequency of 100 kHz (10 μsec period). These clock pulses are shaped by the frequency trigger and are fed to the main gate to be counted. The input signal whose period is to be measured is amplified, shaped by the period trigger (Schmitt trigger), and applied to the gate control flip-flop. The gate control provides the *enable pulse* and the *stop pulse* for the main gate. The main gate will open for a longer period (10^5 times greater). The readout logic is so designed that the decimal point will be automatically positioned to display the proper units.

DDAs work on the same principle as DCAs. They are multivibrator circuits, wherein the time period of the input pulse is changed (see Fig. 5.27; the output of a Schmitt trigger is a square wave).

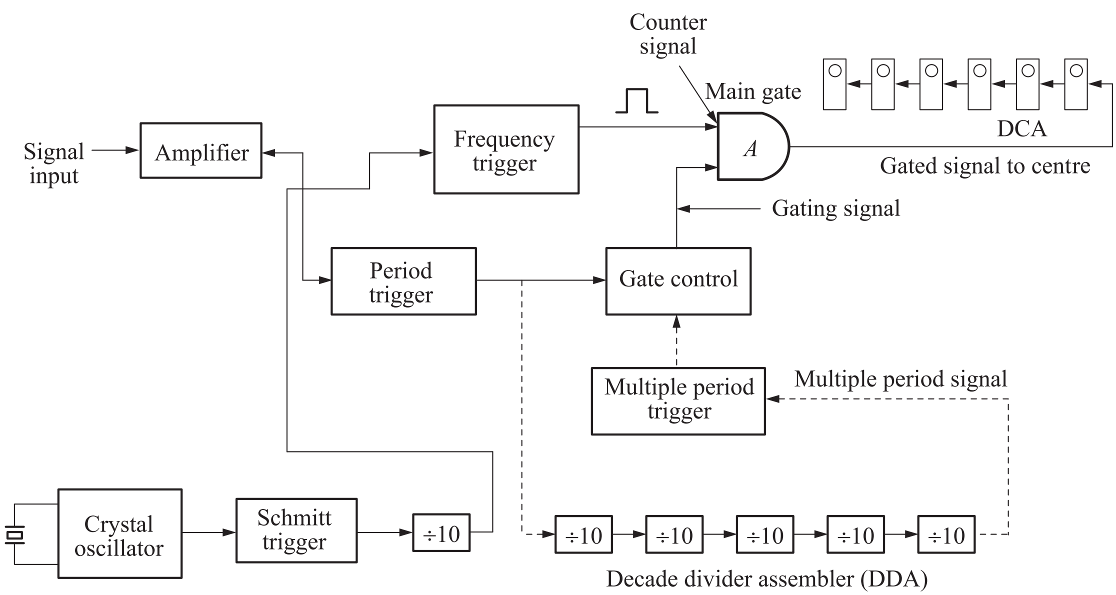


Figure 5.26 Block schematic for period measurement

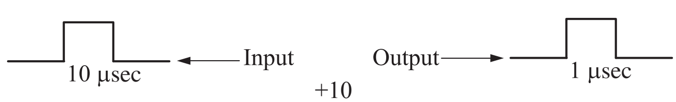


Figure 5.27 Division by 10

Crystal oscillators: When certain solid materials are deformed, they generate an electric charge within them. This effect is reversible in that, if a charge is applied, the material gets mechanically deformed in response. This is called the *piezoelectric effect*. The materials that produce this effect are

1. Quartz.
2. Rochelle salt.

3. Lithium sulphate.
4. Ammonium dihydrogen phosphate (ADP).

If the crystal is properly made, deformation takes place within the crystals and an electromechanical system is formed, which will vibrate when property excited. The resultant flow will depend upon crystal dimensions.

The dimensions of a 90 kHz crystal are 30 × 4 × 1.5 cm.

5.9.1 Advantages

1. No parallax error.
2. High-frequency range is 10 MHz.
3. Resolution is 1 Hz, error ± 1 Hz.
4. Can be used as a counter.

5.10 ERRORS IN FREQUENCY/PERIOD MEASUREMENTS

Frequency and time measurements made by electronic counters suffer from inherent errors.

The main error is because of the gating of the AND gate, which is done independently by the clock pulses and input pulses. Hence, a gating error occurs. The gating time is not synchronised with the input signal when this error occurs.

As shown in Fig. 5.28 in Case (1), four pulses will be counted. In Case (2), in the same period, three pulses will be counted. Therefore, the count will be ± 1 and there will be ambiguity. If a 10 Hz flow is being measured, the ambiguity is ± 10% because of ± 1 count difference.

In such a case, period measurement is to be made (Fig. 5.28).

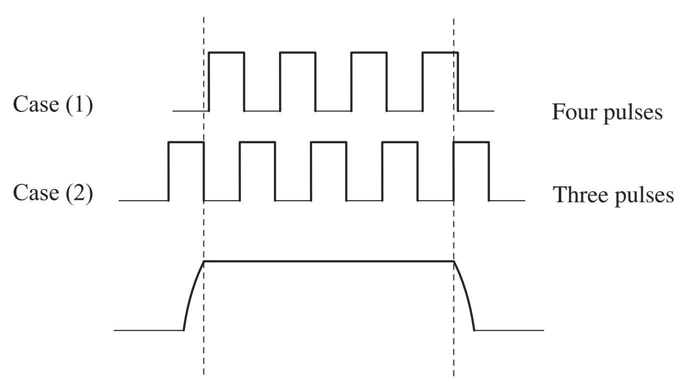


Figure 5.28 Waveforms

If

$$f_c = \text{crystal (or clock) frequency}$$
$$f_x = \text{frequency of the unknown input signal}$$

In period measurement, the number of pulse counted

$$N_p = \frac{f_c}{f_x}$$

In frequency measurement with 1 sec of gate time, the number of pulses counted is $N_f = f_x$.

The crossover frequency (f_o) at which $N_p = N_f$ is $\frac{f_c}{f_x} = f_x$; $f_x = f_o \frac{f_c}{f_o} = f_o = \sqrt{f_c}$.

Signals with a frequency *lower than* f_o should therefore be measured in the period mode. The signal of frequencies above f_o should be measured in the frequency mode in order to minimise the effect of ± 1 count gate error. As $f_x < f_o$, N_p is more. We will get more counts in period measurement since $N_p = f_c/f_x$ and therefore the measurement is accurate. In frequency measurement, the gating signal is supplied by α^{tal}, and the input pulse is measured as shown in Fig 5.29.

In period measurement, the gating signal is provided by the input signal itself. Pulses are counted for one period of the input waveform.

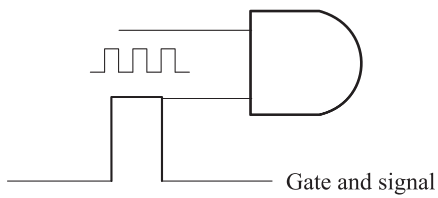


Figure 5.29 Gating signal

5.10.1 Errors Because of Crystal Stability

1. Short-term crystal stability.
2. Long-term stability.

Short term: Errors occur because of momentary frequency variations due to voltage transients, shock and vibration, ageing of crystal oven, electrical interference, etc.

Long term: Stability errors are because of aging. The stress undergone by crystal in manufacture is received. Any particles sticking to the crystal may be relieved. Hence, this may result in an increase in crystal frequency f_c.

However, accumulation of dust and cooling of crystal if it is not kept in an oven, will cause changes in the capacitance of crystal and, therefore, frequency f_c changes.

1. Frequency measurements above $\sqrt{f_c}$ and period measurements below $\sqrt{f_c}$ should be made.
2. Calibration of the counter should be made frequently and electronic counters are used to determine delay times of the relay, as shown in Fig 5.26.

The relay functions control the opening and closing of the signal gate and number of cycles of the time base generator is counted by DCAs. The various response times are measured as follows.

Delay time: The gate is opened by the coil voltage upon its application. The gate is closed by the normally closed contacts when they open.

Transfer time: The gate is opened by the normally closed contacts when they open.

Pick-up time: The gate is opened by the application of the coil voltage.

Drop-out time: The gate is opened by the removing of the coil voltage.

Totalisers: They are electronic counters that indicate the total decimal number in a given period of time.

Scalers: They are also totalisers with scaling facility to indicate in tens or dozens or hundreds, etc.

5.11 UNIVERSAL COUNTERS

In this type of counter, both period and time can be measured. Period measurement for relays, block schematic.

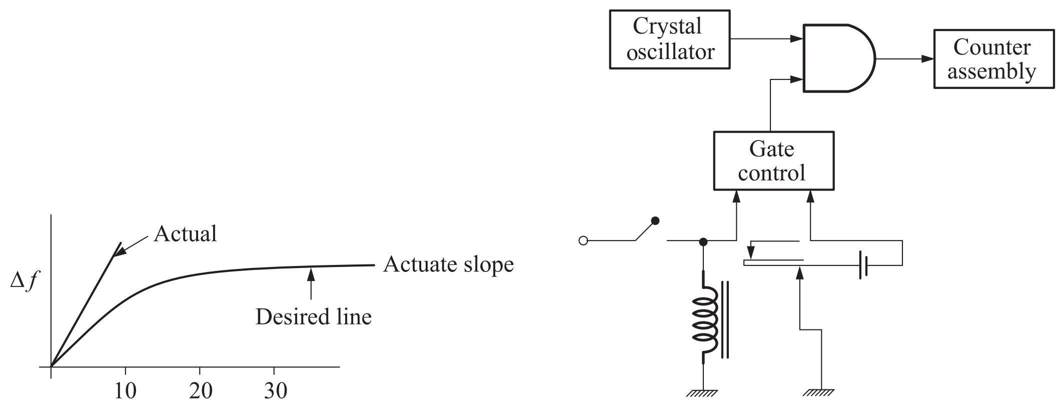


Figure 5.30 Period measurement for relays, schematic

5.12 EXTENDING THE RANGE OF FREQUENCY COUNTERS

Prescalers

Prescalers are used to enter the range of frequency counters. Suppose you have 10 MHz frequency counter, but the frequency upto 100 MHz is to be measured. How do you use the same instrument? A prescaler or frequency divider is used before the frequency counter. The input signal frequency is divided by 10 and then measured. (See Fig. 5.31.)

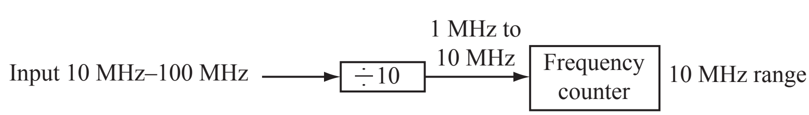


Figure 5.31 Frequency division

However, what is the drawback? The resolution of this frequency measurement deteriorates by a factor of 10. If the error of a 10 MHz frequency counter is 1 Hz, and it is used for 100 MHz, the error will be 10 Hz. By dividing the input signal frequency by 10, the number of pulses being counted in the Decimal Counting Assemblies (DCAs) is reduced and hence the error increases DCA:

$$= \frac{\sqrt{3}}{4} V_{rms} = 0.433 V_{rms}$$

Meter current for saw-tooth input voltage

$$I_{av} = 0.433 \left(\frac{V_{\text{rms}}}{R}\right)$$

$$\text{Error in meter indication} = \frac{0.433\, V_{\text{rms}} - 0.45\left(\frac{V_{\text{rms}}}{R}\right)}{0.45\frac{V_{\text{rms}}}{R}} \times 100$$

$$= 3.8\% \text{ low}$$

Example 5.10

Calculate the velocity of an electron beam in an oscilloscope if the voltage applied to its vertical deflection plates is 2000 V. Also calculate the cut-off frequency if the maximum transit time is one-fourth of a cycle. The length of the horizontal plate is 50 mm.

Solution

Velocity of the electron

$$V = \sqrt{\frac{2eE_a}{m}}$$

$$= \sqrt{\frac{2\times16\times10^{-19}}{9.1\times10^{-31}}} \times 2000$$

$$= 26.5 \times 10^6 \text{ m/s}$$

Cut-off frequency

$$f_C = \frac{V}{4l} = \frac{26.5\times10^6}{4\times50\times10^{-3}}$$

$$= 132.5 \times 10^6 \text{ Hz} = 132.5 \text{ MHz}$$

5.13 GLOSSARY

Active probe: A probe containing transistors or other active circuits that normally use an external power source.

Attenuator probe: A probe that attenuates the input signal to increase the input range of an oscilloscope.

Circuit Loading: Distortion of the circuit by the interaction of the probe and oscilloscope with the circuit being tested.

Compensation: An adjustment made to a probe that balances its capacitance with that of the oscilloscope.

Impedance: The process of impeding or restricting an AC signal flow.

Loading: The process whereby a load applied to a source draws current from the source.

Passive probe: A probe whose network equivalent consists only of resistive (*R*), inductive (*I*), or capacitive (*C*) elements.

Probe: An input device for transmitting the signal to the oscilloscope.

Probe power: Power that is supplied to the probe from some source such as the oscilloscope, a probe amplifier, or the circuit under test.

Reactance: An impedance element that reacts to an AC signal by restricting its current flow based on the signal's frequency.

Ringing: Signal oscillations that result when a circuit resonates.

Shielding: A grounded conductive sheet of material placed between a circuit and an external noise source to intercept and conduct away noise.

SNR (signal-to-noise ratio): The ratio of signal amplitude to noise amplitude, usually expressed in dB: SNR = 20 log (V signal/noise).

Source: The origination point or element of a signal voltage or current.

Source impedance: The impedance seen when looking back into a source.

5.14 THE ABC'S OF OSCILLOSCOPES

AC (alternating current): A signal in which the current and voltage vary over a period of time.

AC coupling: Useful for observing an AC signal riding on a DC signal, where the DC component is blocked but the AC signal passes through.

Acquisition: The process of sampling signals that digitises signal samples into data points and then assembles the data points into a waveform record.

A/D converter (analog-to-digital converter): An electronic component that converts a signal into discrete binary values.

Aliasing: This phenomenon can occur when an oscilloscope digitises at a sampling rate that is too low to reproduce the input signal accurately. The oscilloscope may then display the waveform at a lower frequency than that of the actual input signal.

ART (analog real-time) oscilloscope: The input signals drive a CRT input time to show a signal's behaviour over time. ARTs cannot store, analyse, or manipulate the signal data and are limited by the writing speed of the CRT and simple triggering.

Attenuation: A decrease in signal voltage during transmission.

Averaging: A signal-processing technique used by digital oscilloscopes to eliminate noise in a signal.

Bandwidth: The continuous band of frequencies that a network or circuit passes without diminishing in power more than 3 dB from the midband power.

Cursors: Markers that you can move across the oscilloscope screen. The oscilloscope displays the waveform values (expressed in volts, amps, time, frequency, VA, etc.) at the point(s) the marker(s) cross the waveform(s) or calculates values based on comparing different cursor locations.

DC (direct current): A signal with a constant voltage and current.

DC coupling: A mode that passes both AC and DC signal components to the circuit.

Digital real-time digitising: A digitising technique that samples the input signal with a sample frequency of four to five times the oscilloscope bandwidth. Combined with sin x/x interpolation, all frequency components of the input up to the bandwidth are accurately displayed.

DPO (digital phosphor oscilloscope): Digital oscilloscopes that display, store, and analyse complex signals in real time, using three dimensions of signal information—amplitude, time, and amplitude over time. Displayed points vary in intensity depending on the frequency of their acquisition and decay as if the oscilloscope had an analog CRT.

DSO (digital storage oscilloscope): Input signals are converted into digital data and displayed on screen (CRT or LCD). The digital nature of these data allows storage, analysis, and sophisticated triggering. DSOs are limited by sample rates and aliasing.

Envelope: The outline of a signal's highest and lowest points.

Equivalent-time sampling: A sampling mode in which the oscilloscope constructs a waveform by capturing a sample of information from each signal repetition. Accurate waveform reconstruction requires a repetitive signal and multiple triggering.

FFT (Fast Fourier Transform): A method of calculating the frequency components of a periodic waveform—often called harmonic analysis.

Floating measurements: Voltage measurements where the voltage reference is not earth-ground.

Frequency: The number of times a signal repeats in one second, measured in Hertz (cycles per second). The frequency equals one period.

Glitch: An intermittent error.

Ground: (1) A conducting connection by which a circuit is connected to the earth to establish and maintain a reference voltage level. (2) The voltage reference point in a circuit.

Hold-off: A specified amount of time that must elapse before the trigger circuit generates another trigger signal. Hold-off helps ensure a stable display.

Interpolation: A 'connect-the-dots' processing technique to build a waveform based on a few sampled points.

Noise: An unwanted voltage or current in an electrical circuit.

Normal trigger mode: A mode where the oscilloscope acquires a waveform provided that a specific trigger event occurs.

Nyquist rate: The minimum at which a signal can be sampled to avoid aliasing and to ensure accurate representation. It is twice the input signal frequency.

Oscilloscope: An instrument used to analyse a signal over time. 'Oscilloscope' comes from the used 'oscillate', since oscilloscopes are often used to measure oscillating voltages.

Peak detect: An acquisition mode that captures spikes and glitches that may occur between sample points.

Period: The amount of time it takes a signal to complete one cycle. The period equals one frequency.

Persistence: The natural decay of waveform points on a CRT.

Phase: A means of expressing (in degrees) the time-related positions of different signals.

Terminology of CROs

Pre-trigger: A specified portion of the waveform load that contains data acquired before the trigger event.

Pulse: A common waveform shape that has a fast rising edge, a width, and a fast falling edge.

Real-time bandwidth: The maximum frequency that can be acquired when sampling the entire input waveform to reconstruct the waveform accurately is the sample rate divided by 2.5.

Real-time sampling: A sampling mode in which the oscilloscope collects as many samples as the signal occurs.

Record length: The number of waveform samples at any time.

Rise time: The time taken for the leading edge of a pulse to rise from its minimum to its maximum value (typically measured from 10 to 90% of these values).

Roll mode: A display mode to view the waveform when it is acquired point by point.

Sample rate: The rate at which the analog input signal is sampled for conversion into a digital value.

Sampling: The process of capturing an analog input, such as voltage, at a discrete point in time so that it can be converted into digital data.

Single shot: Triggering of an oscilloscope to take one record length of data only.

Time base: The time base defines the time and horizontal axis of the display.

Trigger: The circuit that drives the capture of signal information. Different types of triggers can be used such as edge, video, pulse, width, logic, etc.

Trigger hold-off: A control that inhibits the trigger circuit from looking for data for some specified time after the last captured event.

Trigger level: The level that the signal must reach before the trigger circuit starts data conversion.

Voltage: The difference in electric potential expressed in volts between two points.

Waveform (trace): A graphic representation of a signal varying over time.

EY format: A display mode used for studying the phase relationship of two signals.

YT format: The conventional oscilloscope display format. It shows signal behaviour (on the vertical axis) as it varies over time (on the horizontal axis).

Specifications of Frequency Counter

Type No.	**Model No.**	
Sl. No.	Description/Parameter	Typical values
1	Frequency range	0.2 Hz to 100 MHz
2	Gate: 10 msec	100 msec, 1 sec, 10 sec, selectable
3	Resolution	100 mHz to 10 Hz
4	Accuracy	± 1 count
5	Input impedance	1 MΩ‖62 pF
Period measurement		
6	Annunciators LED indication for overflow, gate tune	0.01, 0.1, 1, 10 in secs, display hold
7	Display	0.56”, 8 digit, LED

Type No.	Model No.	
8	Display time	0.8–5 sec
9	Time base frequency	10 MHz
10	Stability	± 1 ppm
11	Temperature stability	± 5 ppm from 0 to 50°C

Courtesy: Tektronix.

5.15 SUMMARY

In dual-trace oscilloscopes, there will be one single beam CRT, a time base generator, and two identical vertical amplifiers with an electronic switch. The two modes of operations are (1) the alternate mode and (2) the chopper mode. In the alternate mode, the electronic switch alternately connects each vertical amplifier to the CRT. The CRO alternately displays the two vertical signals. The switching rate of the CRT is controlled by the sweep rate of the time base generator. In the chopped mode, the electronic switch connects vertical amplifiers irrespective of the time base frequency. The chopping rate is approximately 100 kHz. There are four different types of CRO probes commonly used, namely:

(a) Direct probe.
(b) High-impedance probe.
(c) High-voltage probe.
(d) Detector probe.

In high-frequency CROs, because of the large speed or less time duration, the intensity of the electron beams on the screen decreases. Therefore, the accelerating voltage of the beam must be increased to get reasonable brightness on the screen for high-frequency signals. High-frequency CROs consist of a series of deflecting plates. These plates are so shaped and spaced such that an electron travelling between them will get additional energy in the proper time sequence.

Trigger circuits cause delay because of inherent inductances and capacitances of the circuit. The delay will be of the order of 80–200 nsec. This affects the synchronisation of horizontal and vertical signals.

As such, the leading edge and trailing edge details of a pulse are lost. Therefore, delay lines are used. The crossover frequency f_o in frequency or period measurement is $f_o = \sqrt{f_c}$, where f_c is the crystal or clock frequency. Signals with a frequency lower than 'f_o' should be measured in the period mode. Signals of frequencies above 'f_o' should be measured in the frequency mode.

Points to Remember

- A dual trace CRO consists of a single beam CRT, single time base generator, and two identical vertical amplifiers with an electronic switch.
- The two modes of operation are:
  (a) The alternate mode.
  (b) The chopped mode.
- The different types of probes are:
  (a) Direct probe.
  (b) High-impedance probe.
  (c) High-voltage probe.
  (d) Detector probe.
- In high-frequency CROs, since the period of the signal is small, energy transferred to the phosphor screen is small. Therefore, to overcome such problems, the deflecting plates are shaped and spaced such that electrons travelling between them will gain energy from each set of plates and produce additional deflecting force in the proper time region.

- Owing to inherent delay because of inductance and capacitances associated with trigger circuits, horizontal and vertical signals will not be synchronised. Hence, the details associated within the leading edge and the trailing edge of a pulse will be lost. To overcome such problems, delay lines are used.
- The crossover frequency f_o is $\sqrt{f_c}$ where f_c is the clock frequency.
- Signals with a frequency lower than f_o should be measured in the period mode.
- Signals with a frequency above f_o should be measured in the frequency mode

Objective-type Questions

1. The two modes in which a dual-trace CRO is operated are (A) ________ (B) ________.
2. The chopping frequency in a dual-trace CRO typically is ________.
3. The input impedance of a vertical amplifier, typically is ________.
4. Different types of CRO probes are ________.
5. In high-frequency CROs, one requirement is that the bandwidth of vertical amplifiers must be ________.
6. The delay introduced by the triggering circuit is typically ________.
7. The two types of delay lines are (A)________ (B)________.
8. In a distributed parameter delay line the typical value of the time delay introduced is ________.
9. In a period measurement, the gating signal is derived from the ________.
10. The signal that determines the length of time during which the decimal counting assemblers are allowed to accumulate pulses is called ________.
11. Crossover frequency in time frequency measurement is f_C = ________.
12. Signals with a frequency less than crossover frequency f_o must be measured in ________ mode.
13. In some CROs additional focus control provided is called ________.
14. The purpose of synchronising control in a CRO is ________.
15. The sweep signal frequency of a 5 MHz CRO must be________.
16. In a high-frequency oscilloscope, intensity can be increased by increasing the ________.
17. If a CRO has a continuously running time-base generator, the type of CRO is ________.
18. The type of CRO that has two electron guns is ________.
19. The number of time-base circuits in a double beam CRO: ________.
20. The CRO used to study transients during switching of power supply is ________.
21. Both DC and AC voltages can be used using a CRO in ________ coupling.
22. The parameter that cannot be measured directly by a CRO without using any internal element is ________.

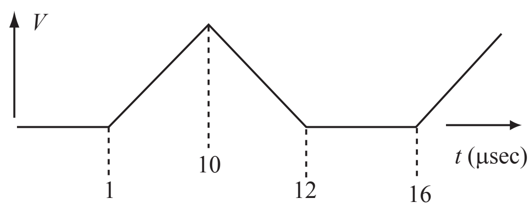
Review Questions

1. Explain the principle and working of a dual trace oscilloscope.
2. Explain about the different types of CRO probes.
3. What are the various circuit design considerations in the case of high frequency CROs?
4. With the help of an equivalent circuit, explain the functioning of Lumped parameter delay lines.
5. Giving the equivalent circuit and construction details, explain about distributed parameter delay lines.
6. With the help of a block schematic, explain the principle and working of a sampling oscilloscope.
7. Give the block schematic of frequency-measuring instruments and explain the principle of working.

8. Draw the block schematic of a period-measuring instrument and explain the principle of operation.
9. Explain about the different types of errors that occur in frequency period measurement.
10. Explain about universal counters and prescalers in counters.

Unsolved Problems

1. In a digital storage oscilloscope, the resolution required is of 4 bits in both horizontal and vertical axes. The CRO is required to display transients at the rate of 1 μsec per division for a display of 10 divisions. What should be the speed of ADC (A/D converter)? What type of ADC (A/D converter) do you suggest?
2. In a dual trace oscilloscope, the electron beam is accelerated through a potential of 3000 V. What is the velocity of the electron beam?
3. Determine the value of the deflecting voltage to be applied between the deflecting plates separated by 1.2 cm to deflect an electron beam by 1°? The length of the deflecting plates is 2.5 cm and the accelerating potential is 1500 V.
4. Determine the secondary emission ratio 'S' of a digital storage oscilloscope if the value of the secondary emission current (I_s) is 10 μA and the primary beam current (I_p) is 100 μA ($S = I_s$ I_p)
5. The wave shape of a sweep voltage is as shown in the figure. Determine the values of the sweep time, the retrace time, and the hold-off time.



6

DC and AC Bridges

Introduction • DC bridges • Wheatstone bridge • Kelvin bridge • Strain gauge bridge circuit • AC bridges • Maxwell bridge • Hay bridge • Schering bridge • Wien bridge • Anderson bridge • Resonance bridge • Similar angle bridge • Radio frequency bridge (substitution technique) • Wagner's ground connection • Twin-*T* null network • Bridged-*T* network • Detectors for AC bridges • Phasor diagrams • Recorders • Strip-chart recorders • Pen-driving mechanism • Other features • Servorecorders• Servobalancing potentiometric recorder • Characteristics of typical servorecorders • Oscillographic recorders • Magnetic tape recorders • Recorders (contd.) • Galvanometer oscillographs • Summary

6.1 INTRODUCTION

To measure parameters R, L, C, f, Q (Quality factor of a coil) and 'D' (Dissipation factor of a capacitor) of electronic circuits, bridge circuits are employed. DC bridges can measure resistance R accurately over wide ranges. AC bridges can be used to determine the unknown values of inductor L and capacitor C and even R and frequency f also. The advantage with bridge-measuring circuits is that some errors which occur in measurements due to parasitic values, temperature effects, errors due to improper grounding and shielding can be eliminated. The measurement range of the parameters is large. Even parameters like *quality factor of a coil Q*, *dissipation factor D of a capacitor* can also be measured using AC bridges in addition to frequency.

In this chapter, AC bridges—Maxwell, Anderson, Schering, Wien bridge—are explained. DC bridges—Wheatstone and Kelvin—are also explained. Errors that occur, elimination of the same, and precautions to be taken are explained. The principle and working of a Q meter is also described.

6.2 DC BRIDGES

DC bridges are used to determine the unknown conducting value or sometimes to determine the conductance associated with conducting wires. Wheatstone bridge and Kelvin double bridge are the two types in this category. AC bridges can also be used for resistance measurements, but they are used to determine inductance, capacitance, impedance, admittance, or the frequency of the AC input. Balancing AC bridges is more difficult than DC bridges, since reactive components are also involved in AC bridges. In bridge circuits, a small difference in the imbalance current can be detected by a sensitive galvanometer and hence measurement, accuracy, and precision improves. Bridge circuits are analogous to difference amplifiers, wherein a small change in the input compared to the reference signal is amplified. AC bridges are frequency dependent, wherein the variation in supply frequency in addition to the voltage can affect the measurement.

6.3 WHEATSTONE BRIDGE

6.3.1 Operation

This bridge circuit is used to determine resistance ranging from approximately an ohm (1 Ω) to several ohms (MΩ). The circuit is as shown in Fig. 6.1. The bridge circuit has four resistance arms together with a source of *emf* (it can be a battery) and a null detector. The null detector is a galvanometer or sensitive current meter. The current through the galvanometer depends on the potential difference between points C and D. The bridge is said to be balanced when the potential difference is zero. Arms AC and AD are called *ratio arms*. The value of unknown R_x depends on the ratio of the elements connected in these arms. Arm CD is called a *standard arm*. A standard precision element is connected in this arm. The unknown element is connected in the arm BD. When the bridge is at balance,

$$I_1R_1 = I_2R_2 \tag{6.1}$$

$$I_1 = I_3 = \frac{E}{R_1 + R_3} \tag{6.2}$$

and

$$I_2 = I_4 = \frac{E}{R_2 + R_4} \tag{6.3}$$

Combining equations (6.1)–(6.3) and simplifying, we obtain

$$\frac{R_1}{R_1 + R_3} = \frac{R_2}{R_2 + R_4} \tag{6.4}$$

From equation (6.4), we obtain

$$R_1\,(R_2 + R_4) = R_2\,(R_1 + R_3) \quad \text{or} \quad R_1R_2 + R_1R_4 = R_2R_1 + R_2R_3$$

Therefore

$$R_1\,R_4 = R_2\,R_3 \tag{6.5}$$

R_4 is the unknown element (R_x)

$$R_4 = R_x = \frac{R_2R_3}{R_1} = \left(\frac{R_2}{R_1}\right)R_3 \tag{6.6}$$

$$\boxed{R_4 = \left(\frac{R_2}{R_1}\right) R_3}$$

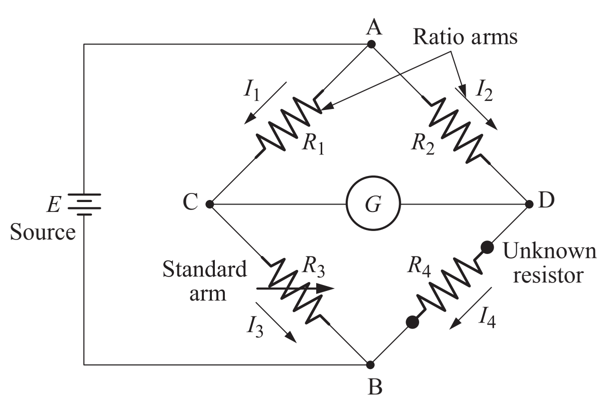


Figure 6.1 Wheatstone bridge circuit

Thus, the value of the unknown R_4 (or R_x) depends on the ratio (R_2/R_1) and the standard element R_3. The null detector must have sufficient sensitivity to indicate current flowing between points C and D because of the smallest potential difference between them. The measurement of the unknown resistance is independent of the characteristics or calibration of the null detector or galvanometer.

6.3.2 Measurement Errors

Errors may occur because of the following:

1. Insufficient sensitivity of the null detector. If the null detector is an ammeter, and if the imbalance current is few mA and if the meter is not able to measure the same, it will indicate null balance. If the detector is a voltmeter and if the potential difference is very small and the voltmeter is not able to read the same, null indication will be given. This can result in errors in determining the value of the unknown resistance *R*.
2. If the value of unknown resistance R_x is very small, the resistances of the wires used for connections in the circuit and also contact resistance can affect the measurement.
3. Thermal electromotive forces (*emfs*) in the bridge circuit can also affect balancing of the bridge. This arises because of dissimilar metals in contact with one another in making connections. The effect is significant for low values of R_x.
4. Changes in the values of resistances in the arms of the bridge circuit because of a rise in temperature can also affect the measurement.

6.3.3 Thevenin's Equivalent Circuit

Different galvanometers used as detectors in bridge circuits may have different values of resolution, that is, the minimum current required by them to cause deflection. The internal resistance of the galvanometers can also be different. It is possible to estimate the current through a given galvanometer for a small imbalance in the bridge circuit by solving Thevenin's equivalent. The Wheatstone bridge circuit and its Thevenin's equivalent circuit are shown in Fig. 6.2.

First, the open circuit voltage, when the galvanometer is removed and the arm CD is open, is determined. Then the equivalent resistance looking into terminals C and D with the battery replaced by its internal resistance is determined as the internal resistance of the battery can also limit the current flowing through the galvanometer. The current through the galvanometer can be determined by dividing Thevenin's open circuit voltage E_{Th} with the total resistance in the circuit, that is, the sum of Thevenin's equivalent resistance R_{Th} and the internal resistance of the galvanometer R_g.

$$I_g = \frac{E_{Th}}{R_{Th} + R_g} \tag{6.7}$$

Thevenin's open circuit voltage is

$$E_{cd} = E_{ac} - E_{ad} \tag{6.8}$$

$$= I_1R_1 - I_2R_2 \tag{6.9}$$

where

$$I_1 = \frac{E}{R_1 + R_3} \quad \text{and} \quad I_2 = \frac{E}{R_2 + R_4}$$

$$\therefore \quad E_{cd} = E\left(\frac{R_1}{R_1 + R_3} - \frac{R_2}{R_2 + R_4}\right) \tag{6.10}$$

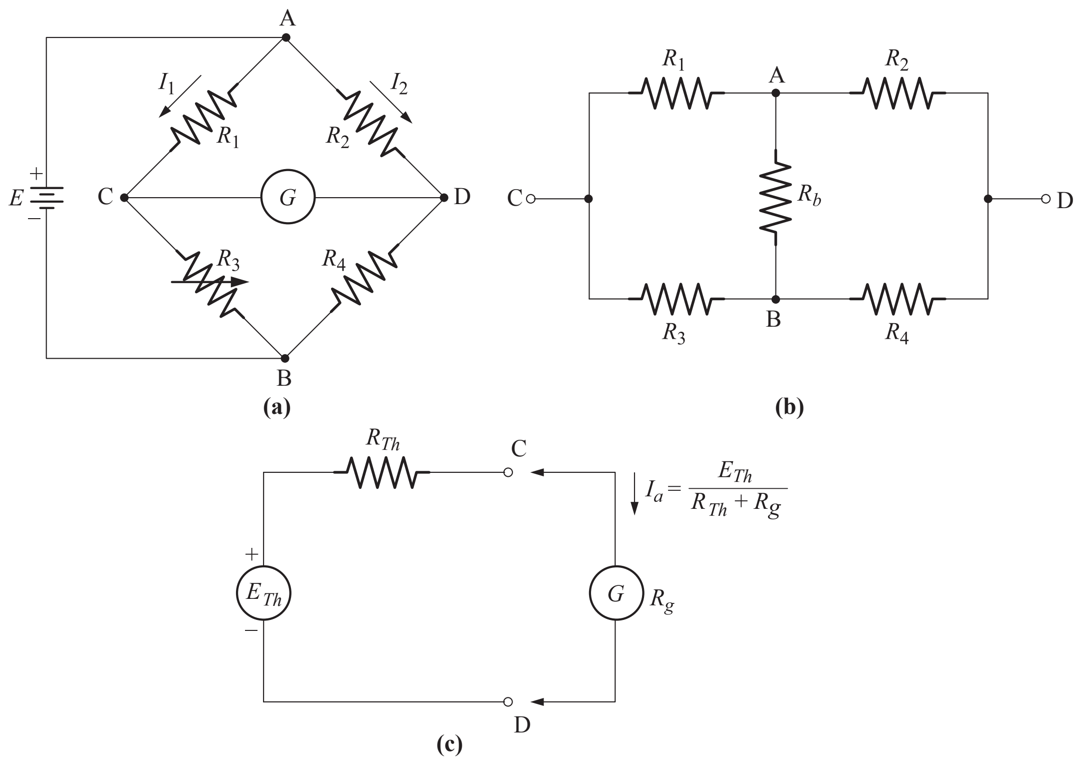


Figure 6.2 Application of Thevenin's theorem to the Wheatstone bridge: (a) Wheatstone bridge configuration, (b) Thevenin resistance looking into terminals C and D, and (c) complete Thevenin circuit with the galvanometer connected to terminals C and D

This is Thevenin's open circuit voltage. In practice, the internal resistance of the battery will be very low compared to R_1, R_2, R_3, and R_4.

Therefore, the circuit can be simplified as shown in Fig. 6.3.

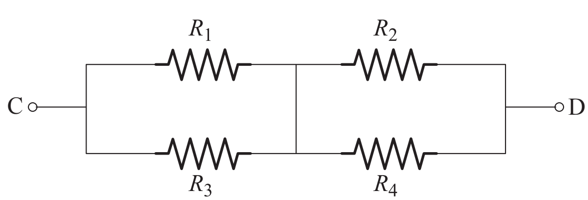


Figure 6.3 Simplified Thevenin's equivalent circuit

R_1 and R_3 are in parallel. This parallel combination is in series with the parallel combination of R_2 and R_4.

Therefore,

$$R_{Th} = \frac{R_1 R_3}{R_1 + R_3} = \frac{R_2 R_4}{R_2 + R_4} \tag{6.11}$$

If the null detector is now connected between points C and D, the current through the null detector or galvanometer I_g is

$$I_g = \frac{E_{Th}}{R_{Th} + R_g} \tag{6.12}$$

where R_g is the internal resistance of the galvanometer. Using this method, we can estimate the sensitivity of the galvanometer required to detect an imbalance for a given bridge circuit.

If the internal resistance of the battery R_b is also to be considered, the delta-wye (Δ–Y) transformation theorem is to be used and the R_{Th} value is determined.

In practice, the Wheatstone bridge is used to determine R in the range *of a few ohms to several M*Ω. If the R to be measured is very high, the current through the galvanometer will be needed. Therefore, the sensitivity is reduced. The lower limit is set by the contact resistances of the leads and the resistance associated with the wires. If the unknown resistance value is also of the same order, the measurement will not be correct.

Therefore, for determining very low values of resistance, Kelvin's bridge is used.

6.4 KELVIN BRIDGE

This bridge is a modification of the Wheatstone bridge circuit. This is used usually to measure *very low values of resistance in the range 1*Ω *to 0.00001* Ω. The effect or load resistances, contact resistances, and the resistances of the wires are also taken into account in the analysis of the bridge circuit. Thus Kelvin's bridge overcomes the limitations of the Wheatstone bridge in determining very low values of resistances.

Consider the bridge circuit shown in Fig. 6.4.

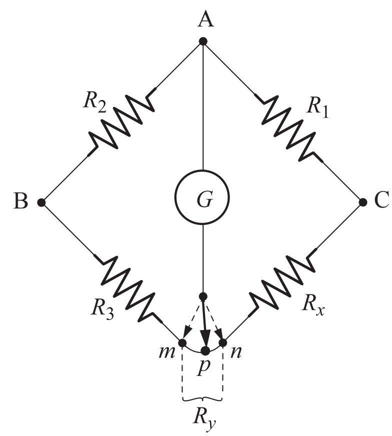


Figure 6.4 A Wheatstone bridge circuit showing resistance ***R*** of the lead from point ***m*** to point ***n***

R_3 represents the resistance of the connecting lead from R_3 to R_x. The galvanometer can be connected to point m or point n. If it is connected to m, the resistance of the connecting lead R_y is added to the unknown R_x. If it is connected to point n, R_y is added to arm R_3, and the resulting resistance R_x will be lower than the correct value. If the galvanometer is connected to point p, the ratio of resistances from n to p and from m to p equals the ratio of resistors R_1 and R_2.

$$\frac{R_{np}}{R_{mp}} = \frac{R_1}{R_2} \tag{6.13}$$

The balance equation is

$$R_x + R_{np} = \frac{R_1}{R_2}(R_3 + R_{mp}) \tag{6.14}$$

Substituting equation (6.13) in (6.14), we obtain

$$R_x + \left(\frac{R_1}{R_1 + R_2}\right)R_y = \frac{R_1}{R_2}\left[R_3 + \left(\frac{R_1}{R_1 + R_2}\right)R_y\right] \tag{6.15}$$

When simplified, this reduces to

$$R_x = \left(\frac{R_1}{R_2}\right)R_3 \tag{6.16}$$

This equation is the same as the one obtained for the Wheatstone bridge. This shows that the effect of the resistance of the connecting lead from point *m* to point *n* has been eliminated by connecting the galvanometer to the intermediate position *p*. Based on this principle, Kelvin's double bridge, which is known as the Kelvin bridge is constructed.

6.4.1 Kelvin Double Bridge

The circuit is shown in Fig. 6.5. This circuit has two additional arms; hence, it is also known as the Kelvin double bridge. The second set of arms *a* and *b* in the diagram connect the galvanometer points to *p* between *m* and *n*. It eliminates the effect of the yoke resistance R_y.

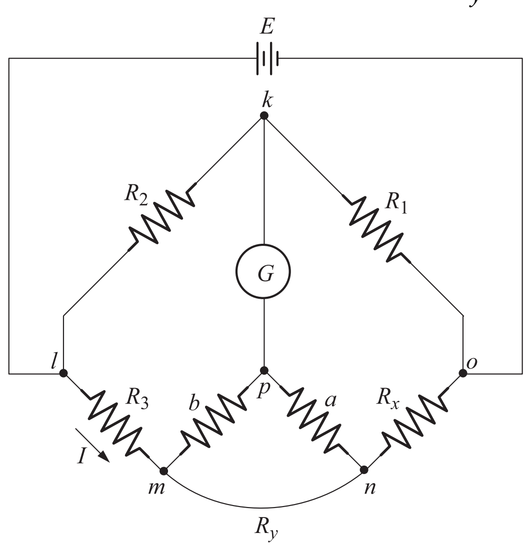


Figure 6.5 A basic Kelvin double bridge circuit

$\frac{R_1}{R_2}$ must be equal to the ratio of the resistances r_a and r_b.

i.e.,

$$\frac{R_1}{R_2} = \frac{r_a}{r_b}$$

Balance is achieved, when

$$E_{kl} = E_{lmp}$$

$$E_{kl} = \frac{R_2}{R_1 + R_2} E = \frac{R_2}{R_1 + R_2} I \left[R_3 + R_x + \frac{(r_a + r_b)R_y}{r_a + r_b + R_y} \right] \tag{6.17}$$

$$E_{lmp} = I \left[R_3 + \frac{r_a}{r_a + r_b} \left(\frac{(r_a + r_b)R_y}{r_a + r_b + R_y} \right) \right] \tag{6.18}$$

The equivalent circuit for networks m, p, n, R_3 and R_y is shown in Fig. 6.6.

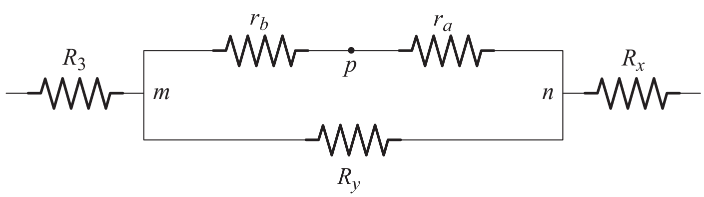


Figure 6.6 Equivalent circuit considering points *m*, *p*, *n,* and R_4

By equating E_{kl} and E_{lmp} we solve for R_x. For zero deflection, $E_{kl} = E_{lmp}$:

$$R_3 + R_x + \frac{(r_a + r_b)R_y}{r_a + r_b + R_y} = \frac{R_1 + R_2}{R_2} \left[R_3 + \frac{r_a}{r_a + r_b} + \left(\frac{(r_a + r_b)R_y}{r_a + r_b + R_y} \right) \right] \tag{6.19}$$

$$R_x = \frac{R_1 + R_3}{R_3} + \left[R_3 - R_3 - \frac{r_a R_y}{r_a + r_b + R_y} - \frac{r_b R_y}{r_a + r_b + R_y} + \frac{r_b R_y}{r_a + r_b + R_y} \right] \tag{6.20}$$

$$R_x = \frac{R_1 R_3}{R_2} + \frac{r_b R_y}{r_a + r_b + R_y} \left[\frac{R_1}{R_2} - \frac{r_a}{r_b} \right] \tag{6.21}$$

By choosing $R_1/R_2 = r_a/r_b$, the second term in the RHS of equation (6.21) becomes zero:

$$\therefore R_x = \frac{R_1 R_3}{R_2} \tag{6.22}$$

Although the lead resistances are considered, by choosing $(r_a/r_b) = (R_1/R_2)$, the unknown resistance R_x can be determined by the equation that is similar to the one obtained in the Wheatstone bridge circuit (Fig. 6.6).

6.4.2 Applications

1. A Wheatstone bridge is used in practice to measure resistance values from a few ohms to few MΩ.
2. Strain gauge bridge circuits are based on the Wheatstone bridge.
3. The insulation resistance of a cable and the leakage resistance of a capacitor, which are usually very high (few 100–1000 MΩ), are measured by a *guarded Wheatstone bridge circuit*.

4. Kelvin's bridge is used to measure resistances of very low values, such as the resistance of the contacts, leads, and connecting wires.

6.5 STRAIN GAUGE BRIDGE CIRCUIT

The change in resistance of a strain gauge owing to external physical parameters such as force, pressure, stress, etc., is measured based on the Wheatstone bridge circuit. The imbalance current or the output voltage is calibrated in terms of strain. In the standard arm, a standard strain gauge identical to the one being used in the unknown arm is connected (Fig. 6.7).

At balance,

$$I_1 = \frac{E}{R_1 + R_3} \tag{6.23}$$

$$I_2 = \frac{E}{R_2 + R_x} \tag{6.24}$$

Owing to strain, let R_x change by ΔR_x. Therefore, the new value is $(R_x + \Delta R_x)$.

Therefore, the imbalance output voltage e_0 is approximately

$$e_0 = I_1 R_3 - I_2 (R_x + \Delta R_x) \tag{6.25}$$

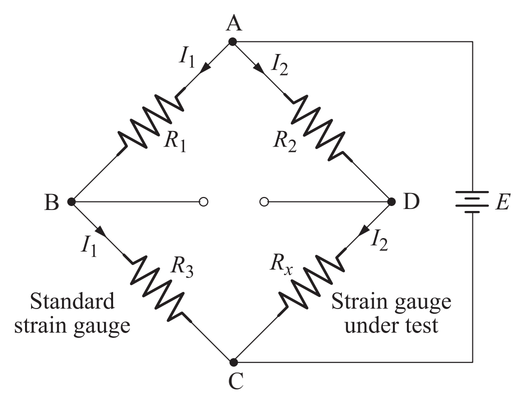


Figure 6.7 Strain gauge bridge circuit

$$e_0 = \frac{ER_3}{R_1 + R_3} - \frac{E}{R_2 + R_x + \Delta R_x}(R_x + \Delta R_x) \tag{6.26}$$

If an equal arm bridge is employed and $R_1 = R_2 = R_3 = R_x$, each is equal to R:

$$e_0 = \frac{ER}{2R} - \frac{E}{2R + \Delta R}(R + \Delta R) \tag{6.27}$$

$$e_0 = \frac{E\cancel{R}}{2\cancel{R}} - \frac{E}{2\cancel{R}\left(1 + \frac{\Delta R}{2R}\right)}\cancel{R}\left(1 - \frac{\Delta R}{R}\right)$$

$$e_0 = \frac{E}{2} - \frac{E}{2\left(1+\frac{\Delta R}{2R}\right)}\left(1+\frac{\Delta R}{R}\right) \tag{6.28}$$

$$= \frac{E}{2}\left[1-\left(1+\frac{\Delta R}{R}\right)\left(1+\frac{\Delta R}{2R}\right)^{-1}\right] \tag{6.29}$$

Applying binomial expansion and neglecting higher order $(\Delta R)^2$ terms, ΔR is very small:

$$e_0 = \frac{E}{2}\left[1-\left(1+\frac{\Delta R}{2R}\right)\left(1-\frac{\Delta R}{2R}\right)^{-1}\right] \tag{6.30}$$

$$e_0 = \frac{E}{2}\left[1-\left(1+\frac{\Delta R}{R}-\frac{\Delta R}{2R}\right)-\left(\frac{\Delta R}{2R}\right)^2\right] \tag{6.31}$$

$$e_0 = \frac{E}{2}\frac{\Delta R}{2R} \tag{6.32}$$

$$e_0 = \frac{E\Delta R}{4R} \tag{6.33}$$

This is an approximate equation for the imbalance output voltage.

Example 6.1

For the bridge circuit shown in Fig. 6.8, R_1 = 1000 Ω, R_2 = 4000 Ω, R_3 = 100 Ω, R_4 = 400 Ω. The galvanometer has an internal resistance of 100 Ω and a current sensitivity of 100 mm/μA. The battery voltage is 3 V. Calculate the galvanometer deflection for an imbalance of 1 Ω in the resistance R_4. The use of a calculator is forbidden.

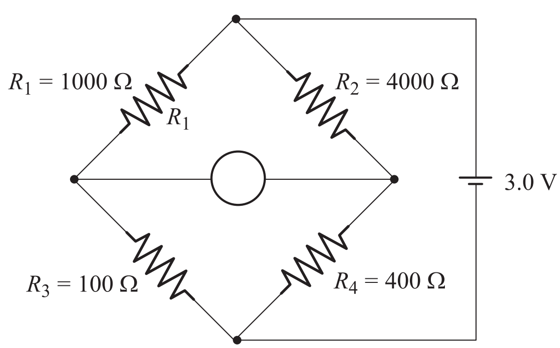


Figure 6.8 For Example 6.1

Solution

The bridge is originally in balance.

Therefore,

$$\frac{R_1}{R_2} = \frac{R_2}{R_4}; \quad \frac{1000\ \Omega}{100\ \Omega} = \frac{4000\ \Omega}{400\ \Omega}$$

Let there be an imbalance in the bridge circuit because of an increase in the R_4 value by 1 Ω.

Therefore,

$$R_4 = (400 + X)\ \Omega$$

Thevenin's resistance

$$R_{Th} = \frac{100 \times 1000}{1000 + 100} + \frac{4000\,(400 + X)}{(4400 + X)}$$

Neglecting X,

$$R_{Th} = \frac{1000}{11} + \frac{4000}{11} = 456\ \Omega$$

$$E_{Th} = \left[\frac{R_3}{R_1 + R_3} + \frac{R_4}{R_1 + R_4}\right] E$$

$$= 3\left[\frac{100}{1100} + \frac{(400 + X)}{(4400 + X)}\right]$$

$$= 3\left[\frac{1}{11} - \left\{\frac{400\left(1 + \dfrac{X}{100}\right)}{4000\left(1 + \dfrac{X}{4400}\right)}\right\}\right]$$

$$= 3\left[\frac{1}{11} - \left\{\frac{1}{11}\left(1 + \frac{X}{400}\right)\left(1 + \frac{X}{4000}\right)^{-1}\right\}\right]$$

Applying binomial expansion and neglecting X_2 terms, X is small:

$$E_{Th} = 3\left[\frac{1}{11} - \left\{\frac{1}{11}\left(1 + \frac{X}{400}\right)\left(1 + \frac{X}{4000}\right)^{-1}\right\}\right]$$

$$= \frac{3}{11}\left[\frac{4400 - 4400 - 10X}{4400}\right] = \frac{10X}{48,400} \times 3\ \text{V}$$

for $X = 1\ \Omega$, $E_{Th} = 620\ \mu\text{V}$.

Galvanometer current

$$I_G = \frac{E_{Th}}{R_{Th} + R_G} = \frac{620\ \mu\text{V}}{(456 + 100)} = 1.114\ \mu\text{A}$$

Deflection

$$D = 1.114\ \mu\text{A} \times \frac{10\ \text{mm}}{\mu\text{A}} = 11.14\ \text{mm}$$

Example 6.2

The ratio arms of the Wheatstone bridge shown in Fig. 6.9 are $R_1 = 1000\ \Omega$, $R_2 = 100\ \Omega$. The standard resistance $R_3 = 400\ \Omega$. The unknown resistance $R_X = 41\ \Omega$. A 1.5 V battery with negligible internal

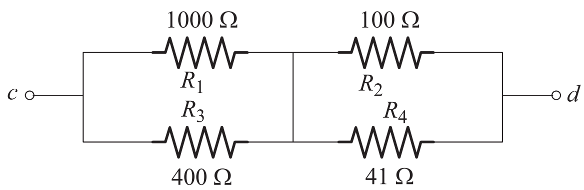


Figure 6.9 For Example 6.2

resistance is connected from the junction of R_1 and R_2 to the junction of R_3 and R_X. The galvanometer with an internal resistance of 50 Ω and a current sensitivity of 2 mm/μA is connected to the other corner of the bridge (Fig. 6.10). Calculate the deflection of the galvanometer caused by the imbalance in the circuit.

Solution

$$R_{Th} = \frac{R_1 R_3}{R_1 + R_3} + \frac{R_2 R_X}{R_2 + R_X}$$

$$R_{Th} = \frac{1000 \times 400}{1000 + 400} + \frac{100 \times 41}{100 + 41} = 315\ \Omega$$

$$E_{Th} = \left[\frac{R_3}{R_1 + R_3} - \frac{R_X}{R_X + R_3}\right]$$

$$E_{Th} = \left[\frac{400}{400 + 1000} - \frac{41}{41 + 100}\right] 1.5 = 7.6\ \text{mV}$$

$$I_g = \frac{E_{Th}}{R_{Th} + R_g} - \frac{7.6\ \text{mV}}{3.5 + 50} = 20.8\ \text{mA}$$

Deflection d = 2 mm/μA × 20.8 = 41.6 mm.

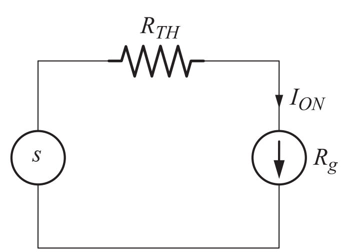


Figure 6.10 For Example 6.2

6.6 AC BRIDGES

To determine the value of unknown reactive components, an AC input is to be used as the source. Such bridges are classified as AC bridges. In addition, the resistance, the Z, Y, X frequency of the source signal, the Q factor of a coil, the dissipation factor of a capacitor D can also be determined using AC bridges. Balancing of AC bridges is more difficult than DC bridges because both magnitude and phase angle conditions have to be satisfied for balance. The detector must respond to imbalance AC current. Headphones are used as new detectors. No sound signal is heard if the bridge is in balance. A Cathode Ray Oscilloscope (CRO) AC amplifier with an output meter or a tuning eye (electron ray tube indicator) are some of the detectors used for AC bridges.

6.6.1 General Form of Bridge Circuit

The circuit is shown in Fig. 6.11.

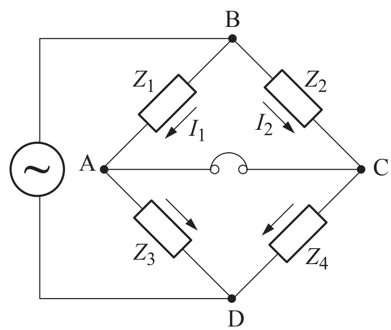


Figure 6.11 General form of the AC bridge

At balance,

$$E_{BA} = E_{BC} \tag{6.34}$$

or,

$$I_1 Z_1 = I_2 Z_2 \tag{6.35}$$

$$I_1 = \frac{E}{Z_1 + Z_3} \tag{6.36}$$

$$I_2 = \frac{E}{Z_2 + Z_4} \tag{6.37}$$

Substituting equations (6.35) and (6.36) in equation (6.34),

$$Z_1 Z_4 = Z_2 Z_3 \tag{6.38}$$

In terms of admittance,

$$Y_1 Y_4 = Y_2 Y_3 \tag{6.39}$$

From the above equations, the product of impedances of one pair of opposite arms must be equal to the product of impedances of the other pair of opposite arms. Note that the impedances are complex numbers.

They have magnitude as well as phase angle:

$$Z_1 = z_1 \angle\theta_1; \quad Z_2 = z_2 \angle\theta_2; \quad Z_3 = z_3 \angle\theta_3; \quad Z_4 = z_4 \angle\theta_4$$

where z_1, z_2, z_3, and z_4 represent the magnitudes.

Therefore, equation (6.37) in terms of magnitudes and phase angles becomes

$$z_1 \angle\theta_1 z_4 \angle\theta_4 = z_2 \angle\theta_2 z_3 \angle\theta_3 \tag{6.40}$$

$$z_1 z_4 = z_2 z_3 \tag{6.41}$$

$$\angle\theta_1 + \angle\theta_4 = \angle\theta_2 + \angle\theta_3 \tag{6.42}$$

Therefore, for the balance of AC bridges,

1. The products of magnitudes of the opposite arms must be equal.
2. The sum of the phase angles of the opposite arms must be equal.

6.7 MAXWELL BRIDGE

This bridge is used to determine the value of an unknown *inductance L* in terms of the known capacitance. The bridge circuit is shown in Fig. 6.12.

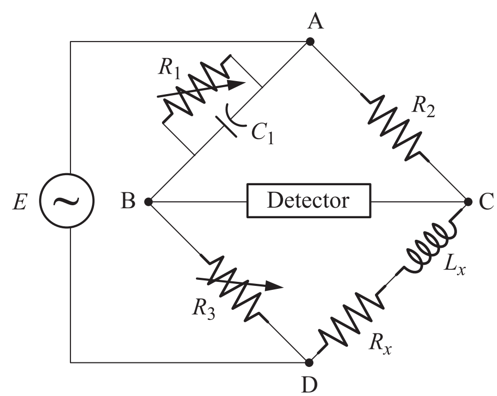


Figure 6.12 Maxwell bridge for inductance measurements

At balance,

$$Z_1 Z_x = Z_2 Z_3 \; 87 \tag{6.43}$$

$$Z_1 = R_1 \text{ parallel with } X_{C_1} = \frac{R_1}{1 + j\omega C_1 R_1}$$

$$Z_2 = R_2$$

$$Z_3 = R_3$$

$$Z_x = R_x + j\omega L_x$$

Substituting in equation (6.42),

$$\frac{R_1}{(1 + j\omega C_1 R_1) R_3} = \frac{R_2}{R_x + j\omega L_x} \tag{6.44}$$

$$R_1 R_x + jR_1 \omega L_x = R_2 R_3 + j\omega R_2 R_3 C_1 R_1 \tag{6.45}$$

Equating real parts,

$$R_1 R_x = R_2 R_3 \tag{6.46}$$

Therefore,

$$R_x = \frac{R_2 R_3}{R_1} \tag{6.47}$$

Equating imaginary parts,

$$R_1 \omega L_x = j\omega R_2 R_3 C_1 R_1 \tag{6.48}$$

Therefore,

$$L_x = R_2 R_3 C_1 \tag{6.49}$$

This bridge is used for coils with the Q factor lying in the range 1–10 (1 < Q < 10).

If the Q factor is large, R_x is small. Therefore, R_1 in the balance equation (6.46) must be very high. Balancing the bridge becomes difficult. It is also unsuitable for coils with $Q < 1$.

6.7.1 Phasor Diagram for the Maxwell Bridge

1. In arm AB, elements R_1 and C_1 are in parallel.
2. Therefore, take the voltage V_{AB} as a reference and draw the phasor diagram.
3. I_c leads V_{AB} by 90°. IR_1 and V_{AB} are in phase. IR_1 is in phase with V_{AB}. The vector summation of IR_1 and I_c is I_{AB}.
4. At balance, $V_{AB} = V_{AC}$. As R_2 is the element in arm AC, I_{AC} is in phase with V_{AC}.
5. At balance, the same I_{AB} flows through R_3 in arm BD
6. Voltage across L_X leads the current through it by 90°. The current through R_X and V_{RX} are in phase. Therefore, I_{CD} current through L_X is as shown in the phasor diagram (Fig. 6.13).
7. Total current I is the vector summation of I_{AB} and I_{AC}.
8. Total voltage is the vector summation of V_{AB} and V_{BD}.

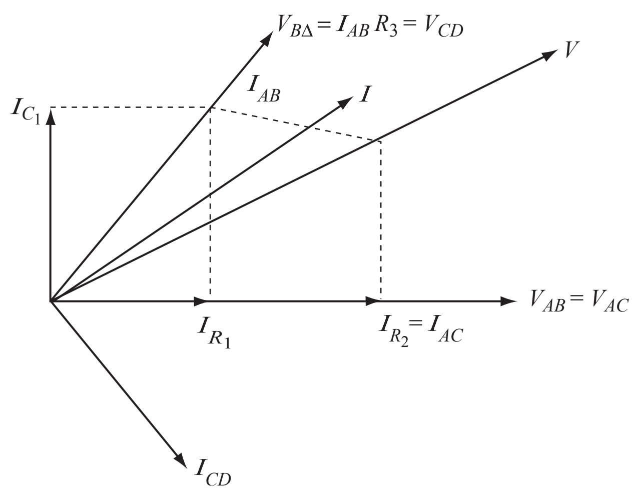


Figure 6.13 Vector diagram for Maxwell's bridge

6.8 HAY BRIDGE

This bridge circuit (shown in Fig. 6.14) is developed to overcome the limitations of the Maxwell bridge. R_1 is in series with a standard capacitor C_1 unlike the Maxwell bridge, where it is in parallel. *This bridge circuit can be used for high Q coils with a Q factor > 10.*

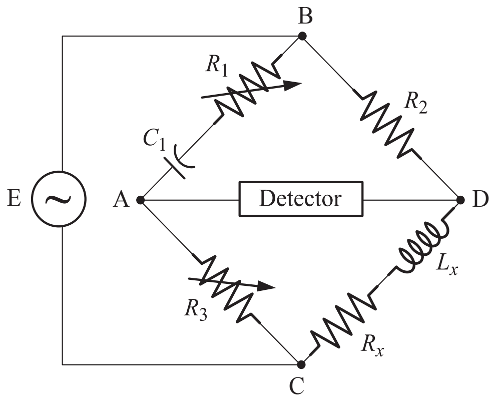


Figure 6.14 Hay bridge for inductance measurements

At balance,

$$\frac{R_1 - \frac{j}{\omega C_1}}{R_3} = \frac{R_2}{R_x + j\omega L_x} \tag{6.50}$$

$$R_1 R_x + \frac{L_x}{C_1} - \frac{jR_x}{\omega C_1} + j\omega R_1 L_x = R_2 R_3 \tag{6.51}$$

$$R_1 R_x + \frac{L_s}{C_1} = R_2 R_3 \tag{6.52}$$

$$\frac{R_x}{\omega C_1} = \omega L_x R_1 \tag{6.53}$$

Both equations contain the unknown R_x and L_x. Solving for the unknowns, eliminating R_x from equation (6.53),

$$L_x = \frac{R_x}{\omega^2 C_1 R_1} \tag{6.54}$$

Substituting this value of L_x in equation (6.54),

$$R_1 R_x + \frac{R_x}{\omega^2 C_1 R_1} = R_2 R_3 \tag{6.55}$$

$$\therefore R_x = \frac{R_2 R_3}{R_1} - \frac{R_x}{\omega^2 C_1^2 R_1^2} \tag{6.56}$$

$$\text{or } R_x = \frac{R_2 R_3}{R_1} \tag{6.57}$$

$$\therefore R_x = \frac{\omega^2 R_2 R_3 C_1^2 R_1}{1+\omega^2 C^2 R^2} \tag{6.58}$$

Similarly

$$L_x = \frac{R_2 R_3 C_1}{1+\omega^2 C_1^2 R_1^2} \tag{6.59}$$

The Hay bridge will not give correct results for $Q < 10$ because at balance,

$$\tan\theta_c = \tan\theta_L \tag{6.60}$$

$$\tan\theta_c = \frac{X_c}{R}; \quad \therefore \tan\theta_L = \frac{\omega L}{R} \tag{6.61}$$

$$\therefore \frac{1}{\omega C R} = \frac{\omega L}{R} \tag{6.62}$$

$$\text{or } \omega C_1 R_1 = Q \tag{6.63}$$

$$\therefore L_x = \frac{R_2 R_3 C}{1+\left(\frac{1}{Q^2}\right)} \tag{6.64}$$

for $Q > 10$, $L_x = R_2 R_3 C$ since $\frac{1}{Q^2}$ can be neglected compared to 1.

If Q is small, this equation will not be valid. The Hay bridge is known as an *opposite angle* bridge since C and L elements are involved, one is a *leading angle* element and the other a *lagging angle* element.

6.8.1 Phasor Diagram for Hay Bridge

1. In arm AB, the elements are in series. Therefore, I_{AB} can be taken as reference. V_{C_1} lags the current through C_1 by 90°. The vector sum of V_{C_1} and $I_{AB}R_1$ is V_{AB} (Fig. 6.15).
2. At balance, $V_{AB} = V_{AC}$; I_{AC} and V_{AC} are in phase, source R_2 is the only element in arm AC.
3. At balance, I_{AB} also flows through arm BD. R_3 is the only element in arm BD. Therefore, $I_{BD} = I_{AB}$ and V_{BD} will be in phase.

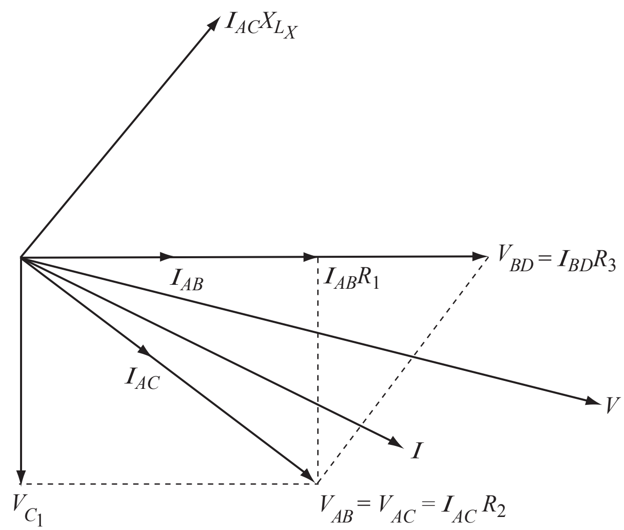


Figure 6.15 Vector diagram for a Hay bridge

4. The drop across L_X in arm CD leads the current through L_X, I_{AC} by 90°, as inductor is the element. The vector sum of $I_{AC}R_X$ and V_{C_X} is $V_{LX,}$ which is $V_{CD} = V_{BD}$.
5. Total current I is the vector sum of I_{AB} and I_{AC}. The vector sum of V_{AB} and V_{BD} is the total voltage V.

6.9 SCHERING BRIDGE

The Schering bridge is used extensively for the measurement of *capacitance, particularly for insulators* with a phase angle of nearly 90º. In the standard arm of the bridge, a high-quality mica capacitor for general measurements and an air capacitor for insulator measurements are used. The circuit is shown in Fig. 6.16.

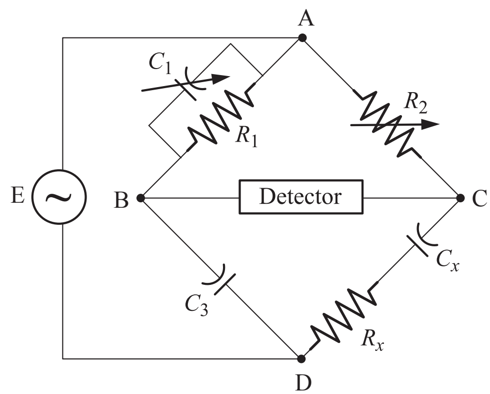


Figure 6.16 Schering bridge for the measurement of capacitance

At balance,

$$Z_x = \frac{Z_2 Z_3}{Z_1}$$

$$R_x + \frac{1}{j\omega C_x} = \frac{1}{j\omega C_3}\frac{\left(\frac{-R_2}{R_1}\right)}{1 + j\omega C_1 R_1} \tag{6.65}$$

$$R_x - \frac{-j}{\omega C_x} = \frac{-j}{j\omega C_3}\left\{\frac{R_2}{R_1}\left(1 + j\omega C_1 R_1\right)\right\} \tag{6.66}$$

$$R_x - \frac{-j}{\omega C_x} = \frac{R_2}{R_3}\frac{1}{\omega C_3} + \frac{C_1 R_2}{C_3} \tag{6.67}$$

$$R_x = R_2 \frac{C_1}{C_3} \tag{6.68}$$

$$\boxed{C_x = C_3 \frac{R_1}{R_2}} \tag{6.69}$$

R_x is the series resistance associated with the capacitor C_x.

Dissipation factor $D = \boxed{D = \frac{R_x}{X_c} \omega C_x R_x}$

Power factor PF = $\boxed{PF = \frac{R_x}{Z_x}}$

The dissipation factor gives the indication of how close the phase angle is to 90°.

6.9.1 Phasor Diagram for a Schering Bridge

1. In arm AB, R_1 and C_1 are in parallel. Therefore, as per convention, voltage V_{AB} is taken as reference. V_{AB} and I_{R_1} are to be in phase. I_{C_1} leads V_{AB} by 90° (Fig. 6.17).
2. At balance, $I_{AB} = I_{BD}$; V_{BD} lags with respect to I_{BD} (or = I_{AB}) by 90°.
3. At balance, $V_{AB} = V_{AC}$; I_{AC} will be in phase with V_{AC}.
4. At balance, $I_{AC} = I_{CD}$; I_{AC} and V_{R_X} will be in phase. V_{C_X} lags I_{C_X} by 90°.
5. Total voltage 'V' is the vector sum of V_{AB} and V_{BD}.
6. Total current I is the vector sum of I_{AB} and I_{AC}.

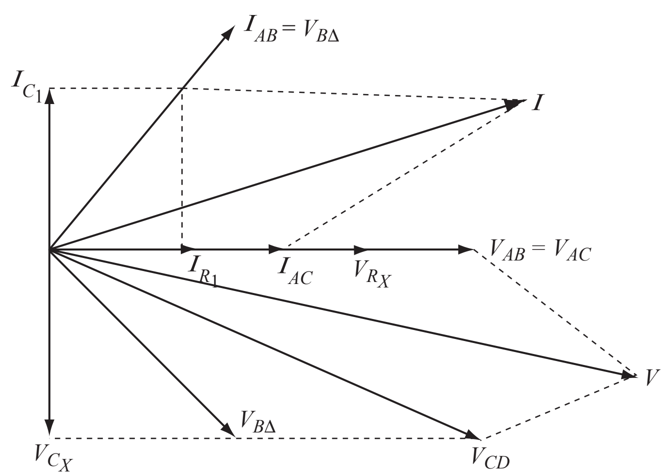


Figure 6.17 Phasor diagram for a Schering bridge

6.10 WIEN BRIDGE

In this bridge circuit, there is a lead-lag network. Balancing of the bridge is easier because satisfying the phase angle equality condition can be achieved. This bridge can also be used to determine the frequency of the AC input in terms of the component values of the bridge circuit. In this AC bridge, there is no inductor. Inductive losses because of stray fields cause problems in balancing of the bridge. Owing to the absence of L in the circuit, shown in Fig. 6.18, this can be effectively used for determining the frequency f of the AC input.

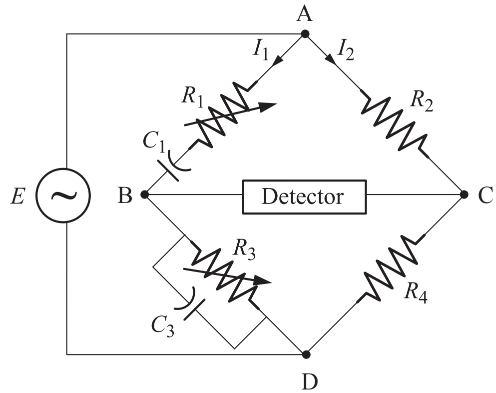


Figure 6.18 Frequency measurement with the Wien bridge

At balance,

$$I_1\left(R_1 + \frac{1}{j\omega C_1}\right) = I_2R_2 \tag{6.70}$$

$$I_1\left(\frac{R_3}{1+j\omega C_3R_3}\right) = I_2R_4 \tag{6.71}$$

$$\therefore \frac{R_1 - \dfrac{j}{\omega C_1}}{\dfrac{R_3}{1+j\omega C_3R_3}} = \frac{R_2}{R_4} \tag{6.72}$$

$$R_2 = \frac{\left(R_1R_4 - \dfrac{jR_4}{\omega C_1}\right)(1+j\omega C_3R_3)}{R_3} \tag{6.73}$$

$$= \frac{R_1R_4}{R_3} - \frac{jR_4}{\omega R_3C_1} + \frac{j\omega R_1R_4C_3R_3}{R_3} + \frac{R_3R_4C_3}{R_3C_1} \tag{6.74}$$

$$R_2 = \frac{R_1R_4}{R_3} + \frac{R_4C_3}{C_1} \tag{6.75}$$

Equating imaginary parts,

$$\omega R_1R_4C_3 = \frac{R_4}{\omega R_3C_1} \tag{6.76}$$

$$\omega_2 = \frac{R_4}{R_1R_4C_3R_3C_1} \tag{6.77}$$

or

$$f = \frac{1}{2\pi\sqrt{R_1R_3C_1C_3}} \tag{6.78}$$

If $R_1 = R_3 = R,\ C_1 = C_3 = C,$

$$f = \frac{1}{2\pi RC} \tag{6.79}$$

Thus, the frequency of the *AC* input can be determined, usually in the audio frequency range. Because of its frequency sensitivity, if harmonics are present in the AC input, the bridge will not be balanced exactly.

6.10.1 Phasor Diagram for the Wien Bridge

1. In arm *AB*, R_1 and C_1 are in series. Therefore, by convention, I_{AB} will be in phase. V_{C_1} lags I_{AB} by 90° (Fig. 6.19).
2. At balance, $V_{AB} = V_{AC}$; I_{AC} will be in phase with V_{AC}.
3. At balance, $V_{CD} = V_{BD} I_{R_3}$ will be in phase with V_{BD}
4. At balance, $V_{BD} = V_{CD} I_{R_3}$ will be in phase with V_{BD}; I_{C_3} leads V_{BD} by 90°. The vector sum of I_{R_3} and I_{C_3} is equal to I_{AB}.
5. Vector sum of I_{AB} and I_{AC} is I.
6. Vector sum of V_{AB} and V_{BD} is V.

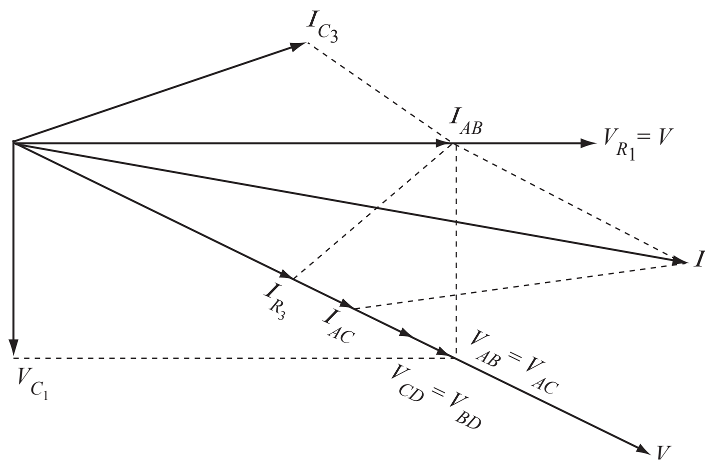


Figure 6.19 Phasor diagram for the Wien bridge

6.11 ANDERSON BRIDGE

In the Anderson bridge, the *unknown inductor's value* is determined in terms of known values of *capacitance* and *resistance*. The value of inductance can be measured over a wide range of values.

The circuit diagram is shown in Fig. 6.20.

Using star-delta transformations,

$$R_{AB} = R_4 R_5 \left(\frac{1}{R_1} + \frac{1}{R_4} + \frac{1}{R_5} \right)$$

$$R_{BC} = R_1 R_5 \left(\frac{1}{R_1} + \frac{1}{R_4} + \frac{1}{R_5} \right)$$

Similarly,

$$R_{AC} = R_1 R_5 \left(\frac{1}{R_1} + \frac{1}{R_4} + \frac{1}{R_5} \right)$$

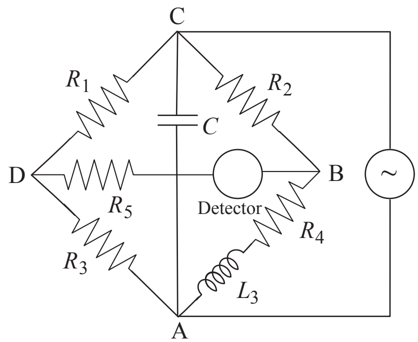


Figure 6.20 Anderson bridge

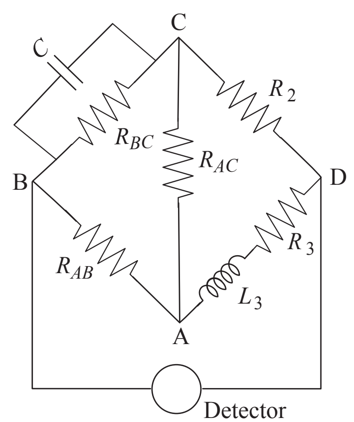


Figure 6.21 Transformed Anderson bridge using λ–Δ transformation.

With star-delta transformations, the redrawn circuit is as shown in Fig. 6.21. Therefore, the bridge can be reduced to the Maxwell–Wien bridge. Using the balance equation of the Maxwell–Wien bridge:

$$R_{BC}R_3 = R_2R_{AB} \quad \text{and} \quad L = CR_2R_{AB}$$

Substituting the values of R_{BC} and R_{AB} in respective equations, we get

$$R_1R_5R_3\left(\frac{1}{R_1}+\frac{1}{R_4}+\frac{1}{R_5}\right) = R_2R_4R_5\left(\frac{1}{R_1}+\frac{1}{R_4}+\frac{1}{R_5}\right)$$

$$\therefore R_1R_3 = R_2R_4$$

Similarly

$$L = CR_2R_4R_5\left(\frac{1}{R_1}+\frac{1}{R_4}+\frac{1}{R_5}\right)$$

or

$$\boxed{L = CR_2\left[R_5\left(1+\frac{R_4}{R_1}\right)+R_4\right]}$$

6.12 RESONANCE BRIDGE

Reactive elements C_d and L_d are connected in one arm to give series resonance so that this arm offers a resistance that can be balanced by varying R_c. This bridge can also be used to determine the frequency in terms of L and C or L in terms of C and f or C in terms of L and f (Fig. 6.22).

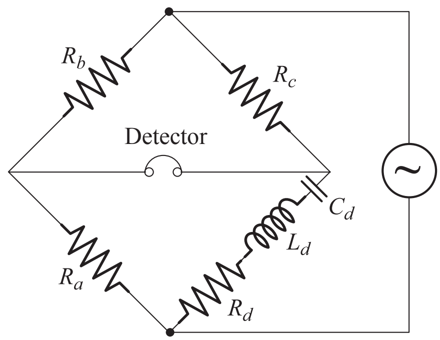


Figure 6.22 Resonance bridge

At balance,

$$\omega L_d = \frac{1}{\omega C_d} \tag{6.80}$$

$$R_d = \left(\frac{R_a}{R_b}\right) R_c \tag{6.81}$$

$$\omega_2 = \frac{1}{L_d C_d} \tag{6.82}$$

$$\boxed{f = \frac{1}{2\pi\sqrt{L_d C_d}}} \tag{6.83}$$

In Fig. 6.22, a *headphone is shown as the detector*.

6.13 SIMILAR ANGLE BRIDGE

This bridge is known as the similar angle bridge because only capacitive elements are involved. It is also called a capacitance comparison bridge (Fig. 6.23).

At balance,

$$R_1 (R_x - jX_{c_x}) = R_3(R_2 - jXC_2) \tag{6.84}$$

$$R_1 R_x - jR_1 X_{c_x} = R_2 R_3 - jR_3 Xc_2 \tag{6.85}$$

$$\therefore R_1 R_x = R_2 R_3 \tag{6.86}$$

$$R_1 X_{c_x} = R_3 Xc_2 \tag{6.87}$$

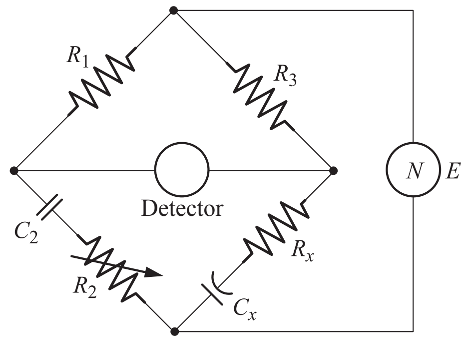


Figure 6.23 Similar angle bridge

$$\therefore \boxed{R_x = \left(\frac{R_1}{R_2}\right) R_3} \tag{6.88}$$

$$\boxed{C_x = \left(\frac{R_1}{R_3}\right) C_2} \tag{6.89}$$

The expression for R_x and C_x does not contain the term f or E. Hence, the measurement of the unknown element is *independent of the supply frequency or the supply voltage.*

6.14 RADIO FREQUENCY BRIDGE (SUBSTITUTION TECHNIQUE)

The bridge is first balanced by shorting Z_x terminals. The values of C_2 and C_3 are noted. Then unknown impedance Z_x is inserted, and the values of C_2 and C_3 are noted again (Fig. 6.24).

Then,

$$R_x = \left(\frac{R_1}{C_1}\right)\left(C\frac{1}{3} - C_3\right) \tag{6.90}$$

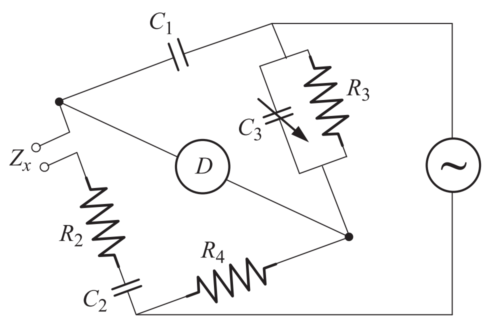


Figure 6.24 Radio frequency bridge

$$X_x = \frac{1}{\omega}\left(\frac{1}{C'_3} - \frac{1}{C_2}\right) \tag{6.91}$$

Before Z_x is inserted,

$$C_3 = \left(\frac{R_2}{R_1}\right) C_1 \tag{6.92}$$

$$C_2 = \left(\frac{R_3}{R_1}\right) C_1 \tag{6.93}$$

The unknown impedance $R_x \pm jL_x$ is inserted and the balance is again obtained.

Then,

$$R_x = \frac{R_1 C'_3}{C_1} - R_2 \tag{6.94}$$

$$X_x = \frac{1}{\omega C'_2} - \frac{R_4}{\omega R_3 C_1} \tag{6.95}$$

Solving,

$$\boxed{R_x = \frac{R_4}{C_1}(C_3 - C_3)} \tag{6.96}$$

$$\boxed{X_x = \frac{1}{\omega}\left(\frac{1}{C'_2} - \frac{1}{C_2}\right)} \tag{6.97}$$

This circuit is also known as the radio frequency bridge. This is used to determine the impedance of both C and L type of loads at high frequencies.

6.15 WAGNER'S GROUND CONNECTION

In the bridge circuit analysis, it is assumed that the bridge elements are simple lumped parameters, without any interaction between them. However, stray capacitances exist between the various bridge elements and the ground and also between the bridge arms themselves. The capacitive reactance shunting an arm of the bridge is a function of frequency. At high frequencies, since $X_c = \left(\frac{1}{2\pi fC}\right)$ the capacitive reactance because of the stray capacitance becomes very small or even approaches zero value. Hence, the arm of the bridge will be shorted causing measurement errors. The problem arises because of large inductors also. Balancing of the bridge becomes difficult because of these stray reactive elements.

This problem can be solved to some extent by shielding the arm of the bridge and by connecting the shield to the ground. By this arrangement, the stray capacitance cannot be eliminated, but the value can be made constant. Then, this stray capacitance can be compensated. Wagner's ground connection a commonly used method to eliminate some of these problems. The capacitance that exists between the detector and the ground can be eliminated by this method.

In Fig. 6.25, C_1 and C_2 represent the stray capacitances that exist between the two terminals of the detector and the ground. It is to be observed that the ground connection in the circuit is different compared to the conventional bridge circuits. The input oscillator terminal is not grounded directly with one corner

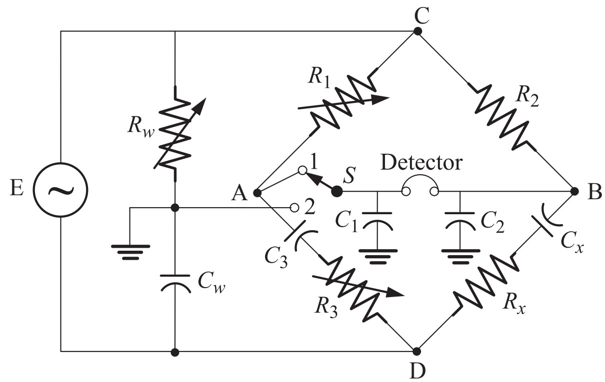


Figure 6.25 Wagner's ground connection eliminates the effect of stray capacitances across the detector

of the bridge circuit. Instead, a series combination of resistor R_w and capacitor C_w are connected across the oscillator. The junction of R_w and C_w is grounded. This is known as *Wagner's Ground Connection*.

The detector is first connected to point 1 and R_1 is adjusted for null or minimum sound on headphones. The switch is then connected to position 2, which connects the detector to Wagner's ground point. In Fig. 6.25, a headphone is shown as a detector. Resistor R_w is now adjusted for minimum sound. When the switch is thrown to position 1 again, some imbalance might be there. Therefore, R_1 and R_2 are again adjusted for balancing the bridge or for a minimum detector response and the switch is again thrown to position 2. This is repeated till null balance is obtained for both positions of the switch at points 1 and 2 by adjusting R_w, R_1, and R_3. When the null sound is finally obtained, points 1 and 2 are at the same potential and this is called the *ground potential.* Stray capacitances C_1 and C_2 are shorted and have no effect on normal bridge balance.

Thus Wagner's ground connection eliminates the effect of stray capacitances at the detector terminals. However, stray capacitance across the arms of the bridge can still cause problems in balancing. This is not eliminated by Wagner's ground connection.

6.16 TWIN-*T* NULL NETWORK

This is also known as the *Parallel-T Network*. It will have two different *T*-networks arranged in parallel with input and output terminals. A zero output is obtained when the circuit impedances of the individual branches are so arranged that the transmission through the two *T*-networks to the output terminal is equal in magnitude but opposite in phase (Fig. 6.26). This is the condition for balance.

The special features of twin-*T* networks are as follows:

1. The input and output terminals are in parallel. In other bridge circuits seen so far, they are in perpendicular directions.
2. The input and output have a common terminal that can be grounded. This minimises the shielding problems. Due to the common ground terminal, a shielding transformer is not required as in the case of the Wheatstone bridge.
3. Because of the lead – lag network used in a twin-*T* network, balancing is obtained easily.

In analysing the twin-*T* networks, it is easy if we assume that output is shorted and determine the condition for which the short circuit current passes through the two *T*-networks equal in magnitude, but opposite in phase (Fig. 6.27).

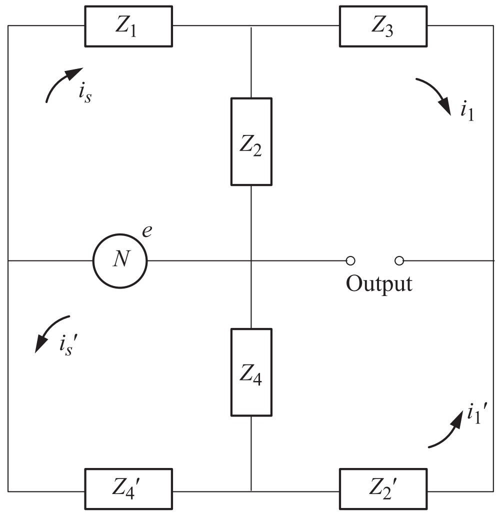


Figure 6.26 Twin *T*-network

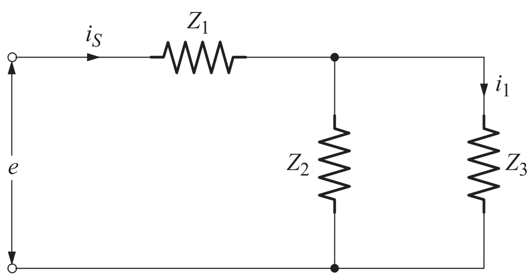


Figure 6.27 Simplified form of one *T*-network

Let i_1 be the output current for the *T*-network Z_1, Z_2, and Z_3. Z_2 and Z_3 are in parallel because it is assumed that output terminals are shorted. The parallel combination of Z_2 and Z_3 is in series with

$$Z_i Z_s = Z_i + \frac{Z_2 Z_3}{Z_2 + Z_3}$$

Therefore,

$$i_1 = \frac{e}{Z_1 + Z_3 + \dfrac{Z_1 Z_3}{Z_2}} \qquad (6.98)$$

Similarly, the expression for i_1' through the second *T*-network is

$$i_1' = \frac{e}{Z_1' + Z_3' \left(Z_1' \, Z_3' \, / Z_2'\right)} \qquad (6.99)$$

$$i_s = \frac{e}{Z_1 + \dfrac{Z_2 Z_3}{Z_2 + Z_3}}$$

$$i_1 = \frac{i_s Z_2}{Z_2 + Z_3}; \quad i_s = e/Z_s$$

If a balance is to be obtained, $i_1 + i_1' = 0$ or

$$i_1' = \frac{e}{Z_1 + Z_3 + \left(\frac{Z_1 Z_3}{Z_1}\right)} = \frac{-e}{Z_1' + Z'_2 + \left(\frac{Z'_1 Z'_3}{Z_2}\right)} \tag{6.100}$$

Cross-multiplying and bringing LHS terms to the RHS,

$$e\left[Z_1' + Z_2' + \frac{Z_1' Z_3'}{Z_2} + Z_1 + Z_3 + \frac{Z_1 Z_3}{Z_2}\right] = 0 \quad \text{or} \tag{6.101}$$

$$\boxed{Z_1 + Z_3 + \frac{Z_1 Z_3}{Z_2} + Z_1' + Z_2' + \frac{Z_1' Z_3'}{Z_2} = 0} \tag{6.102}$$

This is the condition for balance in a twin-T network. The circuit is so named because as can be seen in the circuit, there are two T-networks. These two T-networks are in parallel; therefore, input and output terminals are common. Similar to all other AC bridges, this bridge can also be balanced by changing the supply frequency f of the AC source, in addition to changing the supply voltage or component values such as R, L, and C. However, both the phase angle and magnitude conditions must be satisfied for balance.

6.17 BRIDGED-*T* NETWORK

It is the simplified or degenerated form of a twin-T network. It has lesser number of components compared to a twin-T network. However, it is not as accurate as a twin-T network. This circuit is used to determine incremental values of inductances and L and Q of radio frequency coils. The main advantage of this circuit is that shielding problems are less and no shielding transformer is required, which is similar to a twin-T network.

Comparing with the twin-T network, Z_2' is an open circuit and $(Z_1' + Z_3') = Z_4$ in the bridged-T network (Fig. 6.28). Therefore, equation (6.101) can be modified accordingly to get the condition for balance, in the case of a bridged-T network.

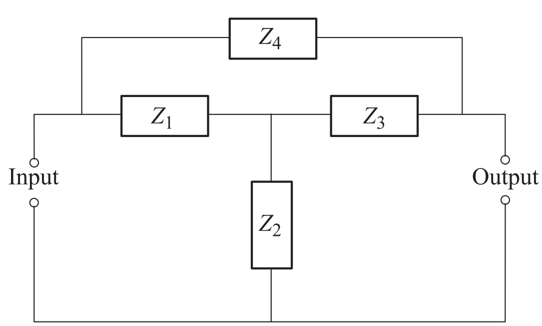


Figure 6.28 Bridged-***T*** network

$$\therefore \quad Z_2' \text{ is open and the term } \frac{Z_1' Z_3'}{Z_2'} = 0$$

$$Z_1' + Z_3' = Z_4$$

Therefore, equation (6.101) gets modified to give the general equation at balance for the bridged-T network as

$$Z_1 + Z_3 + \frac{Z_1 Z_3}{Z_2} = Z_4 \tag{6.103}$$

6.18 DETECTORS FOR AC BRIDGES

In the frequency range of 250–500 Hz, telephone receivers are customarily connected through a transformer for impedance matching. For AC bridges, telephone receivers are commonly used as detectors. The second output becomes minimum as the bridge circuit approaches a balancing condition. At lower and higher audio frequencies, when the ear is not sensitive, a Cathode Ray Tube or a CRO can be used. A tuned amplifier with an indicating device can also be used as a detector. At radio frequencies, ordinary radio receivers can be employed as detectors. The impedance of headphones will be usually in the range of a few hundred ohm to 5 kΩ. The oscillator input power required depends upon the indicator, and generally it is in the range of 50–200 mW. AC voltmeters and ammeters can also be used as detectors if they have the required sensitivity.

6.19 PHASOR DIAGRAMS

AC bridges can be analysed graphically using phasor diagrams. When the bridge is in balance, the phasor diagram is drawn. DC is time invariant. DC voltages and currents remain costant with time. Therefore, no vector or phasor diagrams are drawn for DC bridges. AC quantities vary with time. They have magnitude and phase. Therefore, AC bridges can be analysed graphically with the help of phasor diagrams. The points to note in drawing phasor diagrams are as follows:

1. When two elements of the bridge are in parallel, voltage is taken as the reference in drawing the phasor diagram.
2. If the elements of the bridge are in series, the current is taken as a reference.
3. Phasor diagram is closed by drawing the total voltage across the bridge and total current passing through the bridge.
4. Voltage across a capacitor V_c lags current through it I_c, by 90° for an ideal capacitor:

$$\because V_c = \frac{-jI_c}{\omega c}$$

5. Current through the indictor I_L lags by 90° with respect to the voltage across it for an ideal inductor:

$$\because V_L = (j\omega L)I_L$$

6. Voltage and current through a resistor are in phase:

$$\because V = IR$$

6.20 RECORDERS

6.20.1 Introduction

Recorders are required in instrumentation systems to provide a 'hard' copy or permanent record of measurement data. In particular, the recorder is useful to display a time-varying signal in a form that is easy to examine and study long after the signal has ceased to exist. Thus, the recorder serves as an alternative to a display system, which only provides an instantaneous display of measurement data. The recording may be on paper, punched cards, punched tape, magnetic tape, or in variety of formats. The indicating instruments range from simple panel meters used to measure voltage and current to oscilloscopes and many types of digital indicators. The following types of recorders are popular in instrumentation:

(a) Strip-chart recorders.
(b) Oscillographic recorders.
(c) Magnetic tape recorders.
(d) Digital recorders.

While the electrical signal is converted into mechanical motion for effecting a record in strip-chart recorders, conversion of electrical signal into magnetic remanence is the basis of magnetic tape recorders. However, recording of the signal in oscillographic recorders is performed either by mechanical or electrical means, being dependent on the configuration employed. These three recorders are useful for recording analog data, whereas magnetic recorders can also facilitate digital data recording. In addition, electromechanical recorders serve as digital recorders. The following properties have to be usually taken into account while selecting an analog recorder for a particular application or for comparing the performance of different data recording systems.

(i) Sensitivity: This is the smallest signal amplitude that can be recorded, and is governed by such factors as built-in gain of the amplifier used with the recorder, noise associated with the amplifier, and the conversion characteristic of the recording operation. High sensitivity is usually desired for a good recorder.

(ii) Frequency response: This governs the ability of the recorder to faithfully reproduce rapidly changing signals. The various time constants associated with the recording operation (e.g., mechanical time constant in the case of a strip-chart recorder) control this parameter. A good recorder is expected to possess a reasonably good frequency response, such as upto several kilohertzs.

(iii) Linearity: This is an important characteristic of the recorder, which has a direct bearing on the accuracy of the recording operation and is largely governed by the linearity of the conversion operation, e.g., conversion of electrical data into mechanical motion.

(iv) Accuracy: Linearity of the recording operation and the uniformity of movement of the recording medium (e.g., paper in the case of a strip-chart recorder) govern the accuracy of the recorder and the reproducibility of results.

6.21 STRIP-CHART RECORDERS

This is perhaps the most widely used recorder in instrumentation. In its basic form, it comprises a moving coil meter (e.g., galvanometer) with a writing pen attached to its needle. A strip of paper (sometimes also called chart paper) is pulled under the writing pen at right angles to the direction of its motion so that the time function of a signal applied to the meter (X–T recorder) can be plotted on it. Simple construction, good accuracy, frequency response from DC to a few kilohertz, and low cost are some of the advantages of such a recorder. The use of a servosystem as an integral part of the strip-chart recorder makes it possible to improve its accuracy and ruggedness. Moreover, the use of two servos to drive a single pen over the surface of a stationary chart paper results in an X–Y recorder. This recorder is useful for plotting one signal versus another in the rectilinear coordinate system. Modern X–Y recorders are characterised by good frequency response and high accuracy. Important aspects of these recorders are covered in this section.

6.21.1 Galvanometric Recorders

Basically, a galvanometer is a device capable of measuring current or voltage. Of the many types of galvanometers available to date, the moving coil (D'Arsonval type) galvanometer finds extensive use in strip-chart recorders. Here only important features of the galvanometer relating to its use in recorders are described.

6.21.2 Sensitivity

The sensitivity of a moving coil galvanometer (mcg) depends on dimensions of the coil, flux density in the air-gap, and stiffness of the spring used. Typical mcgs used in strip-chart recorders have a sensitivity of 1 mA or 1 mV for full-scale deflection.

6.21.3 Transient Response

The response of a galvanometer to a step input is very important for its use in recorders. It is dependent on such factors as resistance and inductance of the coil, galvanometer inertia (equivalent to capacitance), torsional stiffness (equivalent to inductance), and friction (equivalent to resistance). Fig. 6.29 shows the response of a typical galvanometer to step input for different values of damping. It is common practice to control the transient response of the galvanometer by changing the damping level (D). Usually D is set between 60% and 100% of the critical value in commercial strip-chart recorders. A value that provides interesting properties is 70.7% of the critical value. As can be seen from Fig. 6.29, the fast rise in the input step is considerably distorted by the galvanometer. The rise time t_r (between 10 and 90 percent points) is used as a figure of merit for the instrument. Typically t_r equals 0.35 t_n when inductance L is 0.707 of the critical value.

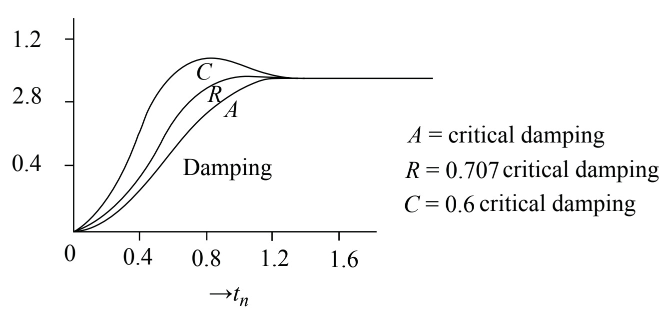


Figure 6.29 Transient response characteristics

The sinusoidal response of the recorder (both amplitude and phase characteristics) can also be used as an index of the transient characteristic following their usual equivalence. Thus, a damping factor of 0.707 of the critical value is the smallest damping for which the frequency response curve of the recorder has the maximum flatness and a linear phase lag.

6.22 PEN-DRIVING MECHANISM

The rotary motion of the galvanometer coil has to be converted into a visible graphical record. This can be done in a number of ways. Figure 6.30 shows a simple system for this purpose in which a pen is attached at the tip of the galvanometer needle. As the coil rotates under the influence of the signal current, the pen also rotates in the same manner.

The tile distance covered by the pen tip from the axis of the system (y) as a result of the coil rotation of 'q' is of the form

$$y = R \sin \theta$$

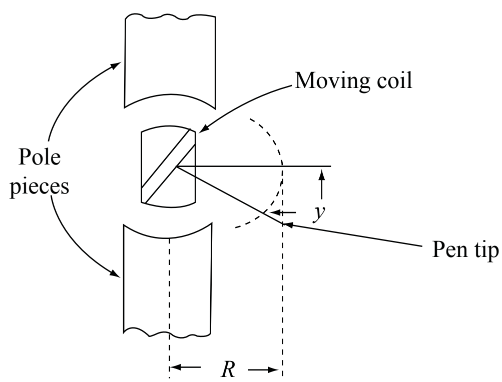


Figure 6.30 Simple pen drive associated with a galvanometer

This can be written using series expansion for the sine function as

$$y = Rq\left(1 - \frac{q^2}{6} + \cdots\right)$$

It is seen that y is directly proportional to q for small values of q. Typically, the departure from the linear law is of the order of 1% when q is 0.25 radians (rad). However, this error can be eliminated by special printing of the chart coordinates in terms of the sine function. However, this is not a common practice as the linear scale is more attractive. The other limitation of the pen-driving mechanism (Fig. 6.30) is that the time lines on the chart paper must be arcs of radius R for avoiding timing distortion of the record. Moreover, the galvanometer shaft has to be located exactly at the centre of a time-line arc.

The rectangular recording arrangement shown in Fig. 6.31 attempts to avoid some of the above limitations but it still suffers from the problem of limited accuracy of record when 'q' is high. In this case the chart paper is pulled over a sharp edge, which defines the locus of the writing pen on the paper. The writing pen or stylus is rigidly attached to the coil of the galvanometer and it is always in contact with the sharp edge of the paper as the coil rotates. The paper used here is somewhat different from that in the previous case. It can be heat sensitive, in which case the writing pen has to be equipped with a heated tip long enough to provide a hot contact with the paper. Alternatively, the paper can be

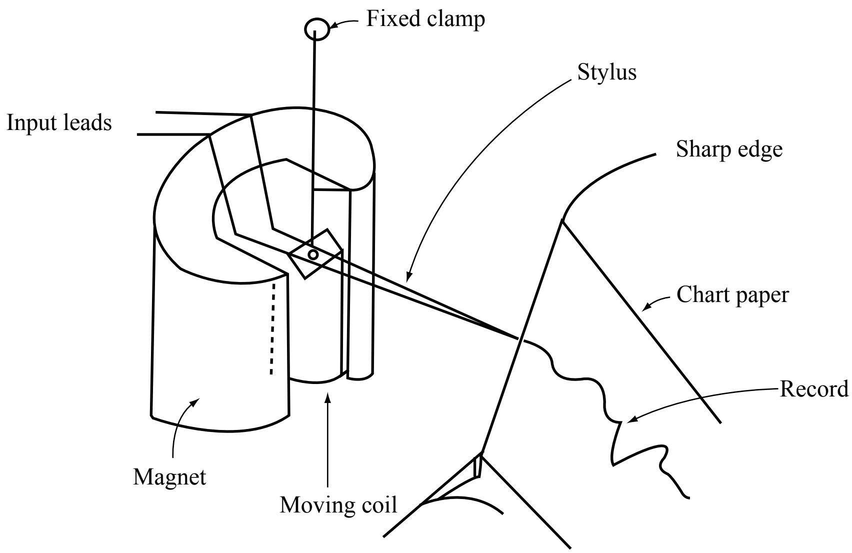


Figure 6.31 A practical rectangular recording arrangement

electrically sensitive. In this case, it is necessary that the pen tip has to carry a current into the paper at the point of contact. In either case a visible mark is left on the paper at the point of contact. The distance covered by the pen tip from the axis of the system (y) is of the form

$$y = R \tan \theta$$

where 'θ' is the angle of rotation of the coil, and R is the distance of the sharp edge from the coil centre. The above equation can be rewritten using series expansion for the tangent as

$$y = Rq\left(1 + \frac{q^2}{3} + \cdots\right)$$

This indicates that y is directly proportional to q only when 'q' is small. Typically when 'q' is limited to 0.25 radians (rad), the departure from the linear law is of the order of 2%, which is more than the case of Fig. 6.30. However, this error can also be eliminated by special printing of the chart coordinates.

Although the pen-driving mechanism of Fig. 6.31 is highly useful for recording in rectilinear coordinates, it can be applied only to heat-sensitive or electrically sensitive chart paper. Therefore, another alternative form of the pen-driving mechanism has to be used for achieving a rectilinear coordinate system for the record when using an ink pen. This is shown in Fig. 6.32. In this case, the pen tip is associated with a mechanical linkage that can convert the rotary motion of the galvanometer shaft into a straight-line motion of the pen tip, while still maintaining good proportionality between the displacement of the pen tip and the angle through which the coil is rotated. Figure 6.32 shows typical linkage for straight-line writing using a galvanometer.

As shown in Fig. 6.32, the galvanometer shaft carries a crank x, which is attached to the writing pen at a pivot. The pen length from this pivot to the writing paper is z, and the length to its rear end is y. The rear end is arranged such that l can slide along a straight line through the galvanometer shaft, which establishes the axis of the linkage. It is easy to see that 'a' is constant, when z is small and y and b are permitted to vary if z = y^2/x. In practice, the pen does not stay perpendicular to the direction of motion

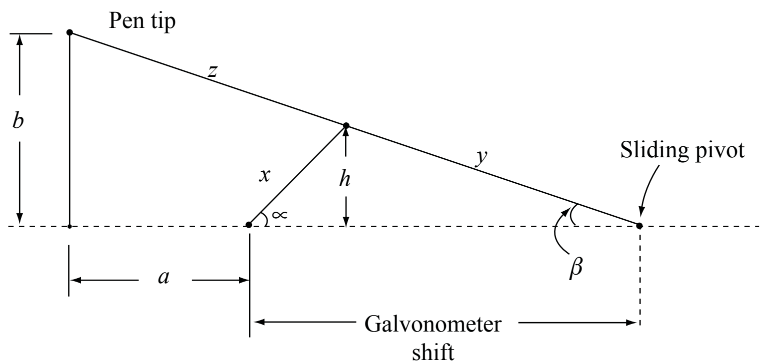


Figure 6.32 A practical rectangular recording arrangement

of the chart paper because the above assumption is not valid when 'α' assumes higher values. However, it is preferable to employ a slightly larger value of z than given in the above equation for good results.

6.23 OTHER FEATURES

Galvanometric recorders are usually associated with suitable amplifiers for improving their sensitivity, as well as for providing buffering with other circuits in the system. Such an amplifier should possess good gain stability, acceptable transient response, high input resistance, low output resistance, an adequate output voltage (and current) capability. A high gain DC amplifier with current feedback is generally preferred for this purpose. At the movement 0 the chart paper governs the time scale of the record; it is necessary that the rotational speed of the paper-driving system is maintained at a constant value. Either a spring-driven or a motor-driven system is employed for driving the chart paper. In both cases, special techniques are necessary to ensure constant speed. Different driving speeds as required (e.g., 2.5, 5, 10, 15 cm/sec, etc.) are obtained through suitable gearing techniques with the basic driving system.

Three types of writing systems are commonly used with present day recorders viz.,

(i) Ink.
(ii) Thermal.
(iii) Electrical.

The ink writing system is perhaps the most popular as disposable ink cartridges are readily available for this purpose. The writing system uses an ink that dries instantly on contact with the paper, a low-pressure ink system, and a rugged stainless steel stylus with hard tungsten carbide tips. On the other hand, the thermal writing system utilizes a hot tip, stylus that provides a trace on contact with the heat-sensitive paper. Thermal writing tips usually possess a long stylus life. However, the system requires a special chart paper that can sustain the burn-in. The electrical writing system also utilizes a chart paper on special design that is conductive, and the stylus tip is required to carry current into the paper at the point of contact for leaving a visible trace. These writing tips also possess a long stylus life.

6.24 SERVORECORDERS

As already explained, two types of servorecorders are commonly required in instrumentation, viz.,

(i) Strip-chart recorder (X–T type).
(ii) X–Y recorders.

Although different types of servosystems are used here, the closed loop null balancing servos are the most popular for both the recorders. In this case, the servomotor itself is used as the pen driver and the input circuit is usually of the potentiometer type. However, other types of input circuits are also used. Their circuitry is often dictated by special needs, such as constancy of input impedance, etc.

6.25 SERVOBALANCING POTENTIOMETRIC RECORDER

The self-balancing recorder derives its name from the popular laboratory DC instrument, viz., a potentiometer that has several advantages. They are

(i) Simplicity of construction.
(ii) High accuracy.
(iii) Sensitivity.
(iv) Excellent stability.
(v) Zero input current under null condition.

These characteristics are retained in the self-balancing potentiometric recorders. Therefore, such recorders have the capacity to handle extremely low-level signals (of the order of microvolts), and their performance is unaffected by lead lengths.

The self-balancing potentiometer (Fig. 6.33) is an electrical device capable of measuring an unknown potential by balancing it wholly or partially against a known emf, read out from the position of a contactor on a slide wire. The slide wire is connected to a DC constant current source. The resulting difference in voltage is applied as an error signal to the amplifier of the continuous balance system. This is followed by repositioning of the slide wire contactor by the balancing motor until such time as the two emfs are equal and the circuit is balanced. Much of the recorded data are generally static. Recorder characteristics that affect its performance under these conditions are

(i) Accuracy.
(ii) Linearity.
(iii) Resultability.

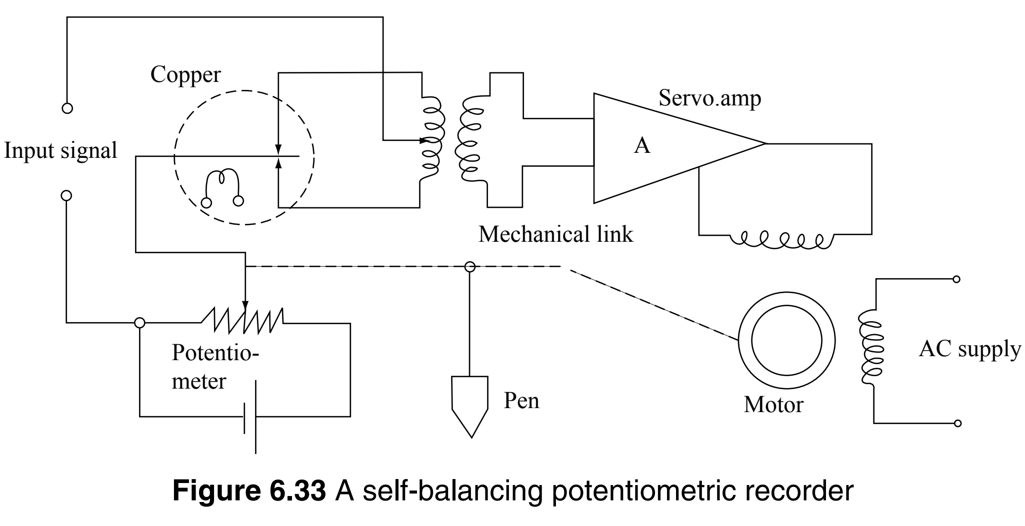


Figure 6.33 A self-balancing potentiometric recorder

Accuracy of the recorder is affected by factors such as precision of the initial calibration, range accuracy, and input to the recorder. The recorder performance in linearity is generally controlled by the slide wire linearity and the loading on the slide wire potentiometer, both of which can be taken care of easily. The parameters controlling the resettability of the recorder are deadband that is bound with the system loop gain, backlash as a result of the mechanical hysteresis between the slide wire wiper and the pen tip. Noise in the servosystem is largely due to the amplifier front end. All the three parameters can be properly adjusted to improve resettability of the recorder.

Therefore, it is easy to realize a self-balancing servorecorder suitable for static signals. However, designing such a recorder for use with a dynamic signal is somewhat difficult because of the additional characteristics such as slow speed and acceleration, which also have to be taken into account. Under dynamic conditions the servo can reach a velocity limit (slow speed) and will be unable to follow the input signal. The pen will lag the input and produces an error. In addition, the servo can reach acceleration limits beyond which it will not respond to the rate of change of the input. It is therefore necessary to see that the velocity of acceleration limits is never reached to facilitate satisfactory operation of the recorder under dynamic conditions.

Basically, the self-balancing potentiometric recorder gives an analog record of data. However, it can be modified for use with digital data handling by the addition of a shaft position encoder. It is, however, necessary that the shaft position encoder should not load the servomechanism.

X – Y recorder: The *X – Y* recorder can plot a given variable against another variable. The schematic diagram is shown in Fig. 6.34.

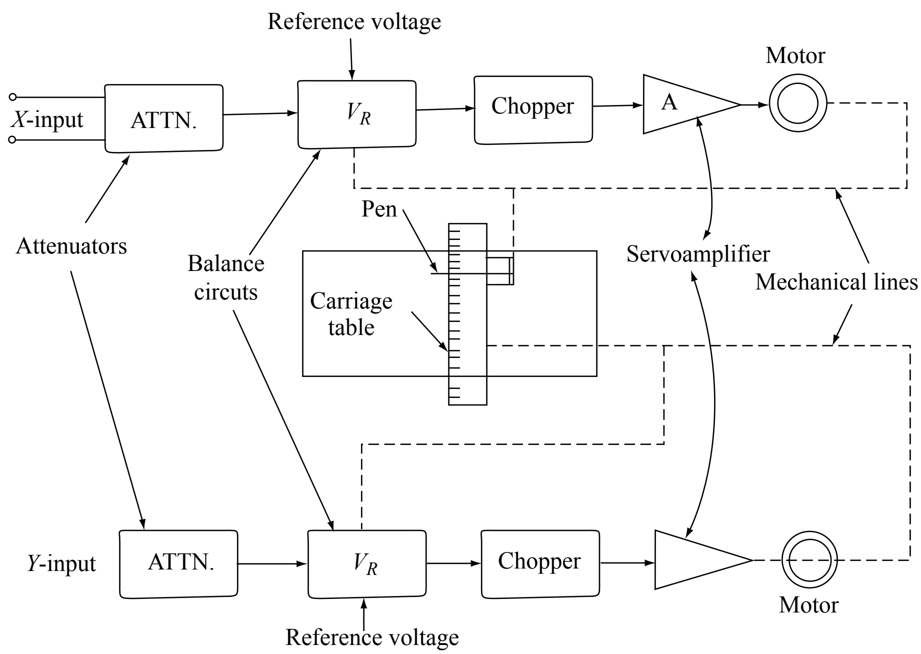


Figure 6.34 Schematic of an ***X–Y*** recorder

As shown in Fig. 6.33, the recorder has two independent but identical self-balancing servosystems. One on the horizontal *x*-axis and the other on the *y*-axis. Comparing with a strip-chart recorder, the *X–Y* recorder is almost similar with an exception that the chart (*y*-axis) is moved in response to changes in the variable rather than at a uniform rate. The resulting relative motion of the two servos produces an *X–Y* graph of the relationship on the chart paper.

As shown in Fig. 6.34, each channel of the *X–Y* recorder comprises the following blocks:

(1) Input attenuator.
(2) Balance circuit.
(3) Chopper.
(4) Servoamplifier.
(5) Servometer.

The input attenuator has the same significance as in any other measuring instruments, viz., range switching. The balance circuit compares the input signal reaching it, with an internal reference voltage and produces a DC signal that is then converted into a 50 Hz AC voltage using a chopper.

The output of the chopper is then amplified and applied to the control winding of a two-phase servomotor. As this is mechanically coupled to the balance circuit potentiometer, it changes the balance voltage initially again and nullifies the new value of the input signal. The rebalancing action is more or less continuous and changes at a rate within the limits of the recorder. Thus, the position of the potentiometer and of the pen, and the carriage to which they are coupled are always directly proportional to the amplitudes of the signals at their respective input terminals. This results in a faithful record of the *X–Y* characteristic.

Important parameters of typical servorecorders; both strip chart and *X–Y* type are given below:

6.26 CHARACTERISTICS OF TYPICAL SERVORECORDERS

(a) *X–T* type (Model 7100, M/S Hewlett-Packard, USA)

Chart width	25 cm
Chart length	36 m
Response time	< 0.5 sec
Linearity	0.1% full scale
Chart speeds	1.25 cm/sec to 30 cm/hr in different steps
Chart speed accuracy	Synchronous with line frequency
Input voltage range	I mV – 100 V full scale in different steps
Input resistance	1 M Ω
Power supply	230 V, 50 Hz AC
Writing system	Servo-actuated ink pen

(b) *X–Y* type (Model 7044 A M/S Hewlett-Packard, USA)

Writing area	38 × 25 cm
Slewing speed	750 cm/sec
Acceleration	y-axis — 2540 cm/sec^2; x-axis — 1270 cm/sec^2
Accuracy	± 0.2% full scale

Dead band	0.1% of full scale
Input voltage range	0.25 mV – 5V/cm full scale in different steps
Input resistance	1 MΩ
Type of input	Floating
Power supply	230 V, 50 Hz AC
Writing system	Servo-actuated ink pen

6.27 OSCILLOGRAPHIC RECORDERS

Oscillographic recorders are essentially *X–T* recorders and are characterized by high sensitivity, reasonably good frequency response, and ease of operation. The heart of an oscillographic (also known as optical oscillographic) recorder is the moving coil galvanometer (mcg), which is built in a miniaturized form. As shown in Fig. 6.35 a small coil is held in a uniform magnetic field by a thin suspension ribbon. A constant current flowing through the coil can produce a torque that is related to the current, the magnetic field, number of turns in the coil, the coil geometry, and its position relative to the magnetic field. A small mirror is attached to the galvanometer suspension. A light beam hits the mirror and gets reflected from it onto the recording medium, which is usually a photo-sensitive material (not shown in the figure). As the coil rotates, the reflected beam will swing through an angle θ. If the distance between the mirror and the recording medium is R, then the deflection recorded 'y' is of the form $y = R \tan \theta$. Figure 6.35 shows the galvanometer arrangement for an optical oscillograph.

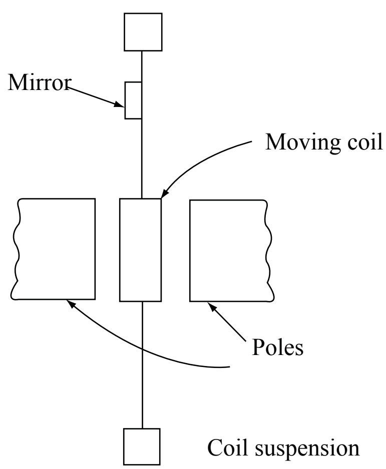


Figure 6.35 Galvanometer arrangement for an optical oscillograph

Commercially available oscillographic recorders usually employ ultra-violet light sources with associated optics to realize a fine spot beam incident on the mirror. The light reflected from the mirror is also focussed onto the recording plane. Other factors usually considered in the design of such recorders are:

(i) Accurate balance against gravity.
(ii) Damping.
(iii) Facilities for trace adjustment.

(iv) Stress-free mounting.
(v) Solid construction, etc.

The chart drive system is very similar to that in the case of the galvanometric strip-chart recorder and different driving speeds are possible. A sensitivity of the order of 1 μV/div, chart width of 50 mm, frequency response upto 125 Hz, and multichannel operation (e.g., 2, 4, etc.,) are typical applications of a commercial oscillographic recorder.

6.28 MAGNETIC TAPE RECORDERS

In data instrumentation, four main types of recording are now recognized: direct AM, FM, Pulse, and Digital recording. The basic elements of the magnetic tape recorder are the electronic encoding circuits and devices, the magnetic heads, and the tape transports. Magnetic tape recording consists of applying a magnetizing force (H) to a magnetic material called the magnetic tape (fine iron-oxide particles coated on a non-magnetic plastic film). The signals on the tape are thus in the form of permanent magnetization after the tape is removed from the source of the magnetizing force. The signal may be recorded directly or converted into other forms before recording. The electronic coding and decoding devices are part of the record and reproduce amplifiers that, respectively, prepare the input information for recording on magnetic tape as converted by the magnetic record heads. The signal is recovered from the tape by means of the reproduce head.

Magnetic tape recording has many unique advantages over the other types of recording methods:

(1) It permits wide frequency range DC to over 4 kHz, and the dynamic range is in excess of 50 db, from 100% down to 0.3%.
(2) It has low inherent distortion characteristics and is not damaged by large overloads.
(3) The signal information is preserved in its electrical form, thus facilitating automatic reduction of data.
(4) It has no processing of the finished recording unlike photographic oscillographs.

By using various multiplexing techniques, thousands of channels of information may be recorded simultaneously. The final recording may be reproduced at a different speed, thus altering the time base that results in multiplication or division of all frequencies involved. This possibility greatly simplifies the reduction of many kinds of data. Magnetic tape equipment is usually more costly than other media, but this is offset by flexibility and reusable capability of the tape.

6.28.1 Direct AM Recording

This type of recording is commonly employed for speech and music. It is the simplest of all recording processes and usually requires one tape track for each channel. The signal to be recorded is amplified, mixed with a high-frequency bias, and fed directly to the recording head as a varying electric current. This bias is introduced in order to eliminate the inherent non-linearity of the typical B-H magnetisation curve. The amplitude of the bias is several times that of the recorded signal. The frequency of the bias is not critical and is usually selected to be at least three and one-half times the highest frequency to be recorded. The bias frequency does not otherwise enter into the recording process or the subsequent playback process (Fig. 6.36).

The reproduce amplifier must have a frequency response characteristic, which is the inverse of the reproduce head characteristic (head output is proportional to the frequency of the recorded signal) to obtain an overall flat frequency response. This is called a playback equalization. The two limitations

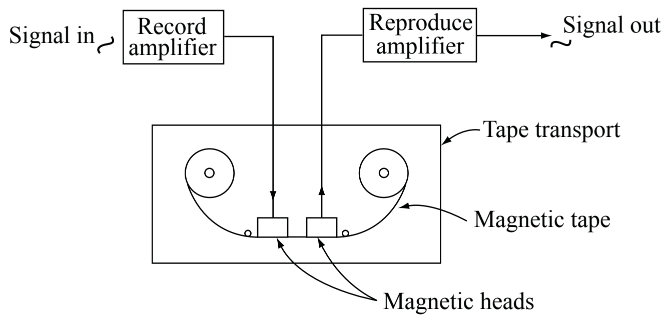


Figure 6.36 Direct recording process

of direct process are low-frequency response and amplitude instability. The lower frequency limit of the direct process is in the order of 50 Hz. This makes it unsuitable for low-frequency recording and makes DC recording impossible. Amplitude instability or signal reduction is caused due to the surface being not entirely smooth and homogeneous. This class of errors is referred to as dropouts. These are intolerable in recording accurate wave shapes of transient phenomena. The direct process has the advantage of the widest frequency spectrum, least amount of encoding and decoding equipment, and wide dynamic range. The major applications of direct process are in recording signals of a wide frequency range (not below 50 cps) and in recording voice commentary on one track of a multitrack recorder for logging and identification.

6.28.2 Frequency Modulation Recording

Frequency modulation overcomes the two basic limitations of direct recording:

(1) Inability to record low frequency.
(2) Amplitude instability caused by tape dropouts.

A particular frequency, selected as the center frequency, corresponds to a zero input signal. A DC signal causes the carrier frequency to vary in one direction. An AC signal causes the carrier frequency to vary on both sides of the center frequency. Thus, all information presented to the tape is presented in the frequency domain, and normal amplitude instabilities have little or no effect on the recording. On playback, the signal is demodulated and fed through a low-pass filter, which removes the carrier and other unwanted frequencies generated in the modulation process (Fig. 6.37).

A widespread application of FM recording is frequency division multiplexing, where a number of individual carrier frequencies are each modulated by a separate input signal. The resulting multiplicity of signals is then mixed linearly and the composite signal is recorded by using the direct recording process. Thus, the wide bandwidth and the linearity of the direct recording process are used to permit the simultaneous recording of many channels of signal information on one track of tape. A wide deviation FM has a wider dynamic range and greater overall accuracy. Here only one channel of signal is recorded on a track of tape and the entire bandwidth is used for this signal. Hence, a deviation of 40% of center frequency is possible.

The FM recording process makes stringent requirements on the tape transport to move the tape at a uniform speed. Any change in speed causes unwanted modulation of the carrier causing system noise. This is the limiting factor in the dynamic range of accuracy of the FM process. The disadvantages of the FM process are: (1) less efficient utilization of tape, additional complexity of the electronic circuitry, and (2) the requirement of constant speed of tape movement. The major applications are:

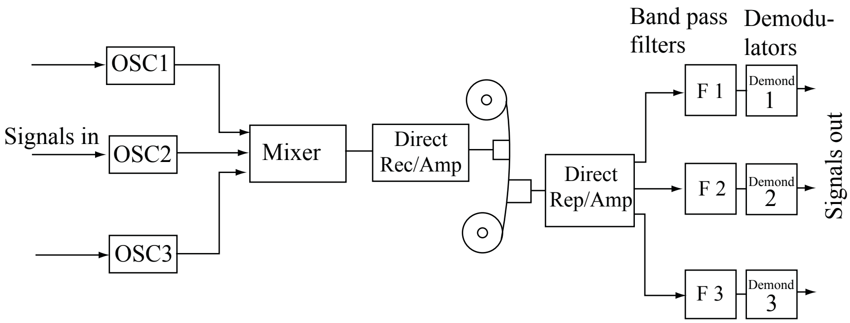


Figure 6.37 Basic frequency multiplexing narrow deviation FM system

(1) Recording of DC and low-frequency signal information.
(2) Recording of transient phenomenon where accuracy of the wave shape is important.
(3) Where data reduction by means of large changes in time base (upto 1000 to 1) are required.

Pulse duration modulation: Pulse duration modulation is a technique in which the duration of a pulse is made proportional to the amplitude of the signal to be recorded with commutation. Therefore, a large number of channels may be multiplexed. This is called time division and requires sampling a number of signal channels sequentially. If a data signal is being sampled at discrete intervals, the time between sampling intervals may be used to sample other signals.

The original waveform can be reconstructed on playback by passing the discontinuous readings through an appropriate filter. An accurate reproduction of a sine wave can be made using as few as six samples per sine wave cycle.

6.29 RECORDERS (CONTD.)

Considering analog-indicating and recording instruments capable of producing a permanent visual record of the three variations of a voltage, a classification with regard to accuracy and speed of a response is useful. There are three types of recorders.

	Recorder	Error %	Frequency response (CPS) Hzs
(1)	Self-balancing potentiometer (*X*–*Y* readers)	0.2	< 5
(2)	Galvonometer oscillographs	2	Upto 10,000 (CPS) Hzs
(3)	CRO	3	Upto 10^9

6.29.1 *X–Y* RECORDERS

X–*Y* recorders are a type of self-balancing potentiometers.

6.29.2 Self-Balancing Potentiometers

The block diagram is as shown in Fig. 6.38. Most instruments utilise AC amplifier and two-phase AC instrument servomotors. AC amplifiers are predefined because they are free from drift, and the cost is reasonable. Two phase motors have a low friction (no brushes needed) and controllability. This is used for DC, a slowly varying null-balancing method.

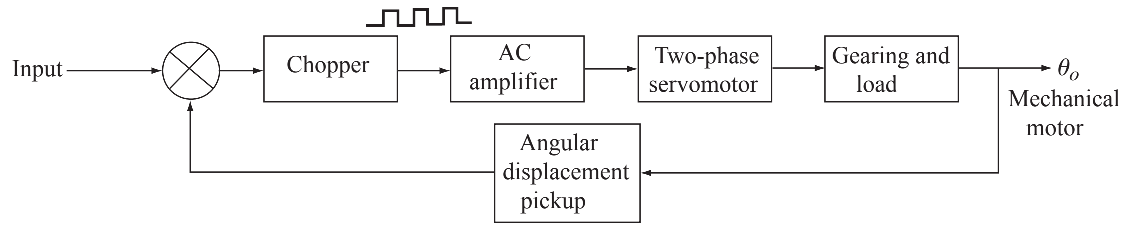


Figure 6.38 Block schematic of a self-balancing-type potentiometer

6.29.3 Working of a Servotype Motor

One of the phases is of fixed amplitude. The amplitude of the other phase that is displaced by ±90° in phase from the fixed phase, controls the directions and the amount of torque developed. When the controlled phase decreases, the torque decreases, the chopper converts, using a low-frequency signal to AC of the input voltage $e_i \neq$ reference voltage, which is converted to AC and amplified, so that it tends to drive the motor until it equals e_i. The output angle θ_0 is made for the translation of a carriage to which the pen is attached, and then the pen calibrated chart paper from a roll under the pen at a fixed and known speed to establish the *time base*. The pen will trace the variation of e_i with time. They will use a Zener diode reference supply voltage. It provides accurate and stable power supply. Full-scale input voltage may be switch selected from 0.1 to 100 mV in 19 ranges. The chart width is 25 cm with a static accuracy of ± 0.25% of the frame scale or 1 μV. The time for a full-scale pen travel is < 0.5 sec. Chart speeds vary from 1.25 cm/sec to 30 cm/hr.

6.29.4 Chopper

It uses a Photodiode or a Light-Dependent Resistor (LDR) or a photoconductor. It will have low resistance when light falls and vice versa. It acts like a switch. Light falls on the LDR from a neon bulb device by a neon oscillator. Therefore, the frequency of output is determined by the flickering of the neon bulb.

6.29.5 Servotype *X–Y* Recorders

In this type of recorder, the servorecorder plots Cartesian co-ordinate graphs from applied electrical signals. However, the graph paper is fixed in position and a pair of servomechanisms are provided to move the pen and the carriage arm. Each servomechanism is self-balancing and each is independent of the other.

The input signal, after passing through an alternator, is applied to the balance circuit where it is cancelled by an internally supplied opposing voltage. In the balanced condition, there is no error signal, and the servosystem is at a null. If the input changes in value, the resulting error signal is applied to a photoconductive chopper that converts the AC value into a corresponding 50 Hz voltage, which is amplified and applied to the servomotor. The balanced voltage will be changed till the input signal is cancelled. The frequency of the chopper is chosen as 50 Hz. If the frequency is more, the servometer may not be able to respond (Fig. 6.39).

The pen assembly in the recorder consists of a drum-type reservoir resting in a pivot mount, which moves along the carriage arm. The pen can be raised and lowered. When the slider of the potentiometer is moved till the balance circuit, the voltage cancels the input voltage. When input is zero, the output of the servometer is zero; the slider is in null position.

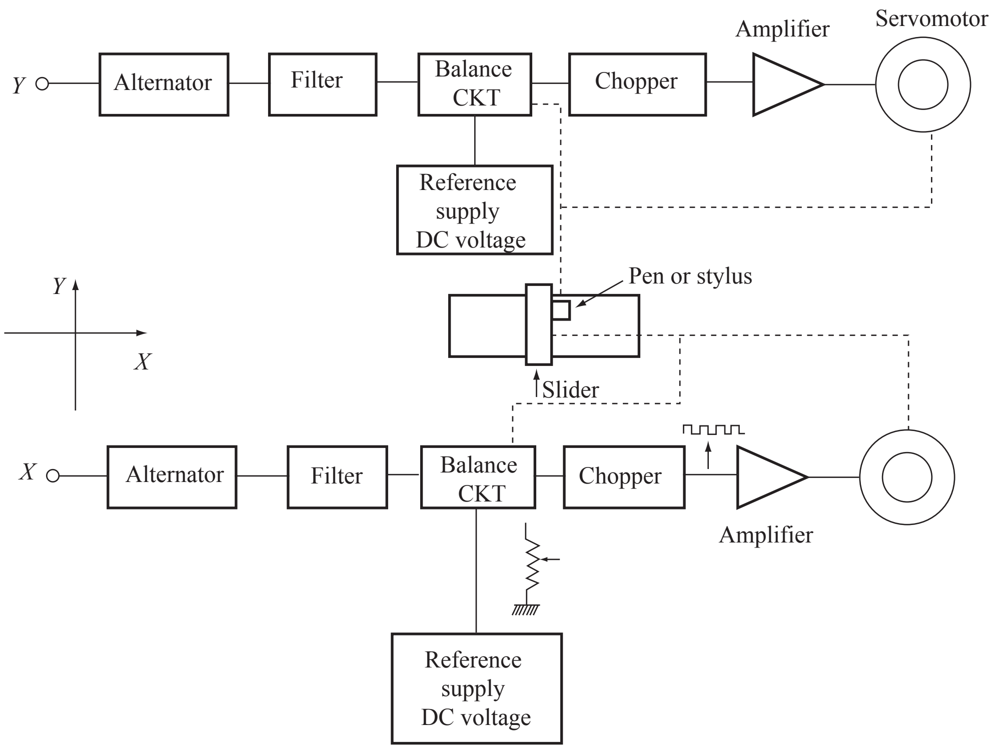


Figure 6.39 Block schematic of an *X–Y* recorder

6.29.6 *Y*-Scale

Let the selection switch be in 1 mV/cm position. The actual variable is velocity 1(m/sec); max. value = 10(mV/sec)→10 mV. Therefore, 1(m/sec)/1 mV.

Y-scale:

$$\therefore\ \frac{1(\text{m/sec})}{1\,\text{mV}} \times \frac{1\,\text{mV}}{\text{cm}} = 1(\text{m/sec})$$

6.30 GALVONOMETER OSCILLOGRAPHS

The input e_s from the source causes a current-carrying conductor in a magnetic field. Thus, the coil experiences an electromagnetic force that, since it has a lever arm, it causes a torque. This torque tends to rotate the coil until it is just balanced by the torque of the tension springs. A pen will be mounted at the end of an arm, and a chart moves at a known speed. To get straight-line motion from the rotation θ_0, speed linkages have been developed.

To realize a high-frequency response (10,000 cps), a tiny mirror is fastened to the moving coil and a light beam is reflected from it. When the coil turns, the light beam that is focused as a spot leaves a trace on the moving chart paper. The paper is a photographic paper (Fig. 6.40).

Commercially available galvanometer oscillographs employ a u.v light source and associated optics to realize a fine spot beam incident on the mirror.

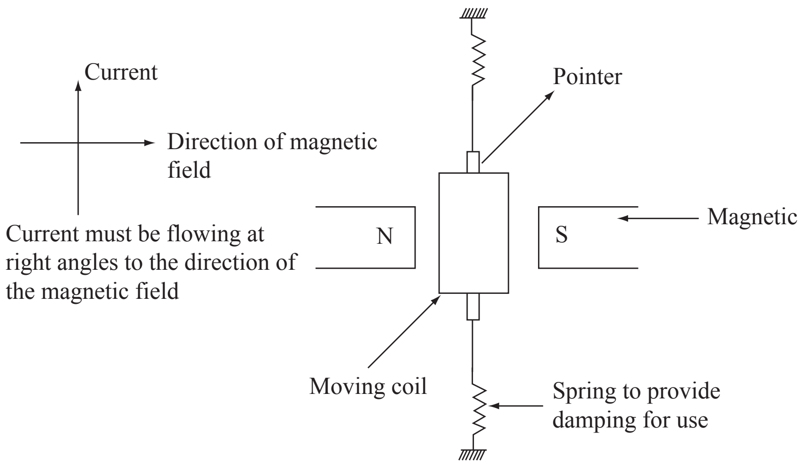


Figure 6.40 Galvanometric oscillograph

6.30.1 Applications

When a permanent record is required, *X–Y* recorders are used instead of a conventional voltmeter and an ammeter. These are used to monitor line voltage and power demands in thermal solutions. *X–Y* recorders are used in mechanical electrocardiograms, in lie detectors, in telemetry to record the incoming signal from a remote transmitter, and in UV and IR spectrophotometers. The coil is suspended by means of a thin suspension ribbon. Sensitivity= 1 μV/dv; width= 50 mm.

Example 6.3

1. An AC bridge shown in Fig. 6.41 has the following constants. Arm *AB:R* = 800 Ω in parallel with *e* = 0.4 μF; *BC:R* = 500 Ω in series with C = 1.0 μF; *CD:R* = 1.2 kΩ; *DA*: pure resistance of unknown values. Find the frequency for which the bridge is in balance and the value of *R* in arm *DA* to produce a balance.

Solution

At balance,

$$\frac{Z_1}{Z_4} = \frac{Z_2}{Z_3}$$

Therefore,

$$Z_3 = \frac{Z_2 Z_4}{Z_1} = Z_2 Z_4 Y_1$$

$$Y_1 = \frac{1}{800} + j\omega\,(0.4 \times 10^{-6})$$

$$Z_2 = 500 - \frac{j}{\omega \times 1 \times 10^{-6}}$$

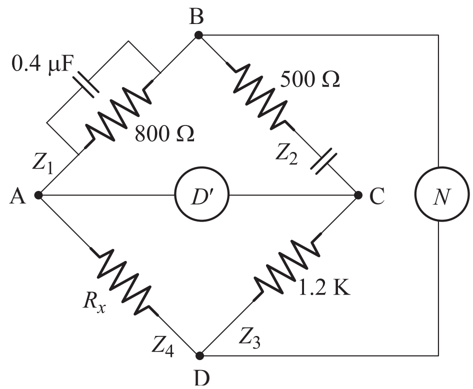


Figure 6.41 For Example 6.3

$$Z_3 = 1200\ \Omega$$
$$Z_4 = ?$$
$$1200 = 500 - \frac{j}{\omega \times 10^{-6}}\ R_x\left(\frac{1}{800} + j\omega 0.4 \times 10^{-6}\right)$$

By equating real parts,

$$R_x = 1170\ \Omega$$

By equating imaginary parts,

$$f = 398\ \text{Hz}$$

Example 6.4

For the circuit shown in Fig. 6.42, obtain the expressions for the unknowns R_s and L_s.

Solution

It is a bridged-T network. At balance,

$$Z_1 + Z_3 + \frac{Z_1 Z_3}{Z_2} + Z_4 = 0$$

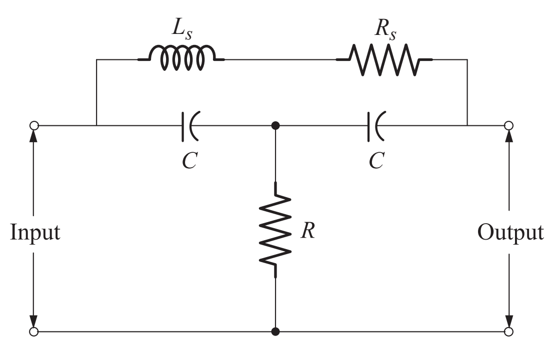


Figure 6.42 For Example 6.4

$$Z_1 = \frac{1}{j\omega C}$$

$$Z_3 = \frac{1}{j\omega C}$$

$$Z_2 = R$$

$$Z_4 = R_s + j\omega L_s$$

Substituting these values in the equations, equating real and imaginary parts, and simplifying,

$$\frac{1}{j\omega C} + \frac{1}{j\omega C} - \frac{1}{\omega^2 C^2 R} + R_s + j\omega L_s = 0$$

Therefore,

$$R_s = \frac{1}{R(\omega C)^2} \quad \omega L_s = \frac{2}{\omega C}$$

Example 6.5

For the circuit shown in Fig. 6.43, obtain the expression for L_p and R_p of the coil.

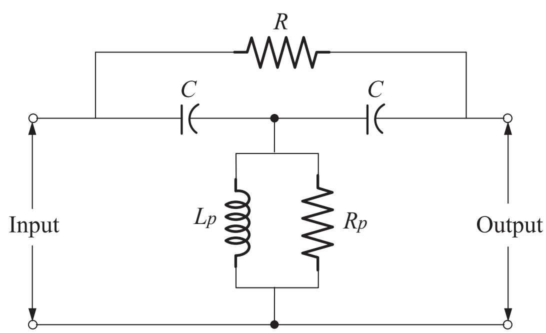


Figure 6.43 For Example 6.5

Solution

This is also a bridged-T network. This circuit is used to compare different coils, L_p and R_p. Using the general equation for a bridged-T network at balance,

$$Z_1 + Z_3 + \frac{Z_1 Z_3}{Z_2} + Z_4 = 0$$

$$\frac{1}{j\omega C} + \frac{1}{j\omega C} - \frac{L_P\left(R_P + \dfrac{1}{j\omega L_P}\right)}{R_P} + R = 0$$

Simplifying the above,

$$\omega L_p = \frac{1}{2\omega C}$$

$$R_p = \frac{1}{R(\omega C)^2}$$

Example 6.6

For the given R.LC. bridge, find the expression for capacitance, resistance, and inductance (Fig. 6.44).

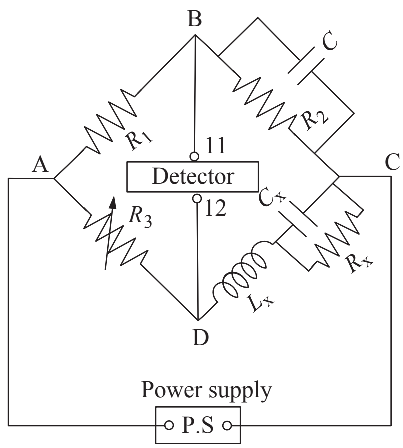


Figure 6.44 For Example 6.6

Solution

(1) Resistance, $R_x = \dfrac{R_2 R_3}{R_1}$.

(2) Inductance, $L_x = R_2 R_3 C$.

(3) Capacitance, $C_x = \dfrac{CR_3}{R_1}$.

6.31 SUMMARY

In this chapter, the Wheatstone bridge circuit and the derivation for R_x at balance are given. The Thevenin equivalent circuit and the derivation of the expression for the imbalance current considering the galvanometer resistance are given. The limitations of the Wheatstone bridge are mentioned. The significance of the Kelvin bridge circuit and the derivation for the unknown resistance R_x are given. Applications of these bridge circuits are mentioned. A typical strain gauge bridge circuit and the derivation for e_o at imbalance are given.

The AC bridges are difficult to balance than DC bridges. Different names have been given to the bridges after the inventors. There are many types of AC bridges, each having some salient features. The student is expected to analyse the bridge circuits clearly and derive the expressions for the unknown elements. AC bridges can also be balanced by varying the input signal frequency. Using these circuits, other characteristics of passive elements such as quality factor Q of the coils, dissipation factor D of the capacitors, loss angle, leakage, etc., can also be estimated. The source for AC bridges is usually an oscillator with the required input power. Shielding and grounding problems cause difficulty in balancing AC bridges. Twin-T and bridged-T networks overcome these problems to a certain extent. AC bridge measurements are also affected by the variation in supply frequency f and input e.

Points to Remember

- DC bridges are used to measure only resistance.
- AC bridges are used to measure R, L, C, or f of AC excitation input. The Q factor of a capacitor can also be determined using AC bridges.
- Wheat stone bridge is used to measure resistance in the range of 1Ω to MΩ.
- Kelvin's bridge is used to measure very low resistance in the range of 1 to 0.00001Ω. Therefore, using this bridge conductance of connecting wires used in circuits, conductors, patch cards, etc. can be measured. It is also called Kelvin's double bridge as two more arms are there in the bridge, in addition to the conventional four arms.
- For balancing AC bridges, both magnitude and phase angle conditions are to be satisfied. Therefore, balancing of AC bridges is more difficult than DC bridges.
- Maxwell's bridge is used to determine the value of inductors (coils) with the Q factor lying in the range 1–10.
- Hay Bridge is used to measure 'L' of coils with Q factor > 10. This bridge is also known as the opposite angle bridge since C and L elements are involved, one is a leading angle element and the other is a lagging angle one.
- Schering Bridge is used to measure C, particularly for insulators with a phase angle of nearly 90°.
- Dissipation factor D of a capacitor gives an indication of how close the phase angle is between I_c and V_c. The current through the capacitor and the voltage V_c across it must be at 90° to each other in an ideal case. Quality factor of a coil or inductor indicates the ratio of the energy restored in the coil, to energy dissipated in it. The higher the 'Q' factor, the better it is because the coil should restore energy in it without dissipation in the ideal case. Therefore, the ideal value of a Q factor is 8.
- Wien Bridge can be used to determine the frequency of AC input excitation to the bridge. In this circuit, there is a lead – lag network.
- Resonance bridge can also be used to determine frequency of AC input.
- Similar angle bridge is so called because it has only capacitive reactive elements. This is also known as the capacitance comparison bridge.

Objective-type Questions

1. The two types of DC bridges are __________.
2. Balancing of __________ bridges is more difficult than __________ bridges.
3. For the measurement of finite resistance associated with conductors __________ bridge is used.
4. The range of resistance that can be measured using Kelvin's bridge is __________.
5. Balancing of __________ bridges is difficult because __________.
6. Maxwell bridge is used to measure __________.
7. Hay bridge is used to measure __________.
8. The bridge circuit that is preferred to measure the frequency is __________.
9. The effect of strong capacitance across the detector can be eliminated by __________.
10. Hay bridge is also known as __________.
11. The disadvantage of Maxwell's Bridge is __________.
12. Anderson bridge is used for the measurement of __________.

13. Anderson bridge is a modification of ______.
14. The electric loss in an imperfect capacitor is given as ______.
15. Inductor's inductance is measured in terms of capacitance and resistance by ______.
16. The bridge suitable for the measurement of inductance of coils with $Q > 10$ is ______.
17. In a Schering bridge the potential of the detector above the earth potential is of the order of ______.
18. The most commonly used null detector in power frequency AC bridge is a ______.
19. The bridge suitable for the measurement of capacitance at high voltage is ______.
20. Wagner's ground connection is used in AC bridges for ______.

Review Questions

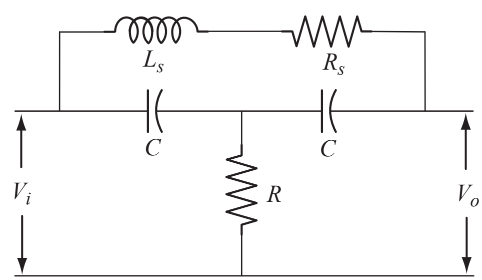
1. What is the significance of bridge circuit measurements over direct meter measurements?
2. Draw the circuit for a Wheatstone bridge and derive the expression for current through the galvanometer due to imbalance in the bridge circuit.
3. What are the limitations of a Wheatstone bridge? Mention its applications?
4. Explain how the Kelvin bridge overcomes the limitations of the Wheatstone bridge. Why do you call it a double bridge?
5. Derive the expression for the unknown resistance R_X in the case of a Kelvin double bridge. Mention the applications of a Kelvin bridge.
6. Compare AC and DC bridges.
7. Draw the circuits for a bridge that can be used to determine L of coils whose Q factor is in the range 1–10. Justify the type of bridge chosen and derive the equation for L_x and R_x at balance.
8. How do you determine the dissipation factor D of a given capacitor using a bridge circuit? Derive the expressions used.
9. What are the salient features of twin-T and bridged-T networks? Derive the necessary expressions.
10. What are the different types of detectors used for AC bridges?

Unsolved Problems

6.1 The description of an AC bridge is as given: Arm AB: R_1 = 1 kΩ; C_1 = 0.2 μF in parallel; Arm BC = R_2 = 800 Ω; C_2 = 0.6 μF in series. Arm CD: R_3 = 1.6 kΩ; Arm DA: R_4 = 1.2 kΩ. Determine the frequency at which the bridge will be balanced.

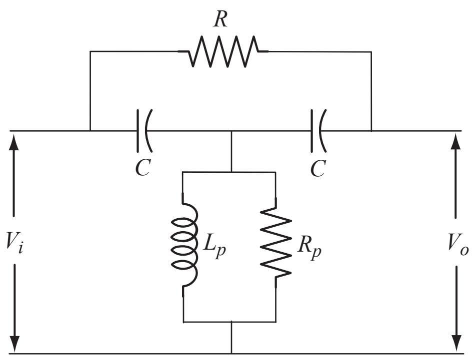
6.2 In the case of an AC bridge, arm AB has a resistance in parallel with a capacitor. The values are R_1 = 1.2 kΩ; C_1 = 0.25 μF. Arm BC has a resistance in series with a capacitor with values of R_2 = 1.0 kΩ; C_2 = 0.5 μF. Arm CD is a pure resistance arm with R_3 = 2 kΩ. Determine the values of the components in arm DA if the bridge is balanced at a frequency of 500 Hz.

6.3 For the circuit shown, determine the values of L_S and R_S. Given, C = 0.1 μF; R = 1.2 kΩ; f = 300 Hz.



6.4 A bridged T- network is at balance at $f = 500$ Hz. Determine the values of unknown inductor L_P and in series resistance R_p if the values of $C = 0.18\ \mu F$ and $R = 2.7\ k\Omega$.

6.5 For the circuit shown, determine the values of R_P and L_P if the bridge is balanced at $f = 1$ kHz. Given $C = 0.1\ \mu F$; $R = 1.2\ k\Omega$.



6.6 A bridged T-network is at balance at $f = 2$ kHz. The values of R and C are 4.7 kΩ and 0.9 μF, respectively. Determine the value of unknown parallel inductor L_P and its shunt resistance R_P.

6.7 In a Hay bridge, the component values with usual notations are, $R_1 = 4.7\ k\Omega$; $C_1 = 0.1\ \mu F$; $R_2 = 10\ k\Omega$; $R_3 = 6.8\ k\Omega$ at balance; $f = 200$ Hz. Determine the value of L_x and its associated resistance R_x.

6.8 In a Schering bridge, the component values with usual notations are, $R_1 = 10\ k\Omega$; $C_1 = 0.1\ \mu F$; $R_2 = 12\ k\Omega$; $C_3 = 0.2\ \mu F$ at balance. Determine the value of C_x and R_x if the AC input given has a frequency of 150 Hz.

7

Transducers

Introduction • Classification of transducers • Active and passive transducers • Force and displacement transducers • Resistance strain gauges • Bonded-type strain gauge • Summary

7.1 INTRODUCTION

Instrumentation systems generally consist of three major elements:

1. An input device.
2. A signal conditioning device.
3. An output device.

The input device receives the quantity under measurement and delivers a proportional electrical signal to the signal conditioning device. Here the signal is amplified, filtered, or otherwise modified to a format, acceptable to the output device. The output device can be a Cathode Ray Oscilloscope (CRO) or a recorder.

The input quantity for most instrumentation systems is non-electrical. However, in order to use electrical methods for measurement, manipulation, and control, non-electrical quantity is generally converted into electrical quantity by a device called a *transducer*.

The definition of a transducer can be given as a device, which when actuated by energy in one form, supplies energy either in the same form or in another form to a second transmission system.

7.1.1 Examples

1. A mechanical force or displacement being converted into an electrical signal.
2. A thermistor reacts to temperature variations.
3. A photo cell changes in light intensity.
4. Measurement of electrical noise.
5. Telemetering system—when input and output are in electrical form.

7.2 CLASSIFICATION OF TRANSDUCERS

Transducers may be classified according to their application, method of energy conversion, nature of output signal, etc. However, a sharp distinction in the classification is difficult and often results in overlapping.

Transducers can be classified as:

1. Active transducers.
2. Passive transducers.

The R, L, and C transducers are passive. They need external power supply. However, *active transducers do not need any external power*. They generate analog voltage or current by themselves when the input is given. Passive transducers can further be classified into (Table 7.1):

1. Resistance transducers.
2. Capacitance transducers.
3. Inductance transducers.

Table 7.1 Types of transducers

Passive Transducer	Principle	Parameter to be Measured
Resistance transducer		
1. Potentiometric	Positioning of the slider by an external force results in variation of resistance	Pressure, force, displacement
2. Resistance strain	Resistance is changed by an elongation torque gauge or compression due to externally applied stress	Force, torque
3. Pirani gauge	Resistance of a heating element is varied by convection cooling of a stream of gas	Gas flow, gas pressure
4. Resistance thermocouple	Resistance of a pure metal with a large positive temperature coefficient varies with temperature	Temperature measurement, radiation, heat
5. Thermistor	Resistance of certain metal oxides with a negative temperature coefficient varies with temperature	Temperature
6. Photoconductive cell	Resistance varies with light intensity	Light intensity
Capacitance transducers		
1. Variable capacitance pressure gauge	Distance between two parallel plates is varied by an externally applied force	Displacement, pressure
2. Capacitor microphone	Sound pressure varies the capacitance between a fixed plate and a movable diaphragm	Speech, music
3. Dielectric gauge	Variation in C by changing to a dielectric constant	Liquid level
Inductance transducers		
1. Magnetic circuit transducer	Self-induction or H of an excited transducer coil is varied by change in magnetic circuit	Pressure, displacement
2. Differential voltage	The differential voltage of two secondary windings of a transformer is varied by positioning the magnetic core	Pressure, displacement
3. Magnetoelastic transducer	Magnetic properties are varied by pressure, and stress changes the value of permeability μ	Force, pressure, sound
4. Magnetostrictive effect	Change in dimensions due to magnetisation	To produce ultrasonic waves

7.3 ACTIVE AND PASSIVE TRANSDUCERS

Table 7.2 Active transducers

Active Transducer	Principle	Parameter to be Measured
1. Thermocouple	An *emf* is generated across the junction	Temperature
2. Photovoltaic cell	Voltage is generated	Light meter solar cell
3. Piezoelectric pick up	An *emf* is generated when an external force is applied to certain crystalline materials like quartz	Vibration sound force

7.3.1 Factors That Affect the Performance of a Transducer

Table 7.3 Factors affecting transducer performance

1. Physical conditions	Mounting provisions, corrosion resistance, mechanical and electrical connections
2. Environmental conditions	Temperature, effects, shock, vibration
3. Compatibility of associated equipment	Impedance matching, insulation resistance, zero balance, etc.
4. Transducer parameters	Type and range of measurand, sensitivity, excitation, etc.

7.3.2 Applications

Transducers can be used to convert mechanical and other physical parameters such as temperature, pressure, force, stress, strain, flow, vibrations, velocity, etc. into electrical signal form. Then the signal can be amplified, modulated, and transmitted, or signal conditioning in any form can be done. Therefore, transducers have intensive usage in many industries such as fertiliser plants, steel plants, thermal power stations, petrochemical industries, and textile mills.

7.4 FORCE AND DISPLACEMENT TRANSDUCERS

7.4.1 Potentiometer

Basic principle: The resistance of a wire is given by the formula

$$R = \frac{\rho l}{A}, \tag{7.1}$$

where ρ is the specific resistance of the material, L the length of the wire, and A the cross-sectional area of the wire. Any stimulus or measurand that charges or affects either L or A or ρ will change the value of resistance. Hence, any variation of that stimulus or measurand can be suitably converted into a variation of an electrical quantity, say voltage. The resistance element can be a sensor for a transducer system.

Some of the quantities that affect the resistance value are temperature, strain, pressure, etc. In addition to these quantities, a slider movement can change the length of the wire under consideration.

As a consequence, displacement—translational as well as angular—can be converted into a variation of resistance.

7.4.2 Potentiometric Transducer

A resistive potentiometer is a resistance element provided with a sliding contact and an auxiliary excitation source, either DC or AC. The motion of the slider can be translational, rotational, or a combination of these two such as helical potentiometer, thus permitting measurement of rotary or translatory displacements. Existing potentiometers can measure translatory motion from about 0.1 to 50 cm and rotary motion from about 10° to 60 full turns.

Resistance elements in common use may be classified as follows:

1. Wire wound.
2. Carbon film.
3. Conducting plastics.

The resistance between the output terminals (Fig. 7.1) is proportional to the displacement.

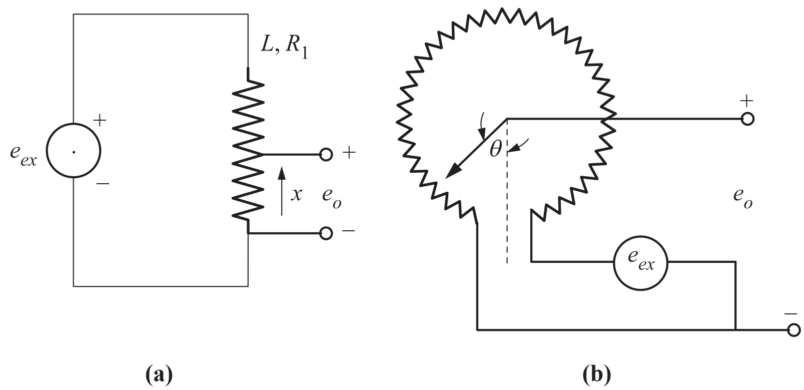


Figure 7.1 Potentiometric transducer: (a) linear type and (b) angular type

$$R_x = \frac{X}{Y} R_t \qquad \alpha X \because \quad R_t \text{ and } L \text{ are constants} \tag{7.2}$$

$$C_o = \frac{R_x}{R_t} e_{ex} \qquad \alpha X \text{ as } R_t \text{ and } e_{ex} \text{ are constants} \tag{7.3}$$

Thus, output voltage is proportional to displacement. If excitation is sinusoidal, the output will also be sinusoidal with no phase shift. The magnitude will be proportional to the displacement. In Fig. 7.1, the output voltage $e_o = \frac{\theta}{360^o} e_{ex}$ as this is a 330º potentiometer.

7.4.3 Loading Effect on a Potentiometer

The output of the potentiometer is normally connected to a meter or a recorder that draws some current from the potentiometer. The practical situation is shown in Fig. 7.2.

The resistance across the output terminals is given by

$$\frac{1}{R_o} = \frac{1}{R_m} + \frac{1}{(X_i / X_t) R_p} \tag{7.4}$$

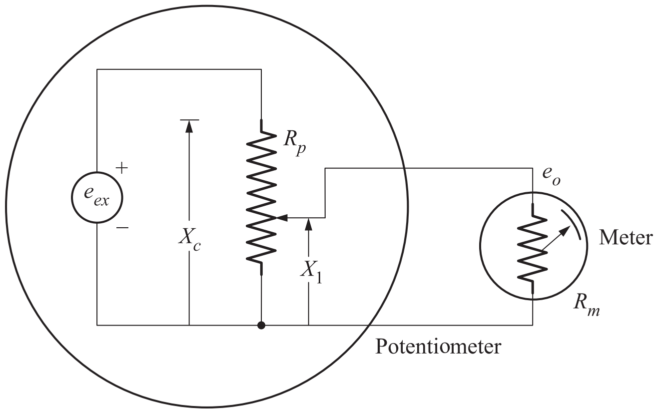


Figure 7.2 Loading effect on a potentiometric transducer

Therefore,

$$R_o = \frac{R_m R_p (X_i / X_t)}{R_m + (X_i / X_t) R_p} \tag{7.5}$$

$$e_o = \frac{R_o e_{ex}}{R_p [1 - (X_i / X_t)] + R_o} \tag{7.6}$$

After a few manipulations,

$$\frac{e_o}{e_{ex}} = \frac{R_m (X_i / x_t)}{R_m + R_p (X_i / X_t) - R_p (X_i / X_t)^2} \tag{7.7}$$

$$= \frac{1}{\frac{1}{(X_i / X_t)} = \frac{R_p}{R_m}[1 - (X_i / X_t)]} \tag{7.8}$$

Now when (R_p / R_m) becomes very small and near zero, then the second term in the denominator can be omitted when compared with the first term. Then

$$\frac{e_o}{e_{ex}} = \frac{x_i}{x_t}$$

Actually the output is given by the above equation if there is no loading. In reality (R_p / R_m) is not very small. Hence, e_o becomes a non-linear function of displacement x_i. This fact is well exhibited in Fig. 7.3.

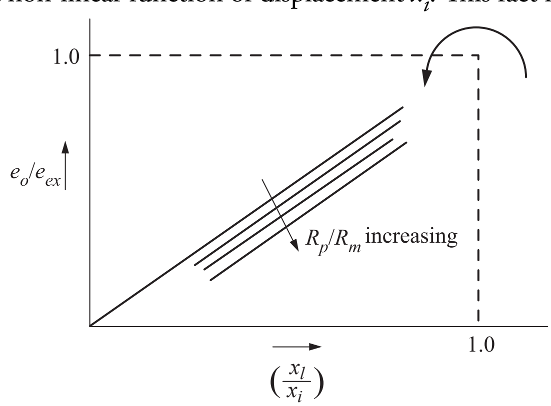


Figure 7.3 Output variation for potentiometric transducer

However, for values of $(R_p/R_m) < 0.1$ the maximum error is approximately 15 (R_p/R_m) percentage of full-scale reading.

This fact is proved here:

$$\frac{e_o}{e_{ex}} = \frac{1}{(1/\alpha)+(R_p/R_m)(1-\alpha)} \quad \text{where } \alpha = (X_i/X_t) \tag{7.9}$$

$$= \text{Percent error} = 100\left[\alpha - \frac{1}{(1/\alpha)+(R_p/R_m)(1-\alpha)}\right] \tag{7.10}$$

$$= 100\left[\frac{\alpha(1/\alpha)+(R_p/R_m)(1-\alpha)-1}{(1/\alpha)+(R_p/R_m)(1-\alpha)}\right] \tag{7.11}$$

$$= 100\left[\frac{(R_p/R_m)\alpha^2(1-\alpha)}{M(R_p/R_m)(1-\alpha)\alpha}\right] \tag{7.12}$$

Now α is always ≤ 1 and $(R_p/R_m) \leq 0.1$ as per the assumption. Therefore, the above equation can be written as

$$\text{Percentage error} = 100\,(R_p - R_m)\alpha^2(1-\alpha) \tag{7.13}$$

$$\frac{d\in}{d\alpha} = 100\,(R_p/R_m)(2\alpha - 3\alpha^2) = 0 \tag{7.14}$$

i.e.,

$$\alpha = 2/3 \quad \text{or} \quad 0$$

$$\alpha = 0 \text{ corresponds to the minimum error point}$$

$$\alpha = 2/3 \text{ corresponds to the maximum error point}$$

$$\text{The maximum percentage error} = 100\,R_pR_m\,\frac{(2)^2}{3}\left(1-\frac{2}{3}\right) \tag{7.15}$$

$$= \frac{400}{27}(R_p/R_m) \tag{7.16}$$

$$\boxed{\text{Maximum percentage error} = 15\,R_p/R_m} \tag{7.17}$$

The above equation tells us that if the error due to loading depends on the ratio of the potentiometer to meter resistance, the linearity will be better. On the other hand, a meter with a resistance R_m lower than the potentiometer resistance also gives better linearity. However, the potentiometer resistance cannot be lowered much as it will reduce the sensitivity of measurement. Sensitivity is defined as the change in the output for a change in input. Here, in the potentiometer $e_o = e_{ex}\,(X_i/X_t)$. Hence, sensitivity is (e_{ex}/e_o). One may think that sensitivity can be increased by just increasing the excitation voltage. However, this is not possible since potentiometers have definite power ratings. The power rating depends on the size, material used, and the configuration, and not on the actual resistance value. The power rating is actually the heat that can be dissipated by the potentiometer. Therefore, the power that can be supplied to the potentiometer is fixed depending on the heat dissipation capacity. Normally manufacturers design a series of potentiometers with a total resistance R_p ranging from 100 to 100,000 Ω. However, all of them may be 5 cm in diameter with the same mechanical configuration and hence have the same heat transfer capability. Therefore, they have the same power rating, say 5 W at 25°C ambience. In addition, this power rating will fix the maximum excitation voltage for R_p.

Let P be the power rating of a potentiometer in watts and R_p the total resistance of the potentiometer. Then

$$P = (\text{Max } e_{ex})^2/R_p \tag{7.18}$$

$$\boxed{\text{Max } e_{ex} = \sqrt{PR_p}} \tag{7.19}$$

Therefore, for a given meter with resistance R_m, if low R_p is chosen to have a better linearity then the maximum excitation will also be low and hence will have a lower sensitivity.

Thus, the choice of R_p is a trade-off between loading and sensitivity considerations. The maximum sensitivity achievable varies considerably from type to type and also with the size of potentiometers. If the size is increased, then the heat dissipation can be increased but the stroke length will also have to be increased. Therefore, the sensitivity cannot be increased easily by increasing the size alone. Suitable design and proper choice of material have to be used to increase the sensitivity. In fact, the shorter stroke devices generally have higher sensitivity.

Extreme values for sector potentiometers are of the magnitude of 15 V/degree and 120 V/cm for potentiometers with a short stroke of about 0.6 cm translational pots. However, these have a very low sensitivity.

Example 7.1

The potentiometer output is to be connected to a recorder of 10 kΩ input resistance. Non-linearity is to be held within 1%. A family of potentiometers having a power rating of 5 W and resistance ranging from 10 to 10,000 Ω in a 50 Ω step is available. Choose from this family of potentiometers that will give the greatest sensitivity. If the potentiometer is of single-turn 360° units, what is the sensitivity?

Solution

Max. error in linearity – Non-linearity = 1%

Therefore, R_p/R_m should be less than 0.1.

If $R_p/R_m \leq 0.1$,

$$\begin{aligned}
\%\ \text{error} &= 15(R_p/R_m) \\
15(R_p/R_m) &= 1 \\
(R_p/R_m) &= \frac{1}{15} \\
R_p &= \frac{1}{15} \times R_m = \ = 666.6
\end{aligned}$$

Therefore, we can choose a potentiometer with a total resistance R_p = 666.6 Ω at the maximum. Any value of R_p less than 666.6 would be all right as far as the non-linearity specification is concerned. However, lower the value of R_p, lower will be the sensitivity. Therefore, we choose 650 Ω potentiometer from the family, which will have maximum sensitivity and at the same time have non-linearity less than 10%:

$$\text{Max } e_{ex} = \sqrt{5 \times 650} = 57$$

$$\text{Sensitivity for } 360° = \frac{57}{360} = 0.16 \text{ V/degree}$$

7.4.4 Resolution

The resolution of a potentiometer is the smallest change in displacement that can be measured or identified. If the excitation is fixed, then it is the smallest change in resistance that can be obtained by slider movement. This factor strongly depends on the construction of the resistance element. To get a high resolution, a single slide-wire can be used as the resistance element of the potentiometer. This will give a continuous stepless resistance variation and hence a very fine resolution. Here the accuracy is limited by other components of the system. These types of potentiometers are available but limited to small resistance values because the length of the wire depends on the desired stroke in a translational device and on the space restrictions in rotational devices. The resistance of a given length of wire can be increased by decreasing the diameter of the wire. However, the lesser the diameter, lesser will be the strength and will get worn out quickly.

The resistance variation for such potentiometers is not linear but is a continuous step change, as shown in Fig. 7.4.

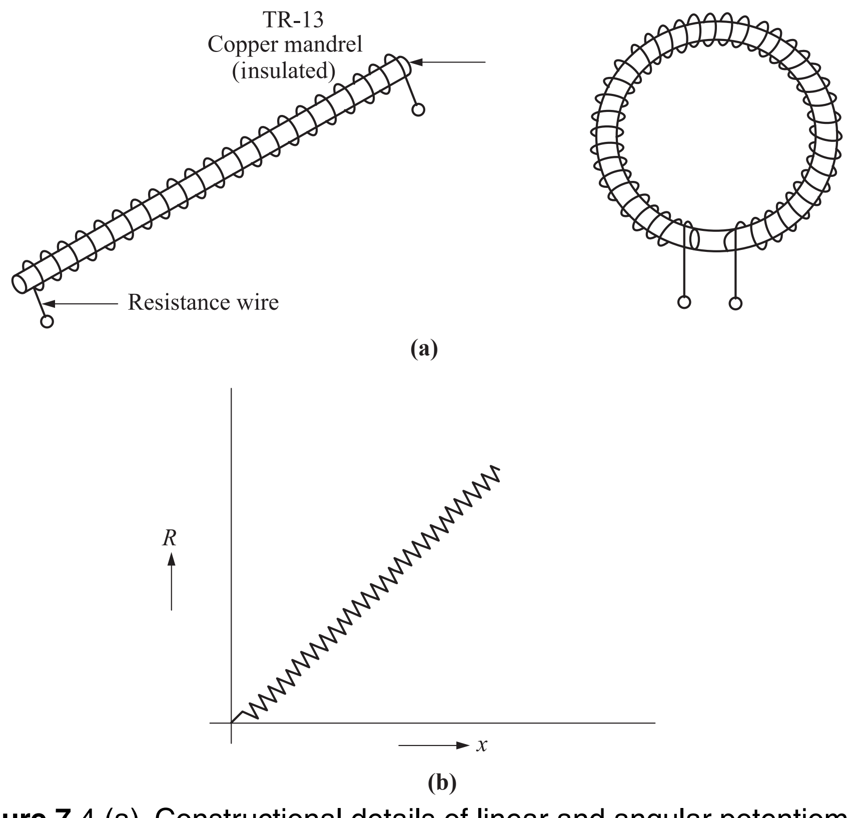


Figure 7.4 (a) Constructional details of linear and angular potentiometers and (b) Output variation for potentiometric transducer

To get fairly high resistance values in a small space, the wire-wound resistance element is made. The resistance wire is wound on a mandrel or cord, which is then formed into a circle or helix.

This means that the resolution is limited here depending on the resistance wire size. For example, if a translational device has 100 turns of resistance wire on a card of 1 cm long, motion changes smaller than 0.01 cm cannot be detected. The practical limit for wire spacing at present times is between 200 and 400 turns per cm. Therefore, for translational devices, the resolution is limited to 0.0025 – 0.005 cm, while a single turn rotational device, on the basis of 400 turns per cm, has the best angular resolution = $360/400\,\pi D = 0.29/D$, where D is in cm.

The resolution is related to total resistance and fine wire is required to get close wire spacing as fine wire has a high resistance. A carbon film or conductive plastic resistance is needed. This carbon film element may have a resolution as fine as 12×10^{-6} cm. However, the overall resolution is a bit low because of mechanical defects in bearings and wiper springs. In carbon film devices, the wiper has a relatively high resistance. Therefore, the amount of current drawn from the potentiometer must be kept quite low.

Another way to improve the resolution is to use multiturn potentiometers. Here the resistance element is in the form of a helix and the wiper travels along a lead screw. However, the number of wires per cm is still limited. The increase in resolution is obtained with the help of gears. The gears make the potentiometer shaft to go through a number of rotations when the shaft of the measurand goes through one rotation. For instance, one rotation of the measured shaft can cause the potentiometer shaft to rotate 10 times. In fact multiturn potentiometers are available to up to about 60 rotations. In a similar manner, motion-amplifying mechanisms can be used for translational devices.

7.4.5 Linear Potentiometers

We have already seen that there can be error in linearity due to loading effects, but there are other factors that give rise to non-linearity such as non-uniformity in the wire area, winding, and mandrel diameter. Unless specified, the potentiometers supplied are non-linear in nature.

Errors in linearity can be corrected by adding fixed resistances in series and/or parallel at proper locations on the winding.

The best linear potentiometers available in the market have a non-linearity of 1% of full scale for $\frac{1}{2}$ in dia. multiturn. In any case the accuracy can be no better than $\frac{1}{2}$ resolution.

7.4.6 Non-Linear Potentiometers

All potentiometers are non-linear in nature. However, to generate a specific non-linear function, a pot is to be designed. If a function $V = kd^2$ is to be generated then a potentiometer as shown in Fig. 7.5 is suitable.

$$h_d = \frac{dL}{D}, \quad \because \text{ Area} = \frac{1}{2} hd \times d = \frac{1}{2}\frac{L}{D} d^2 = kd^2 \tag{7.20}$$

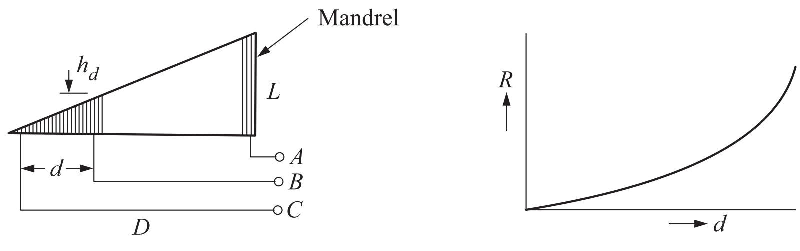


Figure 7.5 Potentiometer to generate $V = kd^2$ relation

To generate sine waves, the potentiometer given in Fig. 7.6 is suitable. In general, any non-linear function can be approximated to a number of linear segments, and each linear segment can be generated between taps of the potentiometer as shown in Fig. 7.7.

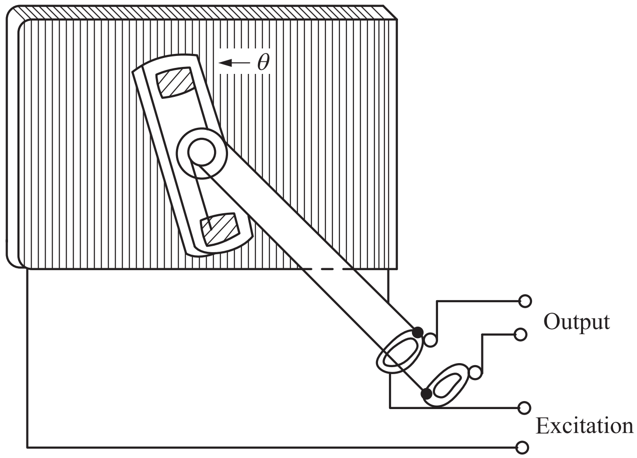


Figure 7.6 Constructional details

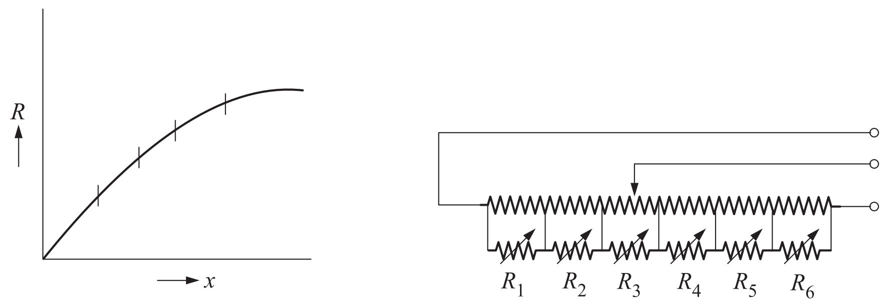


Figure 7.7 Resistance variation with distance

7.4.7 Noise

In potentiometers, noise is the spurious output of voltage fluctuation during the movement of the slider. This spurious output voltage is produced by resolution, and various mechanical and electrical defects. One of the main sources of noise is the vibration of the slider. The motion of the slider over the resistance wires of the wire-wound potentiometers may cause vibration of the contact at a certain speed, thus making only intermittent contact. The magnitude of vibration is significant if the frequency of vibration is near the resonant frequency of the spring-loaded contact. To overcome this effect, two wipers are used with different resonant frequencies as shown in Fig. 7.8.

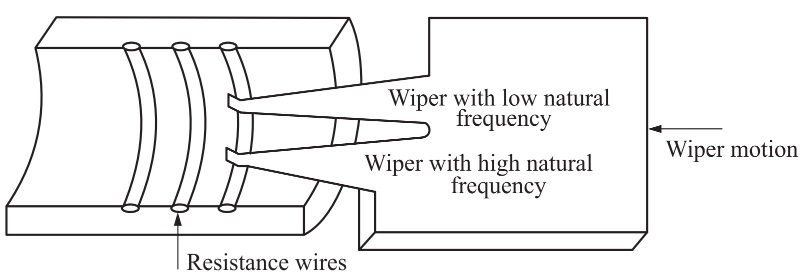


Figure 7.8 Antivibration wiper contact

Here, if one wiper resonates at a certain speed, the other does not, and hence they make good contact with the pot wire. Another manner of reducing this intermittent contact is by filling the potentiometer with a damping fluid. This damping also increases the tolerance of the unit to shock and vibration.

Another source of noise is the dirt and wear-out products of the potentiometer, which come between the wiper and the wires. This makes the contact resistance different at different positions. In addition, the contact resistance varies during the movement of wiper due to unevenness of the resistive wire. This random variation of contact resistance causes a random iR drop when a load current of i flows through. Noise voltage can also appear due to a current i_c flowing from one turn of the resistive wire to the next through the contact (Fig. 7.9).

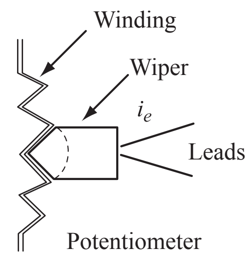


Figure 7.9 Noise voltage generation in potentiometers

The dynamic response of a potentiometer is essentially that of a zero-order instrument if the displacement is considered as input and the voltage is considered as output. The order of an instrument or system refers to the order of the differential equation describing the mathematical model of the instrument or system. Therefore, any change in the input will immediately change the output correspondingly.

The force or torque required to start the movement of the brush and maintain a motion depends on the mechanical loading of the instrument. The mass or moment of inertia and the friction between the wiper and the wires are to be considered to find out the mechanical input impedance of the system. These values vary over a wide range depending on the construction of the potentiometer. Special low-friction rotary pots are available with a starting torque of as small as 200 dyne-cm (0.003 oz. in). However, commercial pots may have 6600 to 33,000 dyne-cm (approximately 1 oz. in – 0.5 oz. in). Translational pots may have friction values 10 – 100 times more, and inertia values vary widely with size.

A typical 22 mm diameter single-turn pot has a moment of inertia of 0.12 g-cm^2, while a 50 mm diameter 10-turn pot has about 18 r-cm^2. The input variable to the pot, which is the measured variable, is displacement, and the loading quantity is the generalised input admittance or compliance. If friction is neglected as it can be analysed if it contains only inertia (Fig. 7.10), then

$$f = Ma; \quad \therefore\ f = \frac{M\,d^2X}{dt^2} = MD^2X \tag{7.21}$$

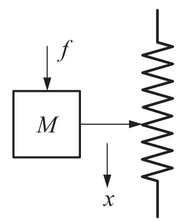


Figure 7.10 Inertia in potentiometers

where d is the differential operator. Then the generalized input admittance or compliance

$$C_i(D) \triangleq \frac{\text{displacement}}{\text{force}} = \frac{x}{f}(D) = \frac{1}{MD^2} \tag{7.22}$$

If we draw the block diagram of the system with its transfer function, we get Fig. 7.11.

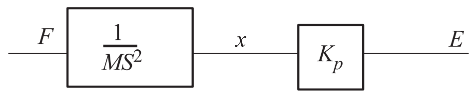


Figure 7.11 Block diagram of the system with transfer function

The frequency response of this system is given by $C_i(j\omega) = \dfrac{1}{M(j\omega)^2}$ (7.23)

$$C_i(j\omega) = \frac{1}{M(j\omega)^2} = \frac{-1}{MW^2} = \frac{1}{MW^2}\angle 180^\circ \tag{7.24}$$

Therefore, for lower frequencies ($j\omega$) is very high. Hence, loading is negligible if the output impedance of the previous stage is not very small. If mass is reduced, the low-frequency range can also be reduced further.

The performance of the instrument may decrease if environmental factors such as temperature, shock, humidity, and altitude are extreme. Special precautions must be taken in selecting the potentiometer, especially, the materials of the potentiometer. Under a specified environment, the life of the potentiometer may be more than 20 million full strokes or rotations.

The main advantage of potentiometer transducers is their accuracy and simplicity. Potentiometers are hardly influenced by acceleration and vibration. However, they require precision machining and their resolution is limited.

7.5 RESISTANCE STRAIN GAUGES

The next important transducer making use of resistance variation as a fundamental property is the *resistance strain gauge*. If a conductor is subjected to a stress, this resistance will change because of dimensional changes and also because of a change in piezoresistance. If we differentiate the equation for resistance R, we will know how a change dR in R depends on the basic parameters.

$$dR = \frac{a(\alpha\, dL + L\, d\alpha) - \alpha L\, da}{a^2} \tag{7.25}$$

Dividing the L.H.S. by R and the R.H.S by $\dfrac{\alpha L}{a} = R$, we get

$$da = -a\left(1-\gamma\frac{dL}{L}\right)^2 - \text{a} \quad \text{where } \gamma \text{ is Poisson's ratio} \tag{7.26}$$

$$\frac{da}{a} = -2\gamma\frac{dL}{L} + \gamma^2\left(\frac{dL}{L}\right)^2 \tag{7.27}$$

Neglecting the second term as it is very small,

$$\frac{dR}{R} = \frac{dL}{L} + \frac{d\alpha}{\alpha} + 2\gamma \frac{dL}{L} \tag{7.28}$$

$$\frac{(dR/R)}{(dL/L)} = 1 + 2\gamma + \frac{d\alpha/\alpha}{dL/L} \tag{7.29}$$

The LHS of the above equation is called a gauge factor. The gauge factor is the unit resistance change per unit strain, which is due to three factors as revealed by the above equation. The first term is the resistance change due to length change, the second term is the resistance change due to area change. The third term is the resistance change due to piezoresistance effect.

If the gauge factor for a strain gauge is known, then a measurement of resistance change allows one to determine the strain to which the gauge is subjected to. The gauge factor for a strain gauge is normally supplied by the manufacturer. This factor can be computed and experimentally verified:

$$\boxed{\text{Gauge factor} = 1 + 2\gamma + \frac{d\alpha/\alpha}{dL/L}} \quad \text{or} \quad (1 + 2\gamma) \text{ approx.} \tag{7.30}$$

Poisson's ratio is always between 0 and 0.5 for all materials. Gauge factor can also be expressed as

$$\boxed{\text{G.F.} = \pi E} \tag{7.31}$$

where π_1 is the longitudinal piezoresistance coefficient and E the modulus of elasticity.

Some of the most common types of strain gauges are made of a material, *advance* or *isoelastic*. Advance is an alloy of 55% copper and 45% nickel. The gauge factor is about 2. Isoelastic is an alloy of 36% nickel, 8% chromium, 4% manganese, silicon, and molybdenum, and 52% iron. This has a gauge factor of about 3.5. Semiconductor materials can be used as strain gauge elements. These semiconductor strain gauges have a very high gauge factor of about 125. This high gauge factor is due to the large change in resistance of the strain gauge material due to piezoresistive effect. This effect is not linear with strain as the longitudinal piezoresistance coefficient is a function of stress. However, in metallic gauges the resistance change under strain is mostly due to dimensional change and the gauge factor is fairly constant.

In metallic gauges, there are essentially two varieties: *unbonded type and bonded type*.

In the unbonded type, the resistance wires of about 0.025 mm dia. (0.001 in) are fixed with some initial tension between two frames that can move relative to each other. This initial tension or preload is necessary to avoid buckling under compression or negative displacement and this preloading should be greater than any expected compression or negative displacement. A simplified figure is shown in Fig. 7.12.

Fig. 7.13 shows another type of unbonded strain gauge for angular motion measurement.

Clockwise angular motion given to the inner member, which is pivoted to the outer stationary member, increases the tension on the wires A and C and reduces the preload on the wires B and D. If they are connected in a bridge, as shown, then the output voltage available is four times the voltage that would have been obtained due to a single wire. This arrangement is useful for measurement of torsional strains and angular displacement, angular velocity, etc. This type of gauge can be used to measure only very small displacements of the order of 0.004 cm full scale. Normally these gauges are used as sensors for force, pressure, and acceleration. In these cases the strain wires serve as the necessary spring element to transduce force to displacement. This displacement is sensed as a resistance variation. The ranges of force and deflection values are decided by the size, length of wires, and the number of wires used.

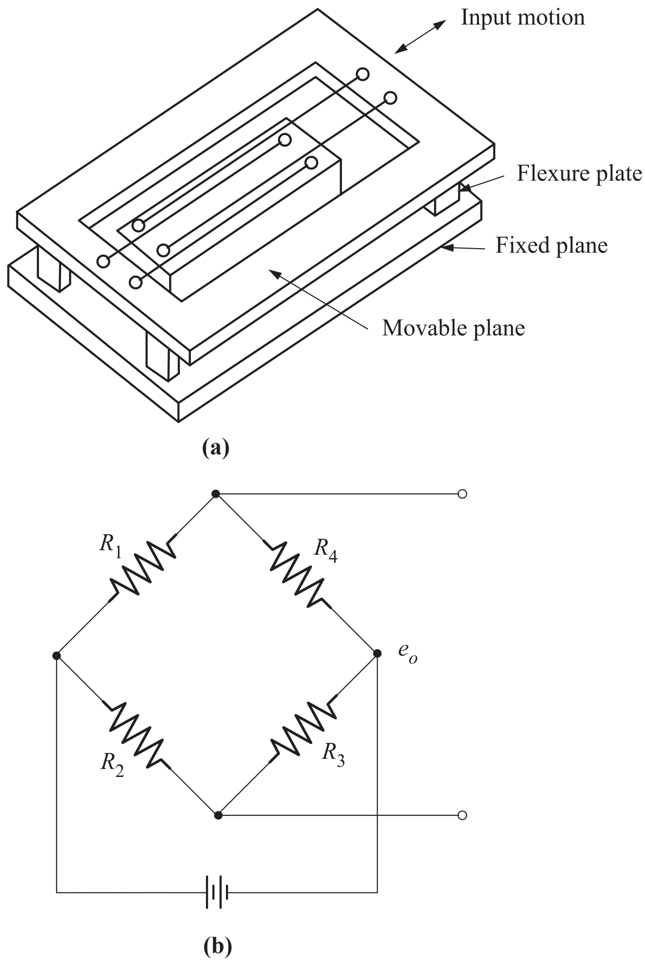


Figure 7.12 (a) Unbond-type strain gauge and (b) bridge circuit

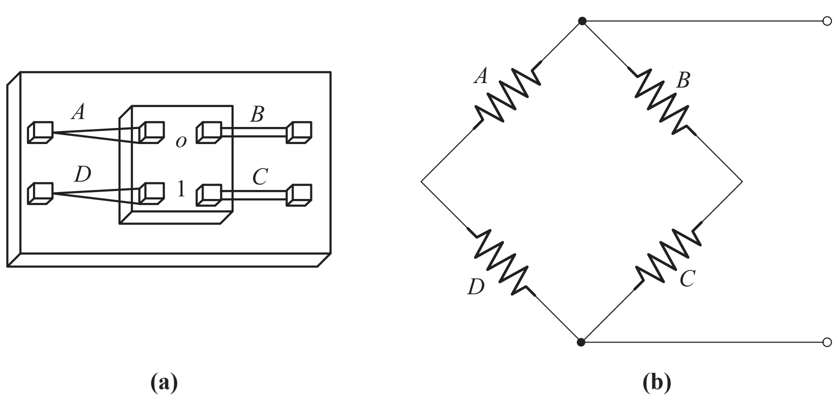


Figure 7.13 (a) Another type of unbonded gauge and (b) bridge circuit

The sensitivity for a bridge excitation of 5 V is 40 mV full-scale output for 0.006 cm full-scale displacement. The nominal value of resistance of the bridge arms is 350 Ω. The thermal sensitivity shift is 0.02% between –18 and 120°C.

The order of the instrument is zero if displacement is considered as the input and the bridge voltage as output. If the input is force then the order is two.

Here the moving mass is M and the spring constant K_s is determined by the resistance wires and the flexure.

$$f \longrightarrow \boxed{\dfrac{1}{M_s^2+K_s}} \xrightarrow{\;d\;} \boxed{K} \longrightarrow e_o$$

Figure 7.14 Transfer function block schematic

7.6 BONDED-TYPE STRAIN GAUGES

In the case of bonded-filament strain gauge, the resistance wire is made into a form of a grid and cemented between two pieces of thin paper. The factors to be considered are as follows:

1. Filament construction.
2. Material of the filament wire.
3. Base carrier material.
4. Cement used to bond the filament to the carrier.
5. Lead wire connections.

7.6.1 Filament Construction

The filament wire is of a bonded type, which is made in the form of a flat grid or flattened helix or a thin foil to give a flat grid pattern as shown in Fig. 7.15.

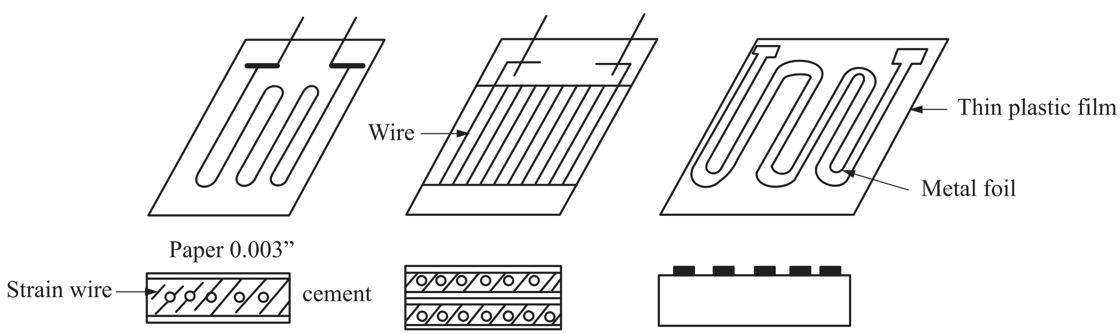


Figure 7.15 Construction types of strain gauges

In the flat grid type, the gauge length is around 1 in and the wire size is 0.001 μ. For a shorter gauge length, up to 1/16 in helical grid type is suitable. In this type, the resistance wire is wound around a thin-walled cylinder in the form of a helix. This cylinder is then flattened and bonded between two sheets of insulating material. For still shorter gauges up to 1/64 in, foil type is suitable. In this type, a metal foil of about 0.00015 in thickness is formed on plastic film of 0.001 in, giving a total thickness of 0.001 in. This foil is etched by suitable processes so that a metallic grid is formed.

These strain gauges are mainly sensitive to the component of strain along their longitudinal axis. However, they are also sensitive to strain to a small extent in the transverse direction. This is because of the loops at the end of each turn of wire. In construction, this end loop should be made as small as possible. For this reason the end loops in the foil type are made thicker to reduce the resistance of this portion of the gauge.

The nominal value of resistance for these gauges range from 40 to 2000 Ω, but the common values are 120, 350, and 1000 Ω.

7.6.2 Materials of the Filament Wire

The material chosen should have a high gauge factor. Common materials used are advance, isoelastic, and nichrome V (nickel–chrome alloy with a GF of 2.2). The choice of material depends not only on the gauge factor but also on its temperature sensitivity. The metal alloys used for strain gauges are temperature sensitive. Therefore, the environmental variations also have to be considered when choosing a particular material. For example, if strains are to be measured over a long period and the temperature is likely to change widely between each test, then low temperature-sensitive material like copper – nickel alloy is to be used despite the fact that it has a low gauge factor. For measurement of strain above 250°C nichrome V is suitable. For strain measurements at still higher temperatures, platinum alloys are used.

As a matter of fact, many more points are to be considered when the grid material is selected. Selection is actually a compromise between the following factors.

1. High gauge factor.
2. High resistivity.
3. Low-temperature sensitivity.
4. High electrical stability.
5. High yield point.
6. High endurance limit.
7. Good workability.
8. Good solderability.
9. Low hysteresis.
10. Low thermal *emf*.
11. Good corrosion resistance.

7.6.3 Base Carrier Material

The carrier material or the support material should have the following properties:

1. Good adherence to cement used.
2. High dielectric strength.
3. High mechanical strength.
4. Minimum thickness consistent with other factors.
5. Minimum temperature restriction.

For room temperature applications, the carrier material used for the filament wire is a nitrocellulose-impregnated paper and for higher temperatures, phenolic-plastic-impregnated paper is used. Nichrome foil gauges are available in the phenolic-glass carrier construction with integral welded leads.

7.6.4 Strain Gauge Cements

The cement to be used for joining the strain gauge to the straining member should have the following properties:

1. High mechanical strength.
2. High creep resistance.
3. High dielectric resistance.
4. Good adherence.
5. Minimum moisture attraction.
6. Minimum temperature restriction.
7. Ease of application.
8. The capacity to dry fast.

Lead wire connections: The vulnerable point for failure in the wire strain gauge is at the discontinuity formed at the junction between grid and lead. A soldered junction between the fine filament and heavier lead wire or ribbon may be alright for static or slowly varying loads. However, for dynamic strain measurements, the joint is welded. Lead wire materials should have a low stable resistivity and a minimum temperature coefficient of resistance. These lead wires are to be insulated properly with materials of the same thermal classification as the gauge carrier and bonding elements. The recommended lead wire insulation material for the temperature range is given below:

Below 75°C – Nylon.

Between 75°C and 65°C – Vinyl.

Between 75°C and 95°C – Polyethylene.

Between 75°C and 260°C – Teflon.

Above 260°C – Glass sleeving or glass impregnated silicon.

7.6.5 Temperature Effect on Strain Gauges

Temperature effect is one of the troublesome factors in the use of resistance strain gauges. The strain gauge is affected by temperature variation in two manners:

1. The strain produced by differential thermal expansion existing between the grid support and the gauge.
2. The variation in resistance due to temperature variation, unless proper correction or compensation is done, will be very different from the actual strain. This temperature effect can be cancelled in various ways. One way is to use a dummy gauge identical to the active gauge. This dummy gauge is cemented to a piece of the same material, which is unstrained and put in the vicinity of the active gauge so that the same temperature variation is experienced by the dummy gauge. This dummy and the active gauge are placed in the adjacent arms of a Wheatstone bridge, so that resistance changes due to differential thermal expansion and temperature co-efficient of resistance will have no effect on the bridge output voltage. Another way is to construct a special inherently temperature-compensated gauge.

Temperature may change the gauge factor also. This effect is negligible in metallic gauges but seriously affects semiconductor gauges. This temperature problem has been successfully solved in the past for the temperature range 7–1100°C by using liquid Relium.

7.6.6 Measurement of Strain

The strain measured with the help of the strain gauge will be a sort of average strain as strain gauge spreads over an area. However, strain is normally defined with respect to a point. Our measurement will be correct only if the strain gradient is constant and the strain is uni-axial. The strain gauge is to be fixed so that its axis is aligned with the strain axis, and its midpoint (which is marked by the manufacturer on the gauge) coincides with the point of interest. If the strain gradient is not constant over the area of gauge and its form is not known, then the strain read by the gauge cannot be associated with a point. In such a situation, the smallest practical gauge is very useful. There are situations where the direction and magnitude are both unknown. In such a case, the strain measurement in three different directions from a point is necessary to determine the maximum strain and its direction. If three different gauges are to be fixed at a point, then arrangement to be done one over the other, which may not be very satisfactory for obvious reasons. Fixing them side by side will cover more area and hence the uncertainty of the measurement will increase. To overcome all these difficulties to a certain extent, strain gauge *rosettes* have been developed. They are available both in wire type and also in foil type. Fig. 7.16 shows a two-element rosette.

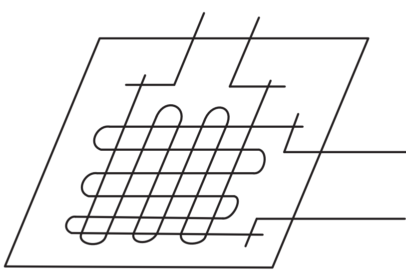
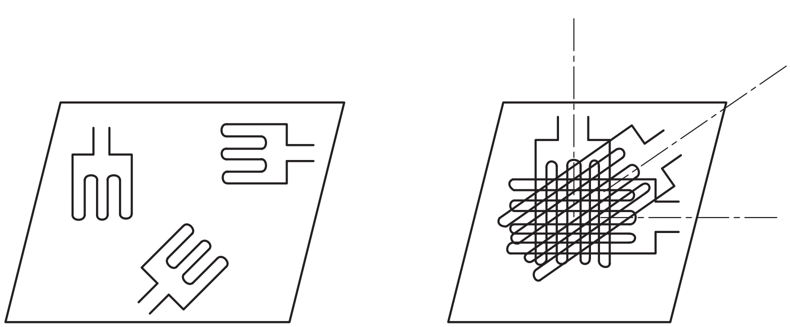


Figure 7.16 Two-element strain gauge rosette

Often two strain measurements at right angles will suffice to determine a stress—both direction and magnitude—in a plane. This rosette can also be used for measurement of strain in a known direction with temperature compensation. The grid at the strain axis will compensate the resistance variation due to temperature change, and resistance variation due to strain is negligible. Fig. 7.17(a) is a delta rosette and Fig. 7.17(b) is a rectangular rosette. These rosettes can be used for any strain measurement when direction and magnitude are completely unknown.



(a) (b)

Figure 7.17 Strain gauge rosette: (a) delta rosette and (b) rectangular rosette

7.6.7 Strain Gauge Circuitry

The strain gauge can transduce a strain into a resistance change only. If a voltage change corresponding to the strain is needed, then a potentiometer circuit or a bridge circuit with an excitation should be thought of. One may think that an ohmmeter can be used to find the change in resistance, which can be properly calibrated to give the corresponding strain. This is not possible because the sensitivity of the ohmmeter is far less than the sensitivity required for strain measurement. As a matter of fact strains of 1 μ cm/cm are detectable with commercial equipment. Typical gauge constants are: gauge factor $F = 2.0$ and $R_g = 120\ \Omega$. For these values the change in resistance that is to be measured is

$$\begin{aligned} dR &= F \sigma R_g, \quad \text{where } \sigma \text{ is the strain} \\ &= 2 \times 1 \times 10^{-6} \times 120 \\ &= 0.00024\ \Omega \end{aligned}$$

No ohmmeter will be able to measure this small resistance change fairly accurately. Therefore, bridge and potentiometric methods are to be used.

7.6.8 The Ballast Circuit

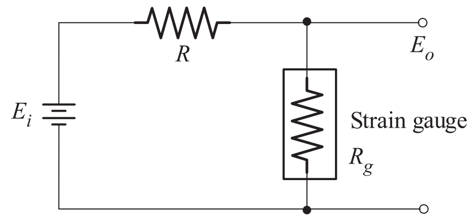


Figure 7.18 Ballast circuit

$$E_o = R_g E_i/(R + R_g) \tag{7.32}$$

$$\frac{dE_o}{dR_g} = \frac{\left[(R_g + R)dR_g - (R_g dR_g)\right]}{(R_g + R)^2} \tag{7.33}$$

$$E_i = \frac{dR_g E_i}{(R_g + R)^2} \quad \text{or} \quad dE_o = \frac{R\, dR_g E_o}{(R_g + R)^2} \tag{7.34}$$

$$= \frac{RR_g E_i}{(R + R_g)^2} \frac{dR_g}{R_g}; \quad \text{but } GF = F = \frac{dR_g}{R_g} I \tag{7.35}$$

$$dE_o = \frac{RR_g E_i}{(R + R_g)^2} F \tag{7.36}$$

The value of R relative to R_g is large so that it acts as a ballast resistance to keep the current I constant regardless of gauge resistance change. The normal value of R is two to four times the magnitude of R_g.

7.6.9 The Wheatstone Bridge Circuit

The simplest form of a Wheatstone bridge for the measurement of strain is shown in Fig. 7.19.

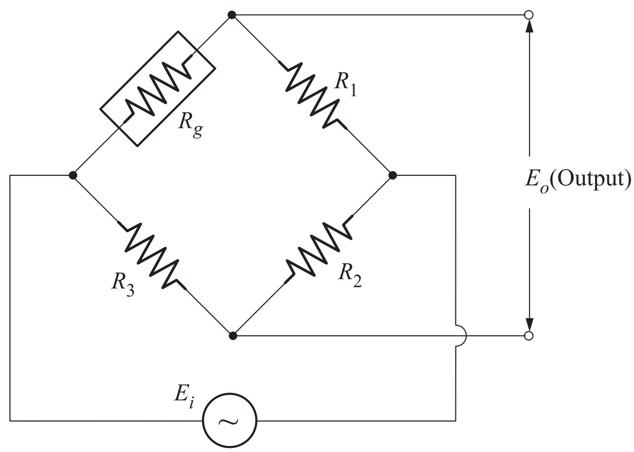


Figure 7.19 Wheatstone bridge circuit for strain gauges

The bridge is balanced initially when there is no strain. When the gauge is strained, the resistance changes and thereby causes an unbalance in the bridge. The voltage output is given by the following expression.

Considering the upper half of the bridge; the potential of the centre point is given by

$$E_c = \frac{R_g}{R_g + R_1} E_i \tag{7.37}$$

The change in this voltage due to a change in R_g will be the output voltage as the potential of the reference point is also initially E_c.

$$E_c = \frac{dE_c}{dR_g} \tag{7.38}$$

As shown in the previous ballast circuit

$$\boxed{E_o = \frac{RR_g}{\left(R_g + R_1\right)^2} E_i F} \tag{7.39}$$

In this basic circuit there is no temperature compensation. To have temperature compensation, another dummy gauge fixed to an unstrained piece of the same material is used as shown in Fig. 7.20.

The active gauge and dummy gauge are the same in all respects and therefore have the same temperature coefficient of resistance and expansion. Any change in R_{g_1} due to temperature is exactly compensated by the same change in R_{g_2}.

The dummy gauge can also be put in the other arm as shown in Fig. 7.21.

In this bridge temperature effect will change R_{g_1} and R_{g_2} to the same extent, and thereby the potential of the centre point is unaltered. The advantage of this method is that the value of R_1 in Fig. 7.22 need not be known accurately. The value of R_{g_2} is normally known with certain accuracy.

If higher voltage output is required, another set of active and dummy gauges can be used as shown in Fig. 7.22.

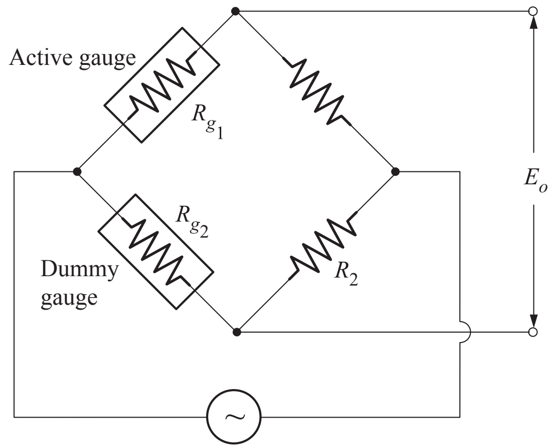


Figure 7.20 Bridge circuit for strain gauges

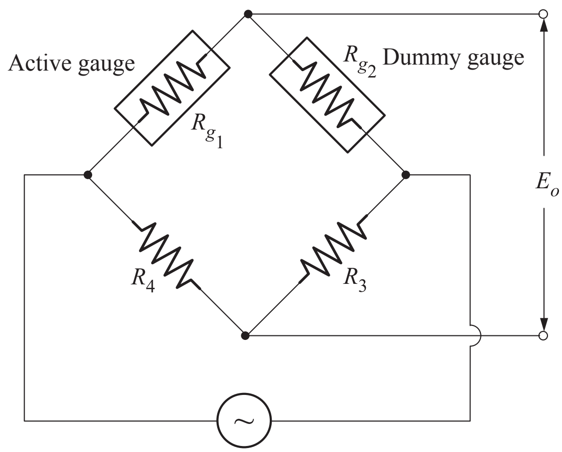


Figure 7.21 Bridge circuit for strain gauges

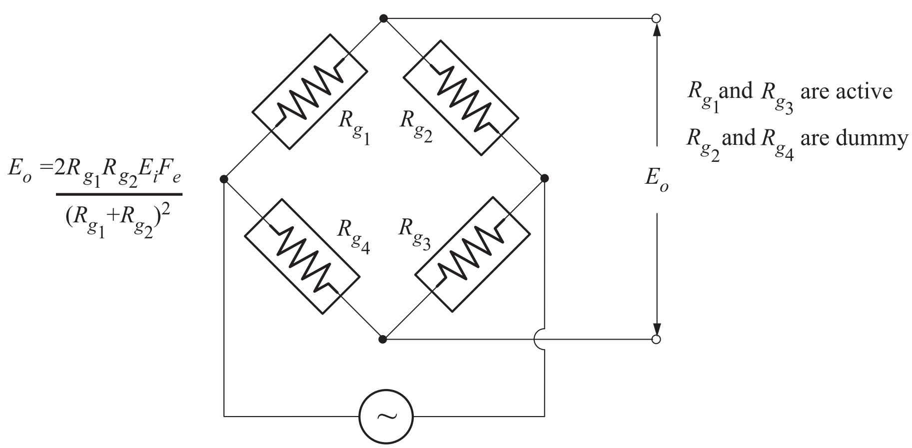


Figure 7.22 Using two sets of active and dummy gauges in bridge circuit

When a bending stress is to be measured the circuit in Fig. 7.22 can be used by just making the dummy gauges R_{g_2} and R_{g_4} active and subjecting them to compression while R_{g_1} and R_{g_3} are subjected to tension as shown in Fig. 7.23. Here the output is $E_o = 4R_{g_1} R_{g_3} E_i F\varepsilon / (R_{g_1} + R_{g_3})^2$.

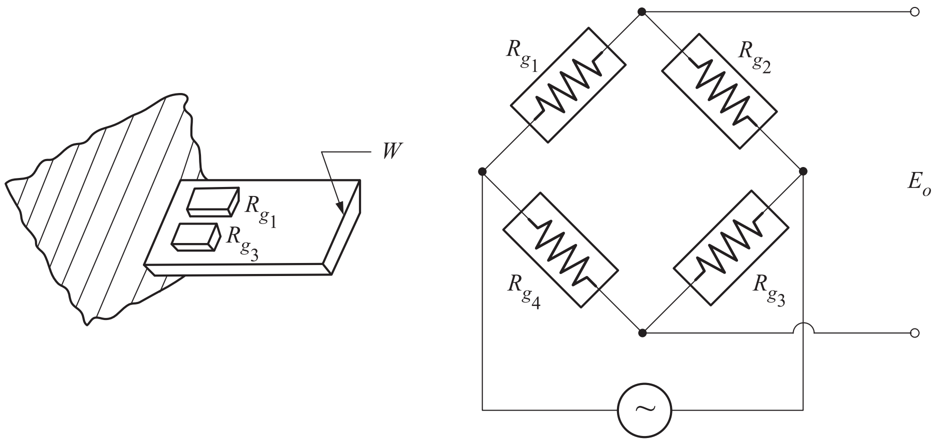


Figure 7.23 Strain gauges subjected to tension and bridge circuit

All the output voltages considered are open circuit voltages. When a load is connected across the terminals, the output voltage is calculated using Thevenin's network theorem as shown in Fig. 7.24.

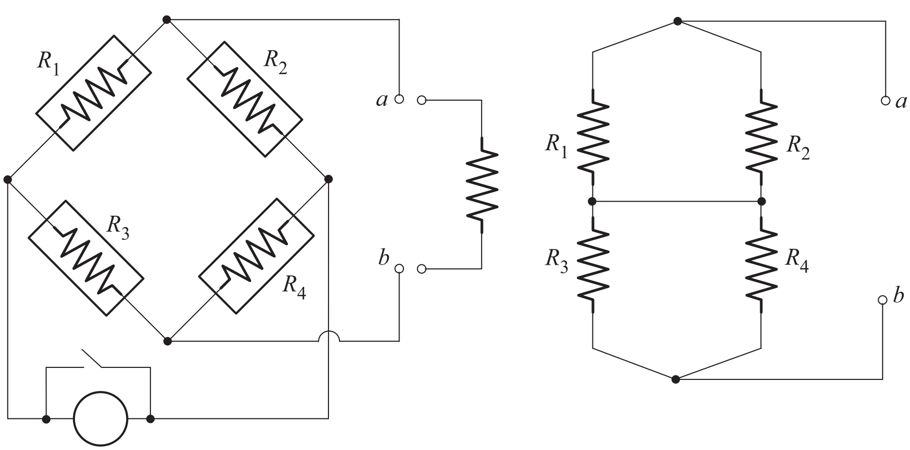


Figure 7.24 Bridge circuit and calculation using Thevenin's network

7.6.10 Circuit for Calibration

Normally calibration of any measuring system is done by applying an accurately known sample of the variable to be measured and observing the measuring system's response. In the case of a bonded resistance strain gauge, this is not possible because once the gauge is bonded to a known strain situation for calibration, it cannot be transferred to the test item. Therefore, some other method is to be thought of.

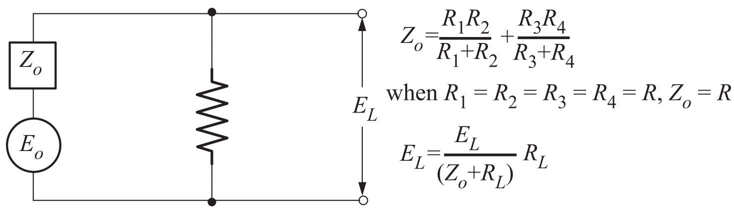


Figure 7.25 Determining system response

Resistance strain gauges are manufactured under carefully controlled conditions, and the gauge factor for each group of gauges is provided by the manufacturer, within an indicated tolerance of about ±0.2% . The nominal resistance value is also known. Knowing these two, a simple method to calibrate is to introduce a known small resistance change at the gauge and determine the system's response (Fig. 7.26). Then the equivalent strain that will introduce the same resistance change can be calculated.

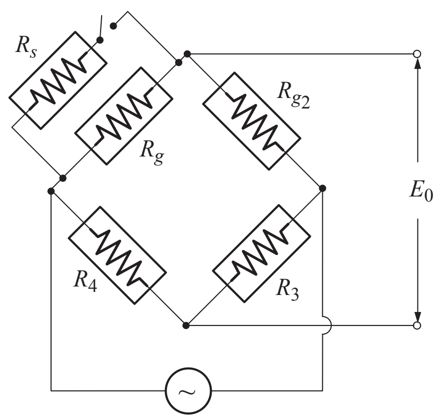


Figure 7.26 Introducing resistance change and determining system response

For a shunt resistance R_s, the change in resistance in arm 1,

$$dR_g = Rg - \frac{R_g R_s}{R_g + R_s} = \frac{R_g^2}{R_g + R_s} \tag{7.40}$$

In terms of change in resistance, the equivalent strain is

$$\boxed{\text{Strain}, \in = \frac{1}{F}\frac{dR_g}{R_g} = \frac{1}{F}\frac{R_g}{R_g + R_s}} \tag{7.41}$$

For dynamic calibration, the electrically driven switch or chopper is made use of to make and break the shunt resistance circuit, and the response is observed on the CRO.

7.6.11 Mounting of Strain Gauges

Mounting of strain gauges is an important problem because an improperly mounted gauge can cause a large error however careful one is in the selection of gauge and circuitry. For the purpose of mounting, the strain gauges may be divided into three groups as follows:

1. Ordinary paper-backed gauge.
2. Phenolic or bakelite-backed gauge.
3. Epoxy-film-supported gauge.

In general, the mounting area should be cleaned of all corrosion, paint, and dirt, and the test material must be exposed. Emery cloth is a good material to use for this purpose. After the bare material is prepared, all traces of greasy film must be removed using a suitable solvent. Acetone is a satisfactory material. The surface should be swabbed with surgical cotton repeatedly until no trace of oil or soil is found on the cotton. The gauge also must be cleaned from any oil. Cleaning of the gauge requires special care as this may damage the gauge itself.

The gauge has to be mounted slowly and carefully. There is no necessity for hurrying up as it takes several hours for the bond to become usable.

7.6.12 Commercial Strain-Measuring Systems

For static strain measurements, a strain indicator is sufficient. In this type, the resistance bridge is generally provided with fine and coarse adjustments for balance. It is excited with an oscillator. Any unbalance is amplified and fed to a phase-sensitive demodulating circuit, which supplies the balancing meter. The bridge is balanced for null deflection and the balancing adjustment is calibrated in terms of strain in μ in/in or μ cm/cm.

For dynamic strain measurements, the indicator type is not suitable. A strain gauge bridge with a CRO is suitable. Basically, this instrument consists of a DC-excited resistance bridge, a chopper, a three-stage amplifier, and a CRO. If the strain is a fast-varying one then the chopper is not necessary. Multichannel strain-measuring systems with an oscillographic recorder is also available.

7.6.13 Stress Measurement on Rotating Members

The real problem in using strain gauges for the measuring stresses of rotating members is to make connection with the rotating members. The common method used is to use slip rings and brushes. There are a lot of problems in this method. For example, the wear and tear of brushes, the variation of contact pressure over usage, and the necessity for readjusting the brush pressure every now and then. All these troubles are not worth taking if there is a temporary method. Normally the setup with strain gauges will be used for 10–20 min and discarded, as the necessary information can be obtained within that time. Of course, there are situations where the experimental setup will be retained for months. However in a majority of cases, the setup is used for a short time and dismantled. In such a case, it is advisable to think of simpler methods. If the rotating member rotates slowly, the leads may be allowed to wrap around and stop the rotation at the end of the lead wire. It is first wrapped into the shaft. This method is not suitable if the rotating member cannot be stopped quickly enough as the end of the cable approaches. Another method is to provide a fast disconnecting arrangement. This needs no more than soldered connections that can be quickly peeled off. Even though this method is somewhat limited, it is quite workable at slow speeds and avoids many of the problems inherent in other methods.

Yet another method is to use a radio frequency transmitter mounted on the shaft, which transmits the strain gauge information. A receiver placed nearby picks up the signal. This method is used successfully by W. R. Campbell and R. F. Suit Jr. The equipment used is bulky and expensive.

The last method would be to use a painted ring. Here an insulating liquid, say Glyptal, is painted on the shaft. Over this, silver paint is brushed on, dried, and polished to form the slip rings. However, this method is suitable only for short-time gauge installations, which are dismantled after the test.

7.6.14 Special Problems in Strain Gauge Applications

1. **Cross-sensitivity:** The majority of the strain-sensitive grids of the bonded-type strain gauge are normally arranged along the sensitive axis of the gauge. However, a part of it is aligned transversely. This transverse portion of the grid senses the strain in that direction and this effect is superimposed upon the longitudinal output. This is known as cross-sensitivity. The error is small, not exceeding 2% or 3% in worst cases. For engineering applications, this error can be tolerated. For scientific and research activities, this error has to be reduced further. Some of the methods used for this purpose are

   a. The transverse portion of the grid is made much heavier in the section than the longitudinal portions. This is easily done in case of etched foil-type gauges.

   b. The grid is made of two wire sizes. The transverse elements are of a relatively large-diameter wire, while the longitudinal strain-sensitive elements are of a small section.

   In both cases, the heavier elements are not appreciably strained because the cement is incapable of supplying the required force and also because the resistance of this heavier part is a small fraction of the total resistance of the gauge.

2. **Fatigue:** Repeated straining of the gauge causes fatigue failure. This failure is mostly due to the fracture at the discontinuity formed at the junction of the grid wires and the lead wires. The endurance can be improved by using a *dual lead* gauge. The construction feature of this type of gauge is shown in Fig. 7.27.

A wire of intermediate size between the grid wire size and the lead wire size is interposed between the grid and the lead. This construction markedly increases the fatigue endurance of the gauge.

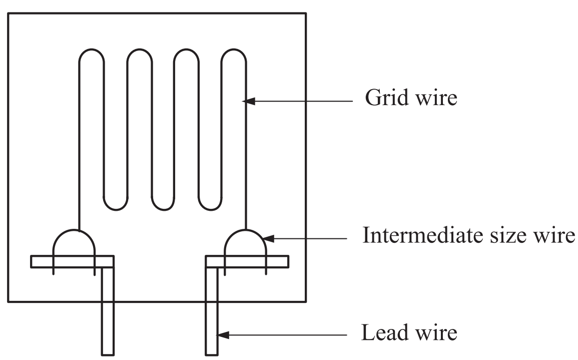


Figure 7.27 Dual lead gauge

7.6.15 Semiconductor Strain Gauges

The gauge factor for semiconductor strain gauges is mainly contributed by the piezoresistance effect. For simple tension or compression, when the current through the gauge is along the stress axis, the per unit change in resistivity ρ is given by

$$\frac{d\rho}{\rho} = \alpha_L \sigma \tag{7.42}$$

where α_L is the longitudinal piezoresistance coefficient and σ the stress. Then the gauge factor is

$$\text{G.F.} = 1 + 2\gamma + \alpha_L \frac{\sigma}{\in} \tag{7.43}$$

$$\boxed{\text{Gauge factor, G.F.} = 1 + 2\gamma + \alpha_L E} \tag{7.44}$$

where γ is Poisson's ratio and E Young's modulus of elasticity. The gauge factor for semiconductive materials can be as high as 200 and hence the usual strain gauge amplifier is not needed.

Several ceramic type of semiconductors have been evaluated, but mostly silicon is employed as the strain-sensitive material. The elastic nature of a single silicon crystal below 1000 °F is nearly perfect and is essentially free of hysteresis, drift, or creep. The magnitude of the piezoresistance effect depends upon the direction of the applied stress and current flow with respect to the crystallographic axes. The strain-sensitive element in these gauges consists of a filament sliced from a single crystal. Typical dimensions of a semiconductor strain gauge are given in Fig. 7.28.

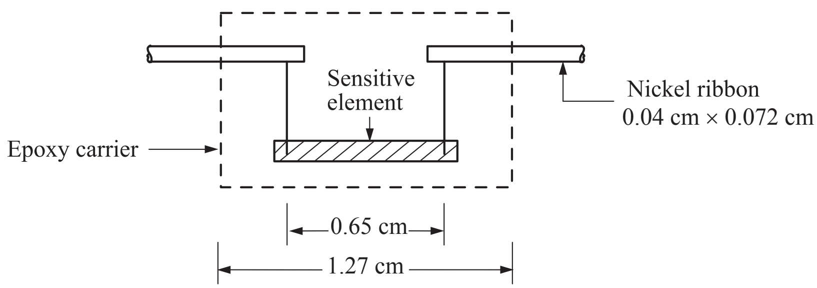


Figure 7.28 Typical dimensions of a semiconductor strain gauge

Semiconductor gauges are highly non-linear. The gauge factor decreases rapidly with increasing temperature due to increase in resistance. However, at a given temperature the gauge sensitivity is constant, better than 0.1%. Temperature compensation can be effected by proper choice of resistive elements or resistance thermometers.

The drastic drop in gauge sensitivity with increasing temperature can be reduced by using a constant current bridge circuit as shown in Fig. 7.29. However, the voltage output is non-linear.

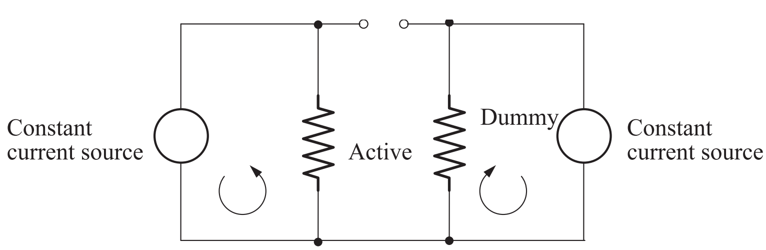


Figure 7.29 Constant current bridge circuit

Example 7.2

A linear resistance potentiometer is 50 mm long and is uniformly wound with a wire of total resistance 5000 Ω. Under normal conditions the slider is at the centre of the potentiometer. Determine the linear displacement when the resistance of the potentiometer is 1850 Ω.

Solution

$$\text{Total resistance} = 5000\ \Omega$$

$$L = 50\ \text{mm}$$

$$\frac{Rt}{L} = \frac{5000}{50} = 100\ \Omega$$

$$\text{Resistance of normal position} = \frac{5000}{50} = 2500\ \Omega$$

$$\text{Change in resistance} = 2500 - 1850 = 650\ \Omega$$

$$\text{Displacement, } y = \frac{650}{100} = 6.5\ \text{mm}$$

Example 7.3

A potentiometer is provided with 50 turns per mm. The gearing arrangement is such that the motion of the main shaft by one resolution crosses 4 resolutions. Determine the potentiometer's resolution.

Solution

$$\text{Resolution of potentiometer} = 1/\text{no. of turns per mm}$$

$$= \frac{1}{50} = 0.02\ \text{mm}$$

4 resolutions of potentiometer with one rotation

$$= \frac{0.02}{4} = 0.005 = 5\ \text{mm}$$

Example 7.4

A thin circular/wire of soft iron has a gauge factor of 3.8. Determine Poisson's ratio.

Solution

$$\text{Gauge factor,} \quad G = 3.8$$

$$\text{Poisson's ratio,} \quad M = \frac{G - 1}{2} = \frac{3.8 - 1}{2} = 1.4.$$

Example 7.5

The wire of a strain gauge is 0.1 m long and has an initial resistance of 120 Ω. On application of force, the wire resistance increases by 0.21 Ω and length of 0.1 mm. Determine the gauge factor of the device.

Solution

$$l = 0.1 \text{ m} = 100 \text{ mm}$$

$$\Delta l = 0.1 \text{ mm}$$

$$\varepsilon = \frac{\Delta l}{l} = 0.001$$

$$R = 120\ \Omega$$

$$\Delta R = 0.21\ \Omega$$

$$G = \frac{\Delta R / R}{\varepsilon} = \frac{0.21/120}{0.001} = 1.75$$

Example 7.6

A resistance strain gauge is used to measure stress on steel. The steel is stressed to 1400 kgf/cm^2 Young's modulus = 2.1×10^6 kgf/cm^2. Calculate the percentage change in resistance of strain gauge assuming the gauge factor to be 2.

Solution

$$\text{Steel stress} = 1400 \text{ kgf/cm}^2$$

$$E = 2.1 \times 10^6 \text{ kgf/cm}^2$$

$$\text{Guage factor, } G = 2$$

$$\varepsilon = \frac{l}{E} = 0.00066$$

$$G = \left(\frac{\Delta R}{R}\right) / \varepsilon$$

$$\Delta R/R = 2 \times 0.00066 = 0.001313$$

$$\text{in \%,} = \frac{\Delta R}{R} \times 100 = 0.001313 \times 100 = 0.1313\%$$

Example 7.7

A strain gauge with a gauge factor of 2 is subjected to stress of 1000 kg/cm^2. $E = 2 \times 10^6$ kg/cm^2. Calculate the percentage change in resistance of the strain gauge. Find Poisson's ratio.

Solution

$$\text{Guage factor, } G = 2$$

$$\text{Stress, } e = 1000 \text{ kg/cm}^2$$

$$E = 2 \times 10^6 \text{ kg/cm}^2$$

$$\varepsilon = \frac{e}{E} = 0.0005$$

$$\frac{\Delta R}{R} = \varepsilon \times G = 0.0001$$

$$\frac{\Delta R}{R} \times 100 = 0.1$$

$$G = 1 + 2\mu$$

$$\mu = \frac{G - 1}{2} = \frac{2 - 1}{2} = 0.5$$

Example 7.8

A strain gauge having a resistance of 200 Ω and a gauge factor of 2.5 is connected in series with a load resistance of 400 Ω across 24 V. Determine the change in o/p voltage when a stress of 140 mgf/m^2 is applied. The modulus of elasticity is 200 GN/m^2

Solution

$$\text{Voltage across strain gauge} = 24 \times \frac{200}{200+400}$$

$$= 8 \text{ V}$$

$$\text{Strain } \varepsilon = \frac{\text{Stress}}{\text{Modulus of elasticity}} = \frac{140 \times 10^{-3}}{200}$$

$$\frac{\Delta R}{R} = \frac{\rho}{\varepsilon}$$

$$\frac{\Delta R}{R} = G.F \times \frac{\Delta l}{l}$$

$$\frac{\Delta l}{l} = \frac{\text{Stress}}{\text{Modulus of elasticity}} = \frac{140 \times 10^{-3}}{200}$$

$$\therefore \quad \frac{\Delta R}{R} = 0.007$$

$$\Delta R = R \times E \times G = 200 \times .0007 \times 2.5 = 0.35 \ \Omega$$

Voltage across strain gauge under

$$\text{Strained conditioned} = \frac{200 + 0.35}{200 + 0.35 + 400} \times 24$$

$$= 8.00933.$$

$$\text{Change in o/p voltage} = 8.00933 - 8$$

$$= 0.00933$$

$$= 9.33 \text{ mV}$$

Example 7.9

A platinum resistance thermometer has a resistance of 120 Ω at 25°C. Determine its resistance at 75°C. The temperature coefficient of resistance is 0.00392 at 25°C. If the resistance 180 Ω, what is temperature T_3?

Solution

$$\text{Resistance at 25 °C,} \quad R_1 = 120 \ \Omega \alpha_T = 0.00392$$

$$R_2 = R_1 [1 + \alpha_T [T_2 - T_1]]$$

$$= 120[1 + 0.00392(75 - 25)]$$

$$= 143.52 \ \Omega$$

for 180 Ω, temp T_3 = ?

$$R_3 = R_1[1 + \alpha_T(T_3 - T_1)]$$

$$= 120\ (1 + 0.00392(+3 - 25))$$

$$T_3 = \frac{180/120 - 1}{0.00392} + 25 = 152.55°C$$

Example 7.10

A copper resistor having a resistance of 15 Ω at 20°C is used to indicate the temperature of a machine. Determine the limiting value of resistance *k*, if the maximum temp is 175°C. The temperature coefficient (T.C.) = 0.00425 at 20°C.

Solution

$$T_1 = 20°C$$

$$T_2 = 175°C$$

$$R_2 = R_1[1 + \alpha_1(t_2 - t_1)]$$

$$\alpha_T = 0.00425$$

$$R_2 = 24.88\ \Omega$$

Example 7.11

A thermistor has temperature coefficient of resistance of –0.05 over a temperature range of 25–50°C. Determine the resistance of the thermistor at 40°C if the resistance of the thermistor at 25°C is 120 Ω.

Solution

$$\alpha = -0.05$$

$$R_{40} = R_{25}\ [(1 + \alpha(T_2 - T_1)]$$

$$= 30\ \Omega$$

Example 7.12

In a variable inductive transducer, the coil has an inductance of 2.5 mH when the effective turns on the coil arc are 50. Determine the inductance of the coil when the effective turns on the coil are 52.

Solution

$$\text{No. of turns on the coil, } N_1 = 50$$

$$L_1 = 2\text{–}5 \text{ mH}$$

$$N_2 = 52$$

$$L_2 = L_1 \times \frac{N_2}{N_1}^2$$

$$L \propto N^2$$

$$= 2.5 \times \frac{J_2}{J_0}^2 = 2.7 \text{ mH}$$

Example 7.13

In a variable reluctance-type inductive transducer, the coil has an inductance of 5 mH. When the iron piece is 1.5 mm and it is moved towards an electromagnet by 0.025 mm, determine the coil inductance.

Solution

$$
\begin{aligned}
L_1 &= 5 \text{ mH} \\
l_{g1} &= 1.5 \text{ mm} \\
\text{Length of air gap} &= 1.5 - 0.025 = 1.475 \text{ mm.} \\
\text{Inductive of coil} &= L_1 + \Delta L = 5 \times \frac{1.5}{1.475} = 5.08 \text{ mH}
\end{aligned}
$$

Example 7.14

An LVDT produces an o/p voltage of 2.6 V for displacement 0.4 mm. Calculate the sensitivity of LVDT.

Solution

$$
\text{Sensitivity, } S = \frac{\text{RMS value of o/p voltage}}{\text{Displacement}} = \frac{2.6}{0.4} = 6.5 \text{ V/mm}
$$

Example 7.15

The output of LVDT is 1.25 V at maximum displacement. At a load of 0.75 kgf, the deviation of linearity is maximum and it is ±0.0025 V. Determine the linearity at a given load.

Solution

$$
\begin{aligned}
\text{Max. deviation of linearity, } D_{\max} &= 0.0025 \text{ V} \\
V_0 &= 1.25 \text{ V} \\
\text{Linearity} &= \frac{D_{\max}}{V_0} \times 100 \\
&= \frac{0.0025}{1.25} \\
&= 0.2\%
\end{aligned}
$$

Example 7.16

An LVDT has a secondary voltage of 5 V for a displacement of ±12.5 mm. Determine the output voltage for a displacement of 8 mm from its central position.

Solution

$$\text{Sensitivity, } s = \frac{5}{12.5} = 0.4 \text{ V/mm}$$

$$\text{for 8 mm displacement, o/p voltage} = 0.4 \times 8 = 3.2 \text{ V}$$

Example 7.17

An LVDT is used to measure the deflection of bellows. The sensitivity of LVDT is 40 V/mm. The bellows are deflected by 0.125 mm by a pressure 0.8×10^6 N/m^2. Determine the sensitivity of LVDT and the pressure when the o/p voltage of the LVDT is 3.5 V.

Solution

$$S = 40 \text{ V/mm}$$

$$V_0 = 0 \times 0.125$$

$$= 5 \text{ V}$$

$$\text{LVDT sensitivity} = \frac{v}{p} = \frac{5}{0.8 \bullet 10^6} = 6.25 \times 10^{-6} \text{ V/N/m}^2$$

$$p = \frac{\text{o/p voltage}}{\text{LVDT sensitivity}}$$

$$= \frac{3.5}{6.25 \times 10^{-6}}$$

$$= 5.6 \times 10^5 \text{ N/m}^2$$

Example 7.18

A capacitive transducer with its plate separation of 0.05 mm under static condition has a capacitance of 5×10^{-12} F. Determine the displacement that causes a change of capacitance of 0.75×10^{-12} F.

Solution

$$d = 0.05 \text{ mm}$$

$$C = 5 \times 10^{-12} \text{ F}$$

$$C = \frac{\in A}{\text{d}}$$

$$\in A = Cd = 0.25 \times 10^{-15} \text{ F}$$

$$\Delta C = 0.75 \times 10^{-12} \text{ F}$$

$$\Delta x = \frac{\in A}{\Delta C} = 0.333 \text{ mm}$$

Example 7.19

A capacitive transducer uses two quartz diaphragms of area 600 mm^2 separated by 2.5 mm. A pressure of 8×10^5 N/m^2 applied to the top of diaphragm, causes a deflection of 0.5 mm. The capacitance is 400×10^{-12} F. When no pressure is applied, determine the value of capacitance.

Solution

$$A = 6 \times 10^{-4} \text{ m}^2$$

$$d = 2 - 5 \times 10^{-3} \text{ m}$$

$$C = 400 \times 10^{-12} \text{ F}$$

$$C = \frac{\in A}{\text{d}}$$

$$\in = \frac{Cd}{A} = \frac{1}{6} \times 10^{-8} \text{ f/m}$$

$$d^1 = d - \text{deflection}$$

$$= 2 - 5 - 0.5 = 2 \text{ mm}$$

$$C^1 = \in A/d^1 = 500 \times 10^{-12} \text{ F}$$

Example 7.20

A capacitance transducer has two parallel plates of an overlapping area of 5×10^{-4} m^2. The capacitance is 9.5 pF. Calculate the separation between the plates and sensitivity.

Solution

$$\in r = 81, \in_0 = 8.88 \text{ JGF/cm}$$

$$A = 5 \times 10^{-4} \text{ m}^2$$

$$C = 9.5 \text{ pf}$$

$$c = \in_0 \in_r A/d$$

$$d = \in_0 \in_r A/C$$

$$= 37.75 \times 10^{-3}$$

$$\text{Sensitivity, } S = \frac{\delta c}{\delta d} = \frac{-\varepsilon_0 \varepsilon_r A}{d^2}$$

$$= 0.025 \times 10^{-8} \text{ F/m}$$

Example 7.21

A 5-plate transducer has plates of dimensions 20 mm × 20mm and are 0.25 mm apart. Determine the sensitivity of the arrangement (Fig. 7.30).

Solution

$$\text{No. of plates, } n = 5$$

$$\text{Area, } A = 20 \times 20$$

$$= 400 \text{ mm}^2$$

$$\text{Relative dielectric constant, } \in_r = 1$$

$$\text{Absolute dielectric constant, } \in_0 = 8.854 \times 10^{-12} \text{ F/m}$$

$$d = 0.25 \text{ mm}$$

$$\text{Sensitivity, } S = \frac{-(n-1)\in_0 t_r A}{d}$$

$$S = \frac{-(5-1) \bullet 8.854 \bullet 10^{-12} \bullet 1 \bullet 20 \bullet 10^{-3} \bullet 20 \bullet 10^{-3}}{0.25 \bullet 10^{-3}}$$

$$= 2.833 \times 10^{-9} \text{ F/m}$$

7.7 SUMMARY

A transducer is a device that converts energy or input in one form to output energy in the same or different form. Active transducers do not need any external excitation to give electrical output from mechanical input. Passive transducers need AC/DC external input to give electrical output. Passive transducers are also classified as resistance, capacitance, and inductance transducers. The performance of a transducer in terms of the output given or sensitivity depends on physical conditions, environmental conditions, compatibility of associated equipment, and transducer parameters.

Using potentiometric transducer, parameters like force, displacement, etc. can be measured. Resistance strain gauges are classified as bonded and unbonded, based on the types of mounting. Strain gauges are sensitive to parameters like temperature, humidity, etc. A Wheatstone bridge circuit is used with a strain gauge as one element of the bridge and a similar strain gauge as the dummy gauge for measurements. Gauge factor is the ratio of unit change in resistance to unit change in length. If Poisson's ratio of the metal is known, the gauge factor (k) of the strain gauge can be determined. The relation between gauge factor and Poisson's ratio (μ)

$$\text{G.F.}(k) = 1 + 2\mu$$

Semiconductor strain gauges have large gauge factors, as high as 200.

Points to Remember

- Transducer converts a physical quantity like force, pressure flow rate, etc. to electrical quantity.
- Active transducers do not need any external excitation to give electrical output. E.g. thermocouples and piezoelectric transducers.
- Passive transducers can give electrical output with electrical input when given.
- Potentiometric transducers are used to determine physical parameters like force, pressure, displacement, etc. with wiper movement of potentiometer coupled to the parameter to be measured.
- The maximum percent error due to nonlinearity of the potentiometer transfer characteristic is

$15\frac{R_p}{R_m}$, where R_p is the potentiometer resistance for a given wiper position and R_m the maximum value of resistance.

- Maximum value of excitation voltage that can be given to a potentiometric transducer is Max $e_{ex} = \sqrt{PR_P}$, where P is the power rating of the transducer.
- Based on mounting of strain gauge wire, transducers are also classified as bonded type and unbonded type.
- Gauge factor = $(\Delta R/R)/(\Delta L/L) = 1 + 2\mu$, where μ is Poisson's ratio of the metallic wire used for the transducer; semiconductor strain gauges have high value of G.F in the range of 100–200.

Objective-type Questions

1. Transducers are broadly classified into _______ types. They are _______.
2. Passive transducers are broadly classified into _______ types. They are_______.
3. The type of resistive transducer that can be employed for the measurement of light intensity is _______.
4. Magnetostrictive effect is _______.
5. _______ types of strain gauges have large gauge factor.
6. The relation between gauge factor and Poisson's ratio is _______.
7. The lead wire that is used for strain gauges below 75°C is _______.
8. Thermistors have _______type of temperature coefficient.
9. The two types of hot wire anemometer are _______.
10. The chemical substance that is used in an electrotype hygrometer is _______.
11. The range of displacement that LVDT can measure is _______.
12. Naturally occurring piezoelectric materials are _______.
13. Strain gauge cements are used for joining the _______ to the _______.
14. For strain gauge lead wires, the recommended lead wire insulation material for the temperature below 75°C is _______.
15. The material that is used for strain gauge lead wire insulation in the temperature range 75°C and 260°C is _______.
16. The material chosen for the filament wire in a strain gauge should have _______ gauge factor.
17. The carrier material or the support material for strain gauges should have _______ dielectric strength.
18, The strain gauge is affected by temperature variation due to the strain-produced differential thermal expansion existing between _______ and _______.
19. The effect of temperature on metallic strain gauges is _______.
20. The parameter that can be measured using dielectric gauges is _______.

Review Questions

1. Derive the expression for estimating the error due to loading effect in the case of a potentiometric transducer.
2. Discuss the linearity, resolution, and noise considerations in the case of potentiometric transducers.
3. Derive the expression for gauge factor in the case of a strain gauge.
4. Give the constructional details of different types of bonded strain gauges.
5. Derive the expression for the output voltage in the case of (i) ballast circuit and (ii) Wheatstone bridge circuit for strain gauges.
6. Explain the principle and operation of semiconductor strain gauges.

Unsolved Problems

7.1 A linear potentiometer-type transducer is 60 mm long. It is uniformly wound with a total resistance of the wire being 6 kΩ. Under normal conditions, for zero displacement, the wiper of the pot is at the centre. Determine the linear displacement, if the measured resistance of the pot is 2 kΩ, when the wiper moves due to the measurement of displacement.

7.2 Determine the resolution of an angular potentiometer if the pot is having 40 turns per mm. The gearing arrangement in the transducer is such that the notion of the main shaft by one revolution causes 4 revolutions.

7.3 If the gauge factor of a strain gauge is 4.1, determine the Poisson's ratio.

7.4 The wire of a strain gauge is 100 mm long. The resistance without the application of any force is 150 Ω. Can the application of force, the resistance of the strain gauge change by 0.3 Ω. Determine the gauge factor of the device. The change in length of the wire on application of force is 0.1 mm.

7.5 A strain gauge is having a resistance of 800 Ω. Its gauge factor is 4. It is bonded to member structure under tensile stress. If its resistance changes by 2 Ω, due to the application of stress, determine the percentage strain.

7.6 A steel member is stressed to 1200 kgf/cm^2; Young's modulus of elasticity of steel is 2.1×10^6 kgf/cm^2. Calculate the percentage change of resistance of strain gauge if the value of the gauge factor is 2.4.

7.7 A strain gauge with a gauge factor of 2.6 is subjected to a stress of 1400 kgf/cm^2. The value of *E*, Young's modulus of elasticity is 2.7×10^6 kgf/cm^2. Determine the percentage change in resistance of the strain gauge.

7.8 A strain gauge is connected in series with a resistance of 500 Ω. The resistance of the gauge under unstrained conditions is 300 Ω. Its gauge factor is 2.4. The supply voltage given is 20 V. What is the change in output voltage when a stress of 150 mN/m^2. *E* of the member under stress is 250 N/m^2.

8

Other Types of Transducers

Introduction • Resistance thermometers • Semiconducting-resistance temperature transducers (thermistors) • Hot wire anemometer • Other variable resistance transducers • Variable inductance transducers • Synchros • Variable reluctance accelerometer • Temperature measurement • Thermocouples • Platinum resistance thermometers • Special resistance thermometer • Thermistors • Digital temperature-sensing system • Miscellaneous transducers • Area flow meters • Positive displacement meters • Magnetic flow meter • Variable capacitance transducers • Piezoelectric transducer • Magnetostrictive transducers • Liquid-level measurement • Ultrasonic-level gauge • Measurement of humidity and moisture • Photoconductive cells • Photo pulse pickup • Digital encoders and encoder transducers • Fibre optic displacement transducer • DC tachnometer generators for rotary velocity measurement • Force measurements • Electromechanical methods • Measurement of pressure • Elastic transducers • High-pressure measurement • Low-pressure measurement (vacuum measurement) • Temperature measurements • Data acquisition systems • Summary

8.1 INTRODUCTION

In the previous chapter, different types of transducers, their classifications, some active and passive transducers, and their principle of working are explained. In this chapter, instruments to measure other physical parameters like force, pressure, velocity, humidity, moisture, speed, proximity, and displacement are described. Their principle of working and operation is explained. A Data Acquisitions System (DAS) that employs various transducers, sensors and measuring instruments to collect data, and to measure various parameters is also described . It analyses and indicates the corrective action to be taken and implemented.

8.2 RESISTANCE THERMOMETERS

A resistance thermometer consists of a resistive element exposed to the environment, whose temperature is to be measured. The electrical resistance of various materials changes in a reproducible manner with temperature, thus forming the basis of a temperature-sensing method. Materials normally used for temperature measurement are broadly classified into

1. Conductors.
2. Semiconductors.

If conductors are used to transduce the temperature they are known as resistance thermometers, and if semiconductors are used then they are known as thermistors.

Conductive elements used in thermometers are

1. Platinum.
2. Nickel.
3. Copper.
4. Tungsten.
5. Nickel/iron alloys.

The variation in resistance R with temperature T for metallic materials can be represented by an equation of the form

$$R_T = R_o (1 + a_1 T + a_2 T^2 + \cdots + a_n T^n)$$

where R is the resistance at $T = 0$.

For engineering purposes and also if the range of variation in temperature is narrow then

$$R_t = R_o \left(1 + \overline{\alpha t - t_o}\right)$$
$$= R_o (1 + \alpha \Delta t)$$

where α is the temperature coefficient at t_o and R_o is the resistance at t_o.

For large temperature ranges, the resistance equation follows the equation form more accurately

$$R = R_o (1 + \alpha \Delta t + \beta \Delta t^2)$$

The variation in resistance is measured and converted into a voltage signal with the help of a bridge circuit. Bridge circuits may employ either the deflection mode of operation or the null (manually or automatically balanced) mode. The output voltage produced for different elements used in resistance thermometer due to change in resistance in shown in Fig. 8.1. Figure 8.2 shows the bridge circuit for null method of measurement. R_4 is varied until balance is achieved. When better accuracy is required the arrangement shown in Fig. 8.2(b) is preferred.

Here the contact resistance in the adjustable resistor has no influence on the resistance of the bridge legs. If long lead wires subjected to temperature variations are unavoidable, then circuits shown in Fig. 8.3(a) and (b)are utilized. To get a fairly linear relationship between the output voltage and the temperature, the values of R_1 and R_2 of the above circuits are made at least 10 times greater than that of the thermometer.

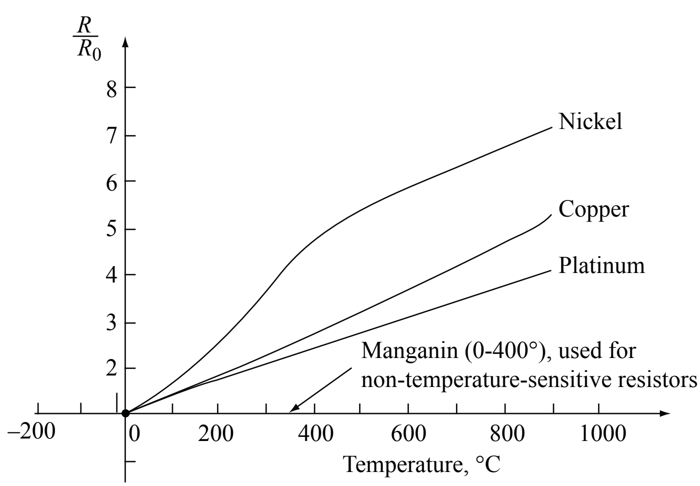


Figure 8.1 Output voltage with temperature for different elements

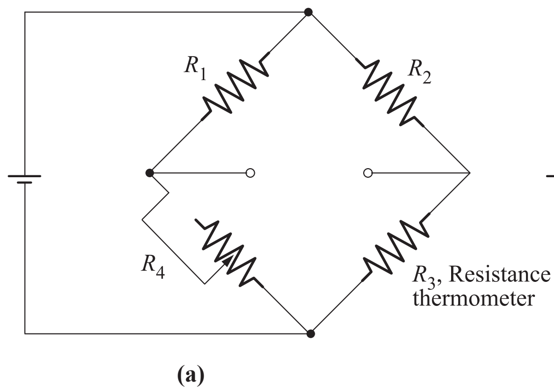


Figure 8.2 Bridge circuits for null method

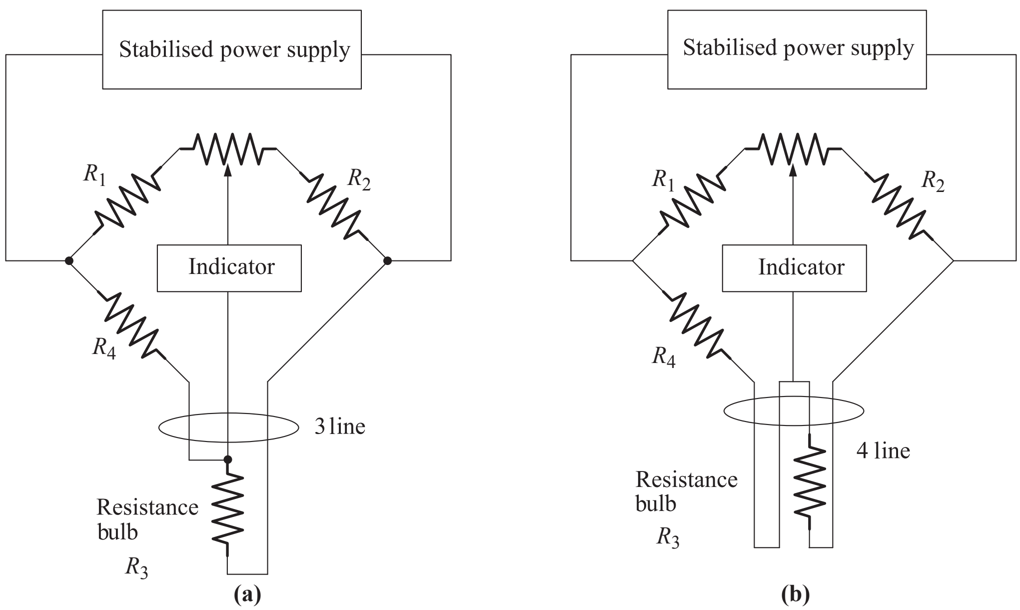


Figure 8.3 Resistance thermometer circuits

8.2.1 Self-Heating

Resistance thermometer bridges may be excited with either DC or AC. The direct or *rms* alternating circuit through the thermometer is usually in the range of 2–20 mA. This current causes an I^2R heating, which raises the temperature of the thermometer more than the surrounding temperature, causing the so-called self-heating error. The magnitude of this error also depends on heat transfer conditions, which is usually small. A 450-Ω platinum element of open construction carrying 25-mA current has a self-heating error of 0.01 °C. To reduce the error due to self-heating to get a higher output voltage at the same time, we can excite the bridge with a pulse whose *rms* value is small compared with its peak value. Such a pulse excitation voltage can be obtained by commutating a DC source as shown in Fig. 8.4, and this also allows timesharing of the bridge among several resistance sensors. As much as a 5-V full-scale bridge output signal can be obtained from resistance sources used in this way. The element resistance of thermometers ranges from about 10 Ω to as high as 25,000 Ω.

The choice of elements for the resistance thermometer depends on the temperature range of operation. Table 8.1 gives the temperature range for different elements.

Table 8.1 Temperature range of operation

Element	Temperature Range
Platinum	230–1010 °C
Copper	160–260 °C
Nickel	260–426 °C
Tungsten	230–1095 °C

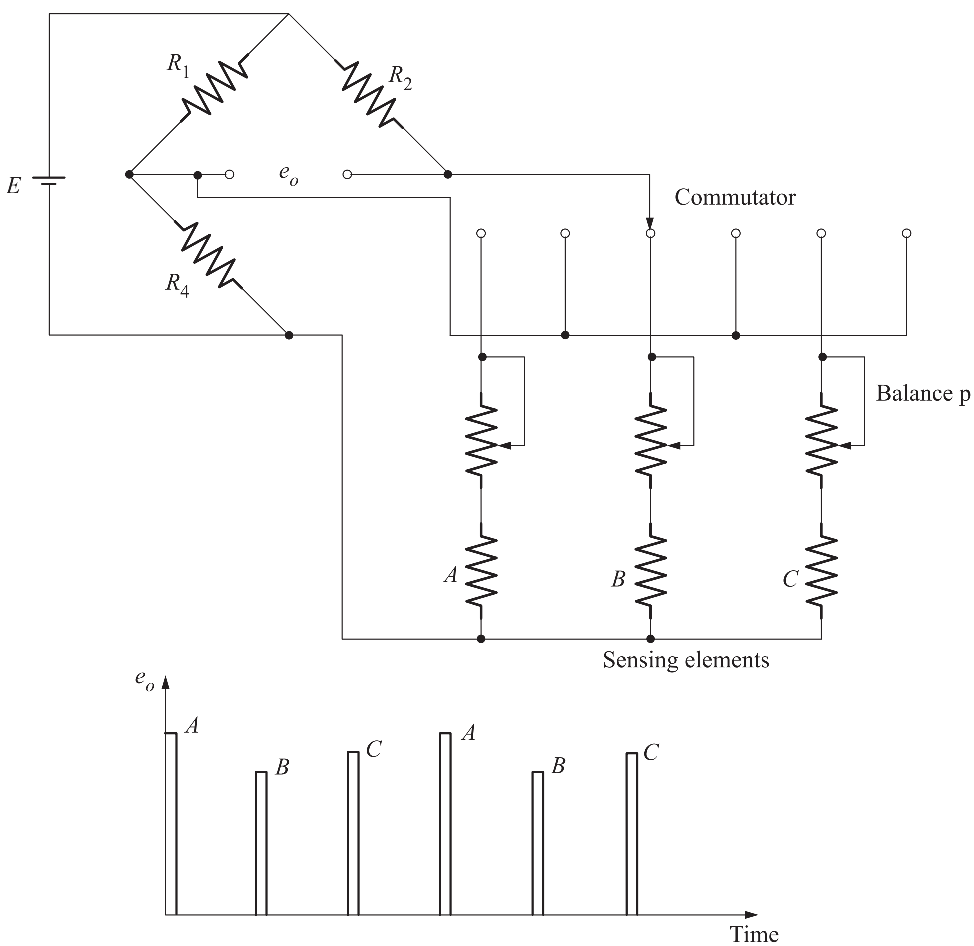


Figure 8.4 Circuits and waveforms

8.2.2 Advantages of Wire Resistance Thermometers

1. They are suitable for measuring large temperature differences and high temperatures.
2. They are very accurate, which makes them suitable for small temperature measurement.
3. Well-designed resistance thermometers have excellent stability.
4. Unlike thermocouples, they do not need a reference junction and this favours them in many aero space and some industrial applications.

8.2.3 Disadvantages

1. The relatively large volume of resistance thermometers compared to thermocouples results in monitoring an average temperature over the length of the resistor, rather than a temperature at a point.
2. They need auxiliary apparatus and power supply.
3. The resistance element is usually more expensive than a thermocouple.
4. Self-heating and thermoelectric effect of the resistive element and connecting leads (dissimilar metal junctions) cause errors.

8.3 SEMICONDUCTING-RESISTANCE TEMPERATURE TRANSDUCERS (THERMISTORS)

Thermistors are thermal resistors with a high negative temperature coefficient of resistance (positive temperature coefficient resistant thermistors are also now available). They are made of Manganese, Nickel, Copper, Iron, Uranium, and Cobalt oxides, which are milled, mixed in proper proportions with binders, pressed into the desired shape, and sintered. The standard forms now available are (Fig. 8.5)

1. Leads.
2. Discs.
3. Probes
4. Washers.
5. Rods.

The resistivity may be from 10^{-1} to 10^{9} Ω-cm.

Thermistors have a large negative temperature coefficient and they are highly non-linear. The resistance at different temperatures can be found out using the following equation:

$$R_T = R_o \, e^{\beta\left(\frac{1}{T} - \frac{1}{T_o}\right)}$$

where R_T is the resistance at temperature T, R_o the resistance at temperature T_o, β a constant that is characteristic of a material, e the base of natural log, and T, T_o the absolute temperature in °K.

The value of β for the semiconductor made of the materials mentioned in Fig. 8.5 is 4000. The temperature coefficient α for a thermistor is expressed as

$$\alpha = \frac{1}{R}\frac{dR}{dT}\bigg|_{R=R_o}$$

$$\frac{dR}{dT} = R_e\beta\left(\frac{1}{T} - \frac{1}{T_o}\right)\frac{1}{T^2}$$

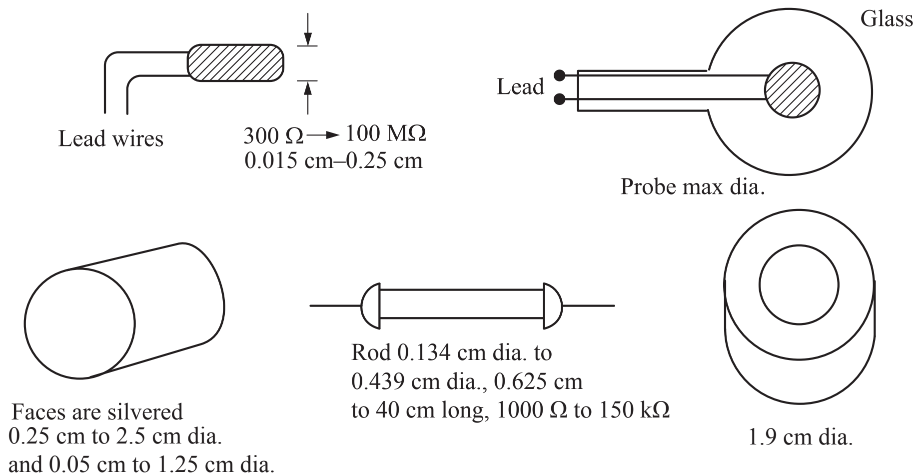


Figure 8.5 Thermistor structures

$$\frac{1}{R}\frac{dR}{dT} = -\frac{\beta}{T^2}$$

at $$25\ ^\circ\text{C}, \alpha = -\frac{4000}{298^2} = -0.045$$

The Resistivity versus Temperature graphs are shown in Fig. 8.6. The voltage to current characteristics of thermistors are shown in Fig. 8.7.

8.3.1 Circuit Employed

The circuit normally used is a Wheatstone bridge with the thermistor in one arm. Individual calibration is necessary for each circuit. A 3-Ω resistance is connected in parallel with the thermistor as shown in Fig. 8.8(a) to have a linearised relationship as shown in Fig. 8.8(b).

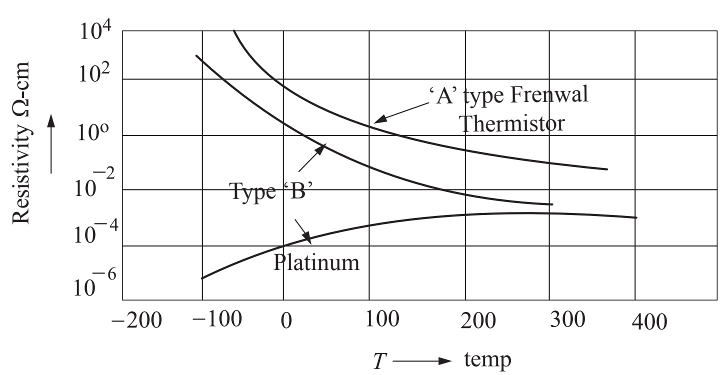
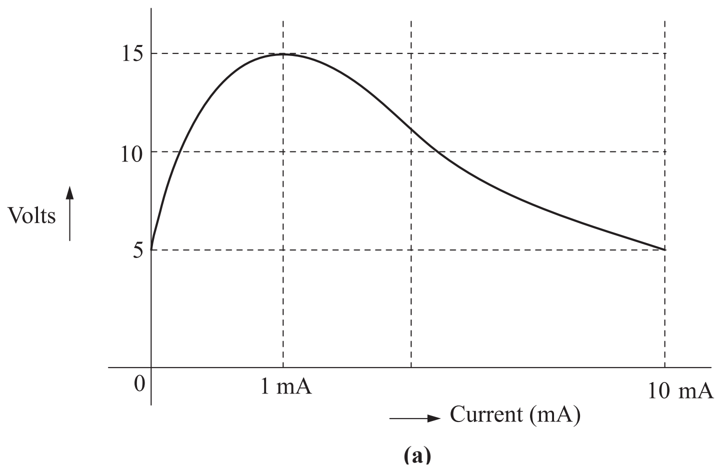


Figure 8.6 Resistivity variation with temperature for thermistors



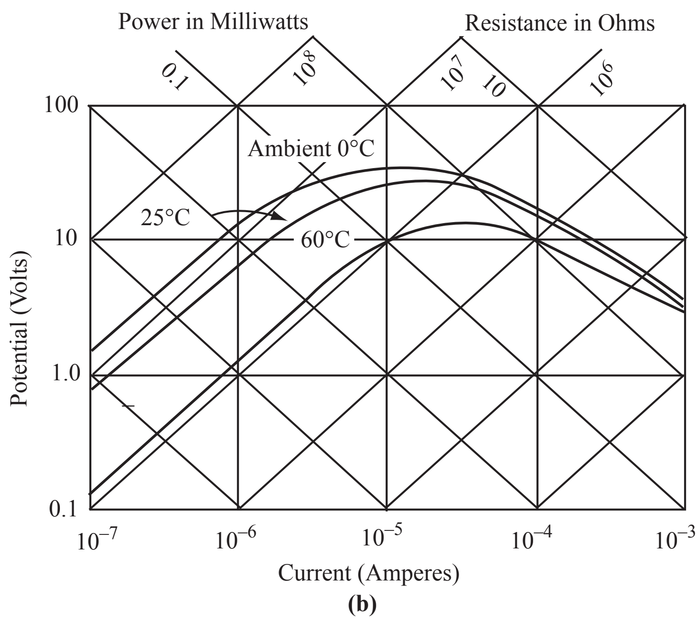


Figure 8.7 (a) *V–I* Characteristics of thermistors and (b) Voltage–current characteristics

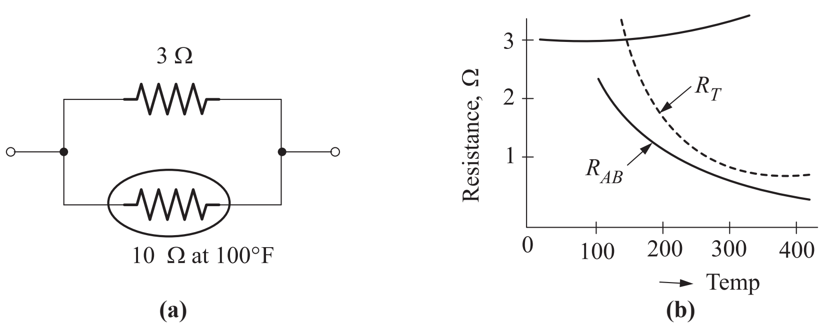


Figure 8. 8 Thermistor circuit and characteristics

8.3.2 Advantages

1. Semiconductor resistance thermometers have an extremely high temperature sensitivity (10 times higher than that of a metallic resistance thermometer).
2. It can be manufactured in almost any size or shape.
3. If a fast response time is desired flake thermistors can be made with thermal relation time in the order of few milliseconds.
4. Since resistance at the normal operating temperature is sufficiently high to make lead resistances appear negligible, thermistors can be used for remote indications.
5. They hold their original calibration for long periods of time, and so their stability at normal temperature is good.

8.3.3 Disadvantages

1. It is highly non-linear.
2. Its low temperature limit is set by insensitivity.
3. Its upper limit is set by instability.

Thermistors are particularly useful for narrow ranges of measurements at extreme temperature such as those of liquid oxygen, which normally vary from –170 °C to –185 °C. They are also useful in fuel and air-conditioning temperature measurements.

8.4 HOT WIRE ANEMOMETER

A hot wire anemometer is a commonly used device for measuring the mean and fluctuating *velocities in fluid flows*. The flow sensing element is a short length of 5 μm diameter platinum–tungsten wire welded between two prongs of the probe and heated electrically as a part of a wheat stone bridge. When the probe is introduced in the fluid stream, it tends to get cooled by the instantaneous velocity and consequently there is a decrease in its resistance. The rate of cooling of the wire depends on the following:

1. Shape, size, and physical properties of the hot wire.
2. Difference in temperature between the heated hot wire and the fluid stream.
3. Physical properties of flowing fluid.
4. Velocity of the fluid stream.

Generally, the first three conditions are effectively constant in the hot wire operation and the instrument's response is then a direct measure of the flow velocity. The wire needs to be strong enough to give adequate resistance and have an extremely small thermal capacity in order to follow the fluctuations in velocities faithfully and with infinitesimal time lag.

The usual sizes employed are of 1–2 mm length and 2–5 m in diameter (Fig. 8.9(a)).

It may be noted that the accumulated dirt on the wire can lead to serious heat transfer errors. Further, the hot wire system is suitable for only gas flow measurements. For liquid flows, hot film probes (Fig. 8.9(c)) are employed but the associated electronic circuitry is the same as for the hot wire probe. The hot film is usually 5-μm, thick and is coated with a thin layer of epoxy to avoid short circuit in liquid flows. Films are usually of 'gold, platinum' deposited by vacuum deposition/sputtering, etc. Pyrex or similar glass is a universal material for the support (substrate) for such probes.

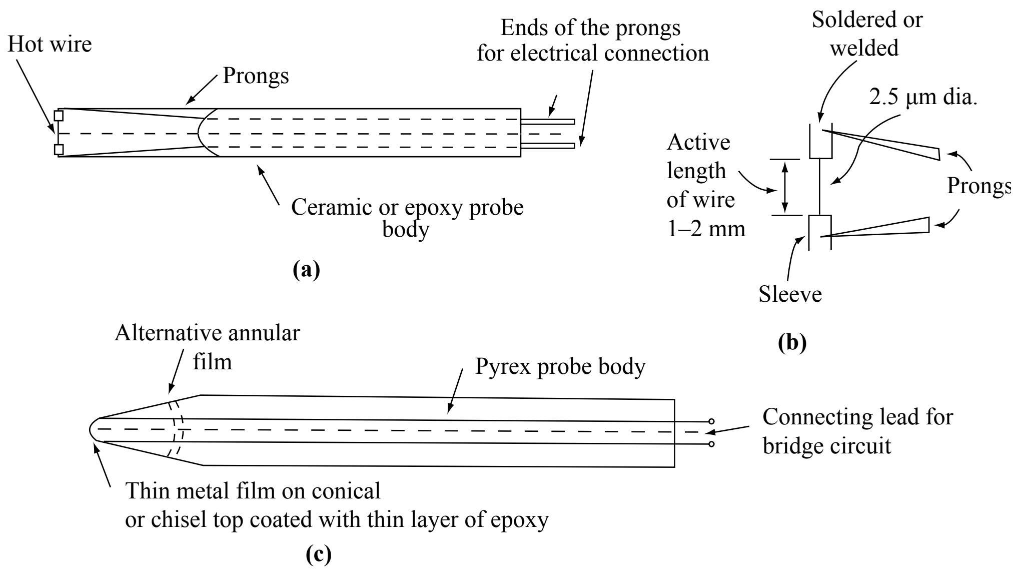


Figure 8.9 Typical hot wire/film anemometer probes: (a) a typical hot wire probe (b) details of the wire of probe and (c) hot film probe

Anemometers are devices used for measurement of velocity of flow. Hot wire anemometers use resistance wire as a sensor. They can be broadly classified into two categories:

(1) Constant current type.

(2) Constant temperature type.

8.4.1 Constant Current Type

In Constant Current Type (CCT) a fine-resistance wire carrying a fixed current is exposed to the flow velocity. The wire attains an equilibrium temperature when the I^2R heat generated in it is equally balanced by the convective heat loss from its surface. Here the circuit is designed so that I^2R heat generated in it is balanced and I^2R heat is essentially constant. Therefore, the wire temperature must adjust itself to change the convective heat until equilibrium is reached. This wire temperature is a measure of the flow velocity, which can be measured in terms of its electrical resistance (Fig. 8.10). Fluctuating flow with a frequency of 160 Hz and above cannot be measured using this type of anemometer.

In constant temperature form, the current through the wire is adjusted to keep the wire temperature constant. The current required to do this is a measure of velocity. For measurement of average velocity, the constant temperature mode of operation is often used. As far as the steady flow is concerned, the setup in Fig. 8.11 is suitable. However when there is large fluctuation in the velocity of flow, we cannot use this circuit.

The constant temperature type (CTT) of instrument used for measuring steady velocities can be extended to measure both average and fluctuating components of velocity by making bridge-balancing operation automatic rather than manual through the use of feedback as shown in Fig. 8.12.

With zero-flow velocity, the bridge excitation is shut off ($i_w = 0$) and the hot wire assumes the fluid temperature. The variable resistor R_3 is then manually adjusted so that $R_3 > R_w$, thereby unbalancing

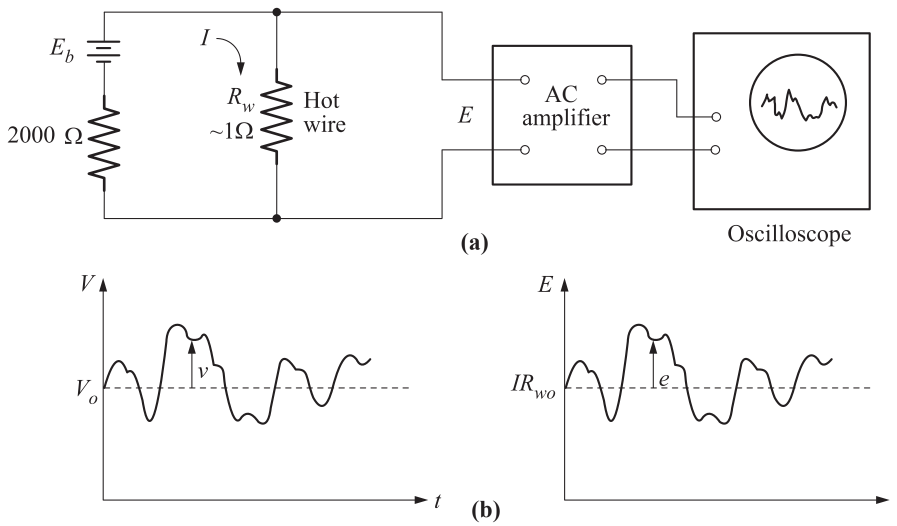


Figure 8.10 (a) Circuit for a hot wire anemometer (CCT) and (b) waveforms

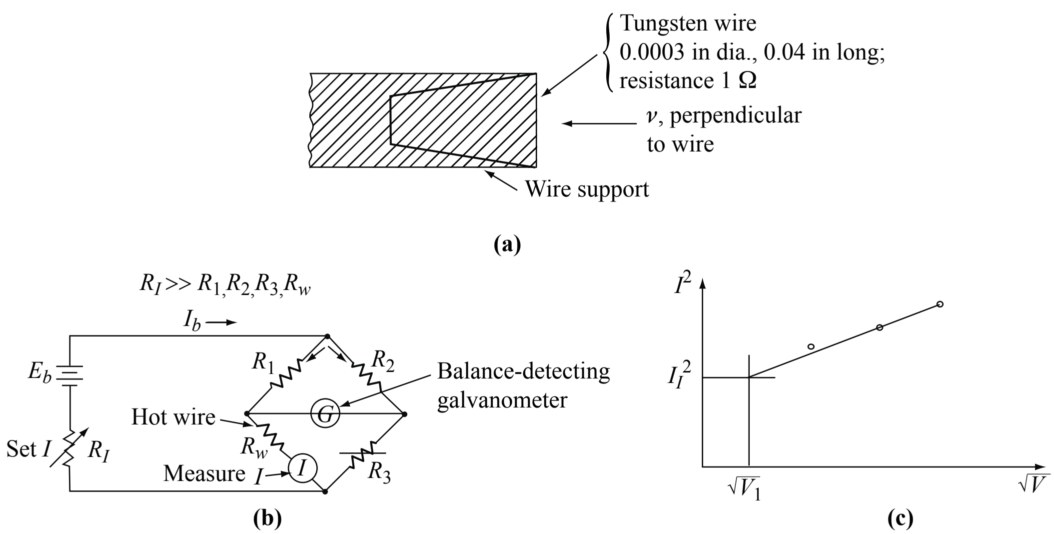


Figure 8.11 (a) Transducer (b) bridge circuit and (c) graph

the bridge. When the excitation current is turned on, the unbalanced bridge produces an unbalance voltage e_0, which is applied to the input of a high-gain current amplifier supplying the bridge excitation current.

The current now flowing through R_w increases its temperature and thus its resistance. As R_w increases, it approaches R_3 and the bridge unbalance voltage e_e decreases. As the actual amplifier has limited gain, the bridge unbalance voltage cannot be zero.

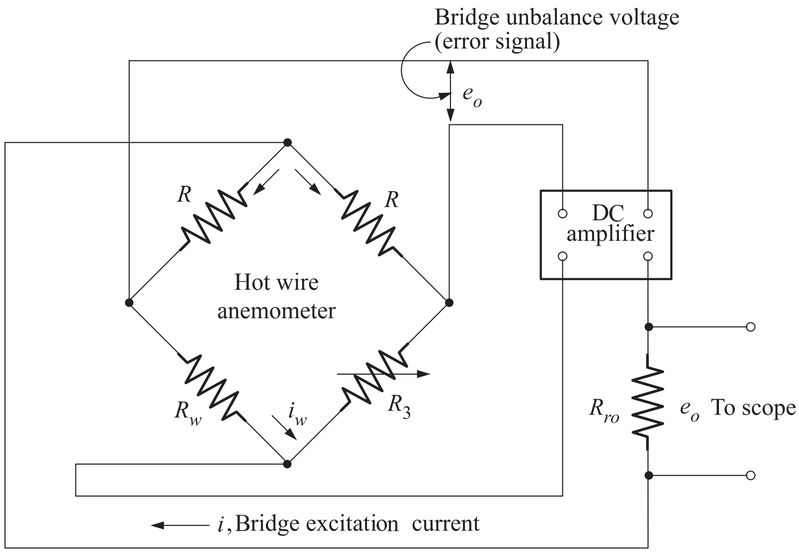


Figure 8.12 Bridge circuit for a hot wire anemometer (constant temperature type)

A typical instrument has a flat (within 3 dB) frequency response up to 17,000 Hz when the average flow velocity is 10 m/sec, 30,000 Hz for 33.3 m/sec, and 50,000 Hz for 100 mps. A commonly used measuring circuit for a hot wire anemometer, CTT type, is shown in Fig. 8.12

8.4.2 Measurement of Direction of Average Flow

A single wire responds essentially to the component of velocity perpendicular to it, if the angle between the wire and the velocity vector is between 90 and 25 °C. For this range V is to be replaced by $V \sin \theta$.

If $\theta < 25°$ the heat loss is greater than that predicted by $V \sin \theta$, if $\theta = 0$ it is about 55% of that for $\theta = 90°$. When the probe is mounted on a rotatable base the flow direction angle could be found by determining the probe rotation angle, which gives a maximum value of 1. This method is quite inaccurate as θ changes very slowly near 90°. A better procedure (if the flow angle is roughly known) is as follows. The wire is at about 50° from the flow direction and I is measured. The probe is then rotated in the opposite direction until an angle is found at which the same value for I is measured. The bisector angle between the two locations determines the flow direction. This method is more accurate since the rate of change of I with θ is at a maximum near 50°.

Figure 8.13 illustrates a special construction of a rotatable probe. The angle between the two wires is 90°. These two wires are connected in adjacent arms as shown in the figure. The probe is rotated until the bridge is balanced and the bisector of the two wires at the null of the bridge is the direction of flow.

Comparative features of constant current and constant temperature anemometers are as follows:

1. In CCT, the current must be set high enough to heat the wire considerably above the fluid temperature for a given average velocity. If the flow should suddenly drop to a much lower velocity or come to rest, the hot wire will burn out since the convection loss cannot match the heat generation before the wire temperature reaches the melting point. This drawback is not there in CTT as the temperature of the wire is to be kept constant irrespective of the flow variations.
2. If the average value of the flow velocity changes, the network must be reset in CCT, whereas this problem is not there in a constant temperature anemometer.

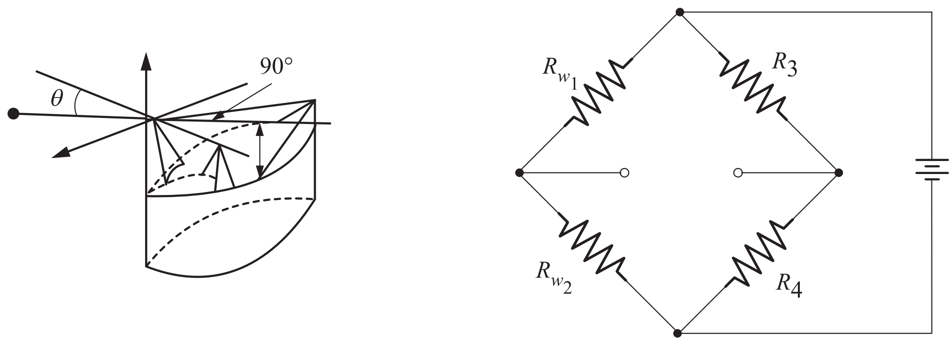


Figure 8.13 Rotatable probe and bridge circuit

3. The error in CCT is more if the fluctuation is more than 5% CTT and cannot be affected by any fluctuation.
4. As the CTT uses a DC amplifier, it is usable down to zero frequency (steady velocity), whereas constant current type is good only to about 1 Hz.
5. In CTT the noise level is more and there is difficulty in designing sufficiently high-gain DC amplifiers without causing instability and drift problem.

So far in the analysis the fluid density ρ was assumed constant. However, if ρ varies, the anemometer actually measures the product ρ_v that is the local mass flow rate:

$$I^2 \ \alpha \ \sqrt{\rho_v}$$

The problems encountered in the applications of hot wire anemometers are as follows:

1. Due to limited strength of fine wire, larger dirt particles striking the wire may actually break them.
2. Unless the flow is clean, significant calibration changes can occur in relatively few minutes of operation due to dirt accumulation on the wire.
3. At high speeds, wires may vibrate because of aerodynamic loads and flutter effects.

To overcome all the above practical difficulties, a hot film transducer was thought of. Here the resistance element is a thin film of platinum deposited on a glass base (see Fig. 8.14). The film takes

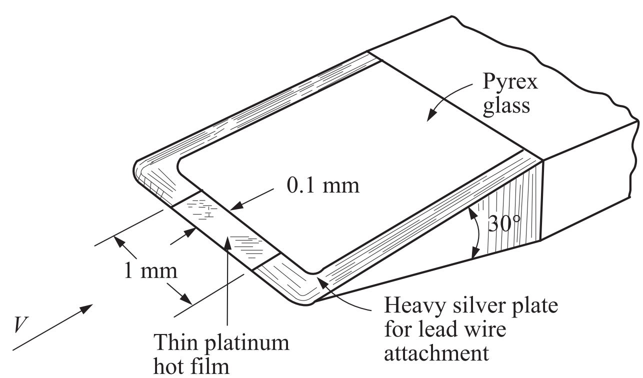


Figure 8.14 Hot film transducer

the place of the hot wire. The basic circuitry is similar to the constant temperature hot wire approach. The film transducers have great mechanical strength and may be used at very high temperature by constructing them with internal cooling water passages.

8.5 OTHER VARIABLE RESISTANCE TRANSDUCERS

8.5.1 Contact Pressure Transducer

The most common transducer based on variable contact pressure between carbon granules is the carbon microphone. The variation in resistance here is due to the variation of the surface area between granules in contact. The main advantage of the carbon microphone is its high output upto 5 V in ordinary types. Their frequency range is only 30–3000 Hz and their signal noise ratio is low.

As a transducer to measure pressure, we can use thin discs of carbon, graphite, or moulded carbonaceous material, which are arranged in stacks:

$$R = R_o + V_p$$

where R_o is the internal resistance of the solid disk, c is a constant, and p the pressure on a carbon file with a large number of thin discs. R_o is small in comparison with C/P so that total resistance R is roughly inversely proportional to the pressure p. The carbon diet material should have high mechanical strength, low mechanical hysteresis creep, and low-temperature coefficient of resistance.

In the measurement of pressures above 6 kg/cm^2, a carbon compound miniature resistor has been used quite successfully.

8.5.2 Humidity Measurement

A lithium chloride salt in a 3% concentration on polyvinyl-acetate film was used in the design of an electrolytic hygrometer. *This material changes its resistance with humidity change*. In order to avoid polarization, AC is preferred in the measurement of conductivity in these devices. The conductivity of this material varies to a large extent with temperature. Therefore, proper temperature compensation is a must. The materials developed recently as humidity-sensing elements are titanium dioxide, a ceramic material, and a conducting plastic of unspecified composition.

8.5.3 Light

Photoconductive cells are basically variable resistance transducers. The resistance changes due to photon impingement. There are basically of two types:

1. Bulk photo conductors.
2. *p–n* Junctions or rectifier types.

Photoconductive cells cover the entire spectrum from infrared to ultraviolet. Their sensitivity to infrared radiation is of great practical importance.

Lead sulphide can be used up to wavelengths of 3. 5 μ. Lead-telluride can be used up to wavelengths of 5.8 μ. Cadmium sulphide can be used up to 0.9 μ.

Selenium is widely used in indicators of low accuracy but it is less sensitive to infrared radiation than other materials and there is a longer time lag between illumination and resistance variation.

Semiconductor photo cells have non-linear and temperature-sensitive calibration curves and the choice of a suitable type depends very much on the particular job on hand.

8.6 VARIABLE INDUCTANCE TRANSDUCERS

Inductance, one of the three electrical circuit parameters (R, L, C) is related to a coil. Inductance of a coil depends on the number of turns, permeability of the material, and geometric factor that is $L = \mu_o \mu n^2 G$, where n is the number of turns of the coil, G is the coil geometry, μ is the relative permeability, and μ_o the permeability of air. Therefore, any quantity or variable that affects one of these factors can be transduced as an inductance variation, which forms the basis of the *variable inductance transducer.*

8.6.1 Linear Variable Differential Transformer (LVDT)

The LVDT is the most common mutual inductance element. This can be considered to be an optimum transducer element for most electromechanical measuring systems with regard to resolution, hysteresis, dynamic response, temperature characteristics linearity, and life.

LVDT consists of a primary winding, two identical secondary windings, and a ferromagnetic movable core. A carrier excitation of frequency 60 – 20,000 Hz and a magnitude of 3 – 15 V is applied to the primary. The symmetrically spaced secondary is connected externally in a series-opposing circuit. Motion of the non-contacting magnetic core varies the mutual inductance from each secondary to the primary, which determines the voltage induced from the primary to the secondary. If the core is centered between the secondary windings, the voltage induced in each secondary is identical and 180° out of phase so there is no output. If the core is moved off center, the mutual coupling of the primary with one secondary will be greater than with the other, and a differential voltage will appear across the secondaries in series. For off-center displacements within the range of operation, this voltage is essentially a linear function of displacement. The phase shift between excitation and the output varies with frequency of excitation. The origin of this phase shift can be seen from the following analysis, and the equivalent circuit of an LVDT is given in Fig. 8.15.

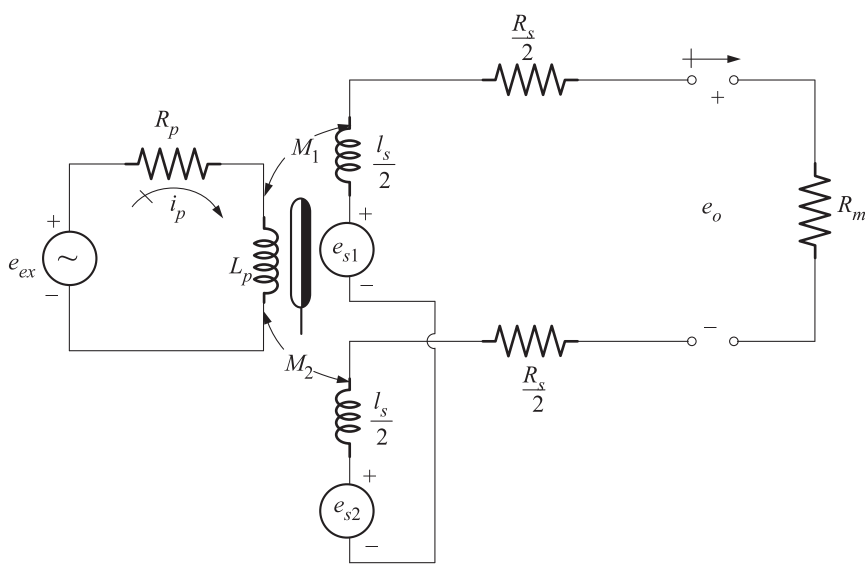


Figure 8.15 LVDT—equivalent circuit

The following equations are in order for the circuit:

$$I_p R_p + \frac{di_p}{dt} - e_{ex} = 0$$

$$e_{s1} = M_1 \frac{di_p}{dt} - e_{s2} = M_2 \frac{di_p}{dt}$$

$$e_s = e_{s1} - e_{s2} = (M_1 - M_2) \frac{di_p}{dt}$$

If the output is open circuited for a fixed core position

$$e_0 = e_s = (M_1 - M_2) \frac{D}{L_p D + R_p} e_{ex} ; \left(D = \frac{d}{dt}\right)$$

$$\frac{e_o}{e_{ex}}(D) = \frac{\left[(M_1 - M_2)/R_p\right]D}{T_p D + 1} ; \left(T_p = \frac{L_p}{R_p}\right)$$

In terms of frequency response

$$\frac{e_o}{e_{ex}}(j\omega) = \frac{(M_1 - M_2)/R_p}{\sqrt{(\omega L_p)^2 + 1}}$$

where 90° – $\tan^{-1} \omega T_p$. Therefore, the phase shift between e_o and e_{ex} is a function of the excitation frequency.

If a voltage-measuring device of input resistance R_m is attached to the output terminals, then the situation is different as follows:

$$\frac{(M_1 - M_2)D}{L_p D + R_p} e_{ex} = (R_s + R_m)\, i_s + L_s = \frac{di_s}{dt}$$

$$e_o = i_s R_m$$

$$\frac{e_o}{e_{ex}}(D) = \frac{(M_1 - M_2)R_m}{R_p (R_s + R_m)(L_p D + 1)(L_s D + 1)}$$

where

$$T_s = \frac{L_s}{R_s + R_m}$$

It is obvious from the above expression that the frequency response of $(e_o/e_{ex})\,(j\omega)$ has a phase angle of +90° at low frequencies and –90° at high frequencies. Somewhere in between it will be zero. By suitably adjusting the excitation frequency, one can achieve a desired phase shift. If for some reason the excitation frequency cannot be adjusted to this value, then one of the methods shown in Fig. 8.16 can be used. Now the output voltage at null may not be zero due to harmonics and stray capacitance coupling between the primary and the secondary. Under usual conditions, this is less then 1% of full-scale output voltage and may be quite acceptable. Fig. 8.17 shows some methods for null reduction. The values of R and R_p are not critical but should be as low as possible without loading the excitation source.

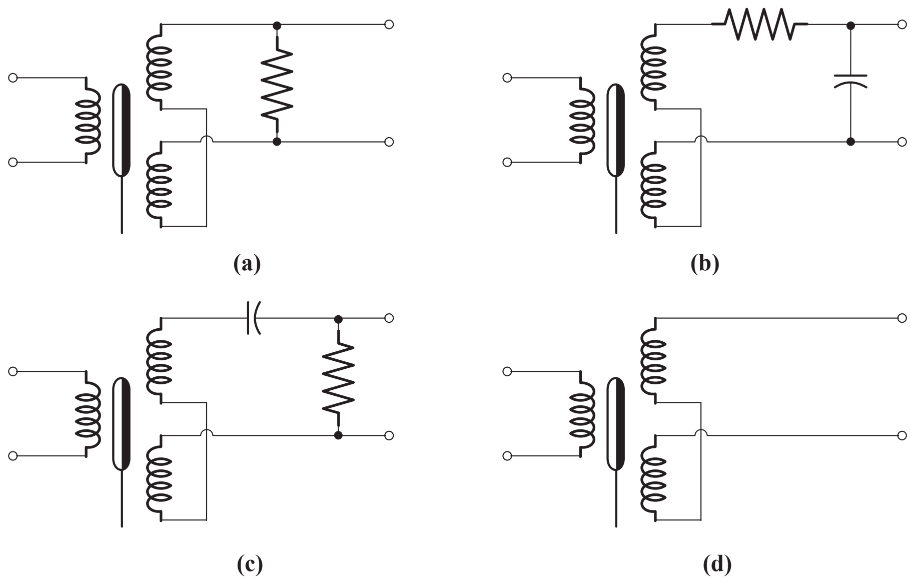


Figure 8.16 Four possible methods for retarding a leading phase angle

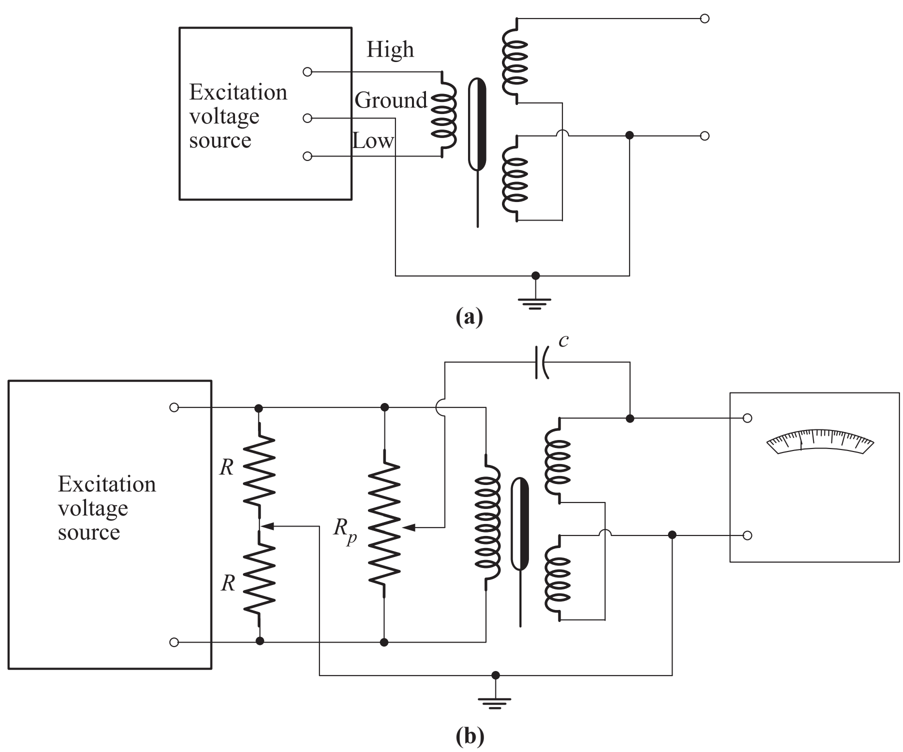


Figure 8.17 LVDT circuits for null reduction

The output of a differential transformer is a sine wave whose amplitude is proportional to the core motion. An AC voltmeter calibrated in motion units is alright for measurement of static or very slowly varying displacement except that the meter will give exactly the same reading for displacements of equal amount on either side. For direction sensitiveness and for rapid core motions, the arrangement shown in Fig. 8.18 is suitable. Figure 8.18 shows the circuit arrangement for phase-sensitive demodulation using semiconductor diodes. If the ratio of the excitation frequency to the core motion frequency is 10:1, a simple RC filter is adequate. The output of this filter then becomes the input to an oscillograph or oscilloscope. For better response of the LVDT system, a sharp cut-off filter may be necessary.

Resonant LVDT: In this device, the secondary coils are wound in a manner that markedly increases the distributed capacitance of the coils. The much higher capacity modifies the impedance of the coils and permits partial resonance at a relatively low excitation frequency. The result is exceptionally high output voltage for small core movement.

Rotary Variable Differential Transformer (RVDT): The heart of this device is literally a toroidal magnetic cam-core pivoted eccentrically on an input shaft across the centre of a coil form. The coil form contains two pairs of identical windings spaced symmetrically above and below the centre. Fig. 8.19 shows the details of this device and its output response characteristics. This device reacts to a rotary position similar to the LVDT, which reacts to linear position inputs. The maximum range of angular position measurement with reasonable linearity is about ±60°.

DC-LVDT: The concept of a DC-operated transducer element having all the advantages and unique features of the LVDT has always appealed to meteorologists and instrument users. In practice, such a transducer element is achieved by incorporating some form of signal conditioning electronics with an LVDT (Fig. 8.20).

The state of the art instrument was dramatically advanced, however, by the development and introduction of the hybrid microcircuit modules. The block diagram in Fig. 8.20 shows that the module includes a complete passive demodulator/DC amplifier signal conditioner. DC-LVDTs incorporating this module feature high sensitivity, low output impedance, excellent reliability, and extreme ruggedness.

8.6.2 Applications of LVDT

8.6.2.1 Displacement measurement and gauging. One common form of a completely assembled LVDT displacement transducer is illustrated in Fig. 8.21. It is commonly known as LVDT gauge head because it is widely used in machine tool inspection and gauging equipment. The core is connected to a spring-loaded probe shaft having a removable tip. The probe shaft is guided in a sleeve bearing that is retained in a case, which also encloses the LVDT coil winding. The case is often threaded externally to simplify mounting.

Lever or finger probe gauge heads are useful in places that are inaccessible to the probe of an ordinary LVDT. Figure 8.22 shows two possible arrangements. A parallel flexure type is preferred as this introduces less error due to core rotation.

Traditional users of LVDT gauge heads are manufacturers of machine tool inspection equipments and quality control departments. In these applications, the LVDT gauge head and its read-out act as more precise substitutes for mechanical dial position indicators. This type of head system can be used to record data electronically, to gather statistical quality-control information on manufactured parts, or to interface with a digital computer or numerically controlled machine tool in a closed-loop control

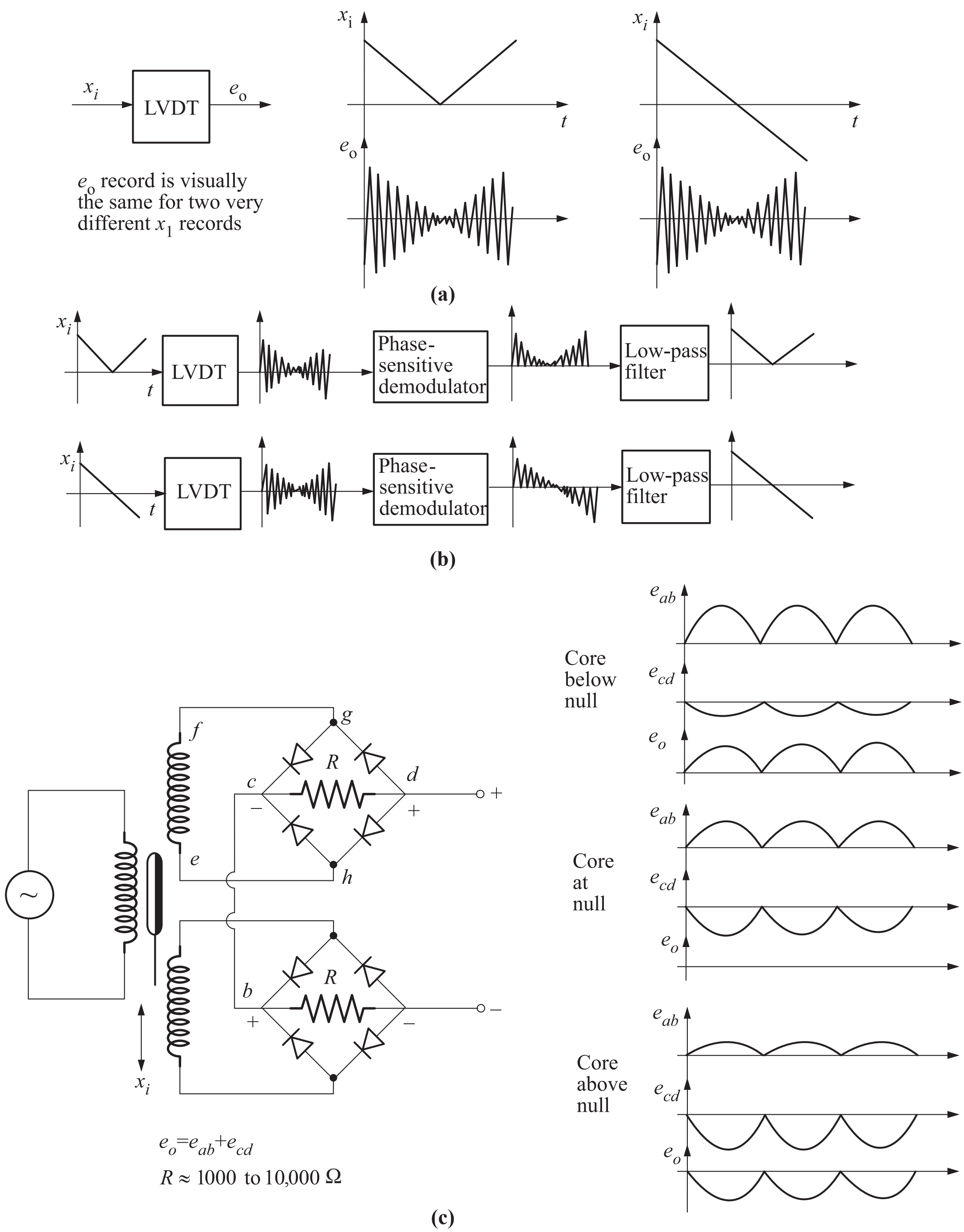


Figure 8.18 LVDT circuits for phase-sensitive demodulation and output waveforms

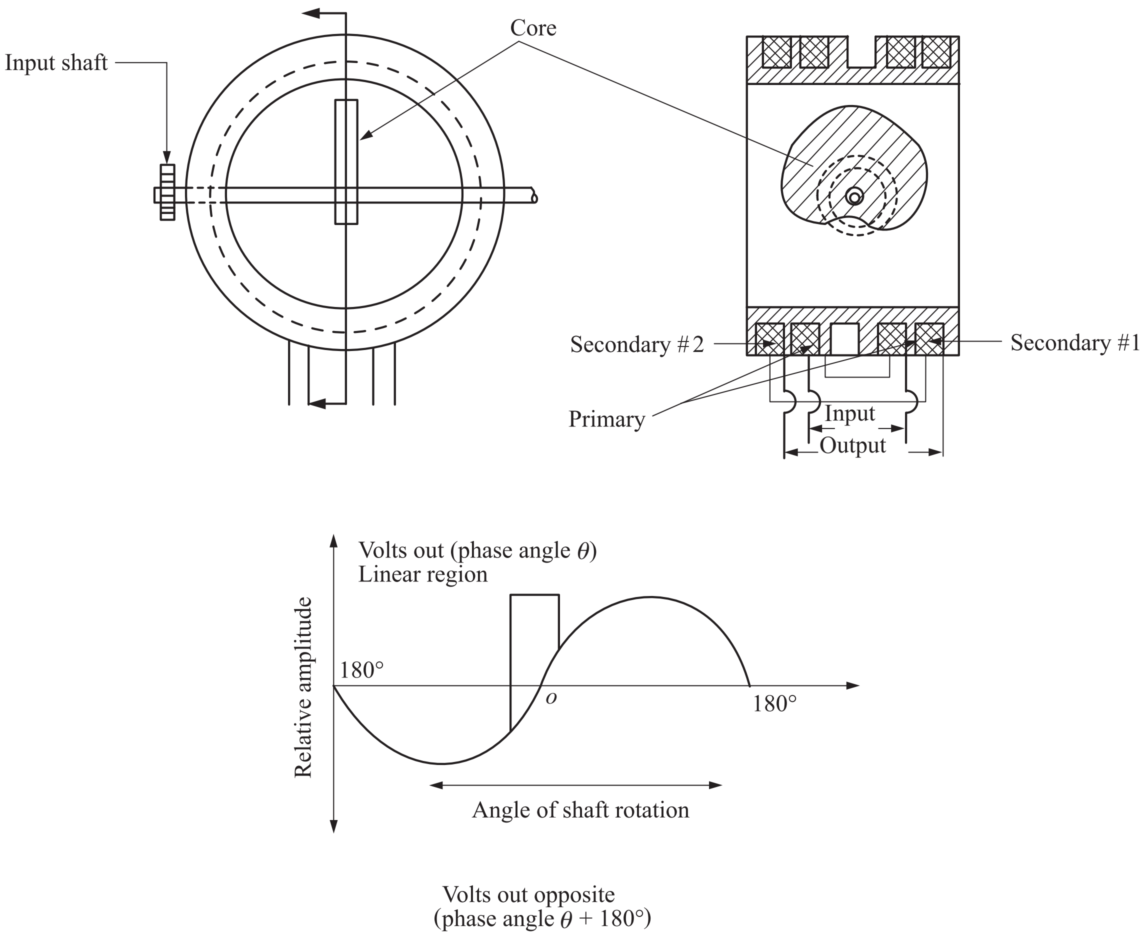


Figure 8.19 RVDT structure and waveforms

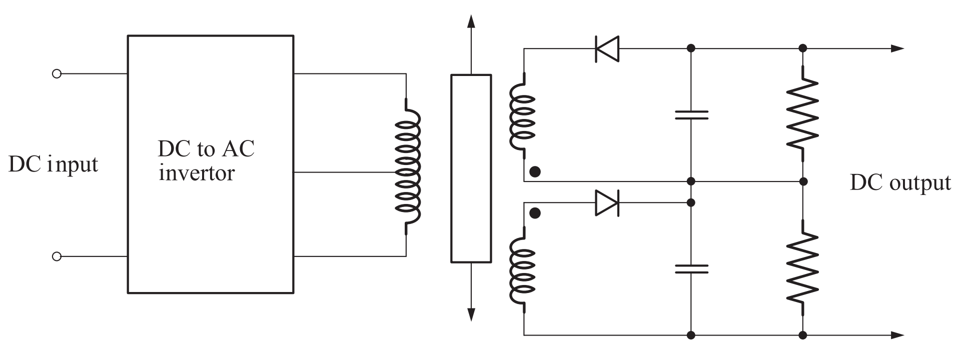


Figure 8.20 DC-LVDT circuit with signal conditioners

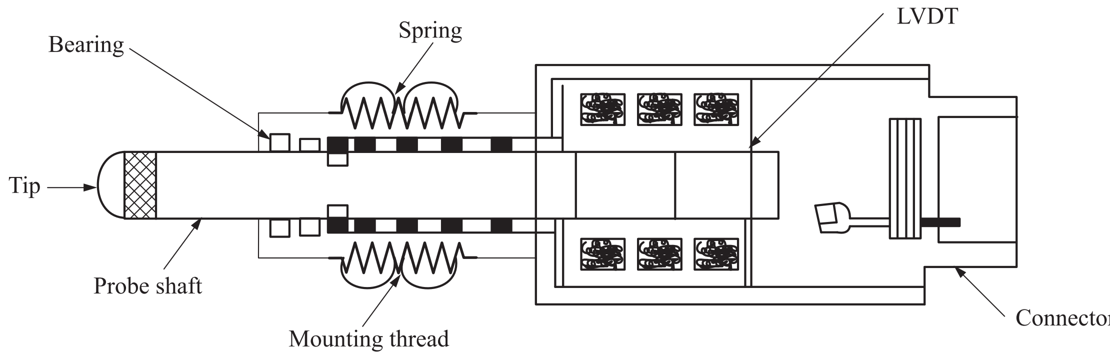


Figure 8.21 LVDT constructional details

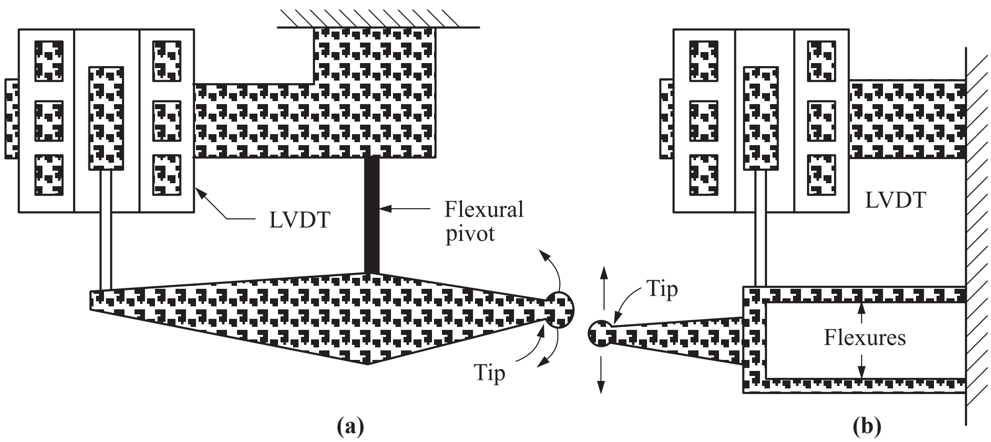


Figure 8.22 LVDT with lever or finger probe gauge heads

system. Another advantage of the LVDT gauging system is that a digital read-out can be incorporated, permitting unskilled or semi-skilled personnel to do production line inspection work. Ultra-precise LVDT gauging systems can make measurements that are otherwise difficult or impossible even with precision mechanical dial indicators or other mechanical measuring devices.

Non-contacting position measurement can be done using the pneumatic servofollower shown in Fig. 8.23.

The pneumatic servofollower consists of a double acting air cylinder having a piston, piston rod and seals, and a special nozzle at one end of the piston rod. A 2 kg/cm^2 pressure air is applied to the reference pressure chamber through the inlet orifice and then exhausted back to the atmosphere through the outlet orifice. Since both orifices are of the same size, the reference pressure developed equals one half of the supply pressure. The supply pressure is also connected to the nozzle where the baffle is the surface being measured. This gap-dependent pressure is fed back to the control pressure chamber. At a particular gap between the surface and the nozzle, typically 0.0075 cm, the control pressure equals one-half supply pressure and the piston does not move. If the gap changes because of dimensional or positional changes in the gauged surface, the control pressure changes accordingly, repositioning the piston and nozzle and restoring the gap. The LVDT coupled to the pneumatic servofollower gives an output either for the displacing or to the duplicating system.

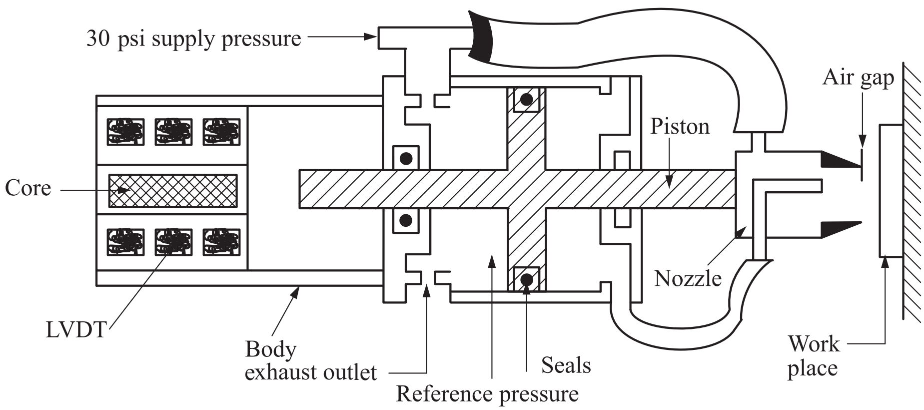


Figure 8.23 Pneumatic servofollowers LVDT

8.6.3 LVDT Load Cells

Many elastic members can be loaded in either tension or compression leading to deflection in either direction. The bidirectional nature of the displacement characteristic of LVDT perfectly complements the bidirectional deflection of an elastic member. Thus, LVDT load cells produce an output voltage linearly proportional to the axial force load. Typical examples of LVDTs combined with elastic members are illustrated in Fig. 8.24. LVDT pressure transducers require an elastically deformable sensing element that responds to omnidirectional fluid pressure. Figure 8.25 illustrates the mechanics of a typical C-tube LVDT pressure transducer. The coil is mounted to the block containing the fixed end of the tube and the pressure inlet. A cantilever spring and non-magnetic core rod maintain the core in the centre of the LVDT coil. As the deflection of the tube is small, a voltage output that is a linear function of pressure is obtained. Pressure transducers can also be constructed using helical or twisted Bourdon tubes connected to a RVDT.

For low-range pressure measurements, axially convoluted bellows, radially convoluted diaphragms, flat diaphragms, and welded pressure capsules are alternatives to Bourdon tubes as sensing elements in low-range pressure LVDT transducers. Figure 8.26 shows a capsule-type LVDT pressure transducer.

Thus, the LVDT has got a wide utility. Schaevitz Engineering has been in the forefront in the development of LVDT with its total system capability. New use is being discovered everyday for this device.

8.7 SYNCHROS

The term synchro means a family of AC electromechanical devices, which in various forms perform the functions of angle measurement. Voltage and/or angle addition and subtraction, remote angle transmission, and computation of rectangular components of vector synchros for angle measurement are most utilised as components of servomechanisms.

The other names for this device are Selsyn and Autosyn, which are acronyms for *self-synchronising* and *automatic synchronising*. Basically, synchros are mutual-inductance transducers. The synchro contains a rotor and a stator. The rotor carries a winding as shown in Fig. 8.27(a) and the connections to this winding are made through slip rings. The stator is wound exactly like an AC motor containing three windings. They are connected in a star shape as shown in Fig. 8.27(c). The rotor sets up a flux in the stator space. The flux induces voltage in the stator windings. The magnitude of the voltages induced in each stator winding

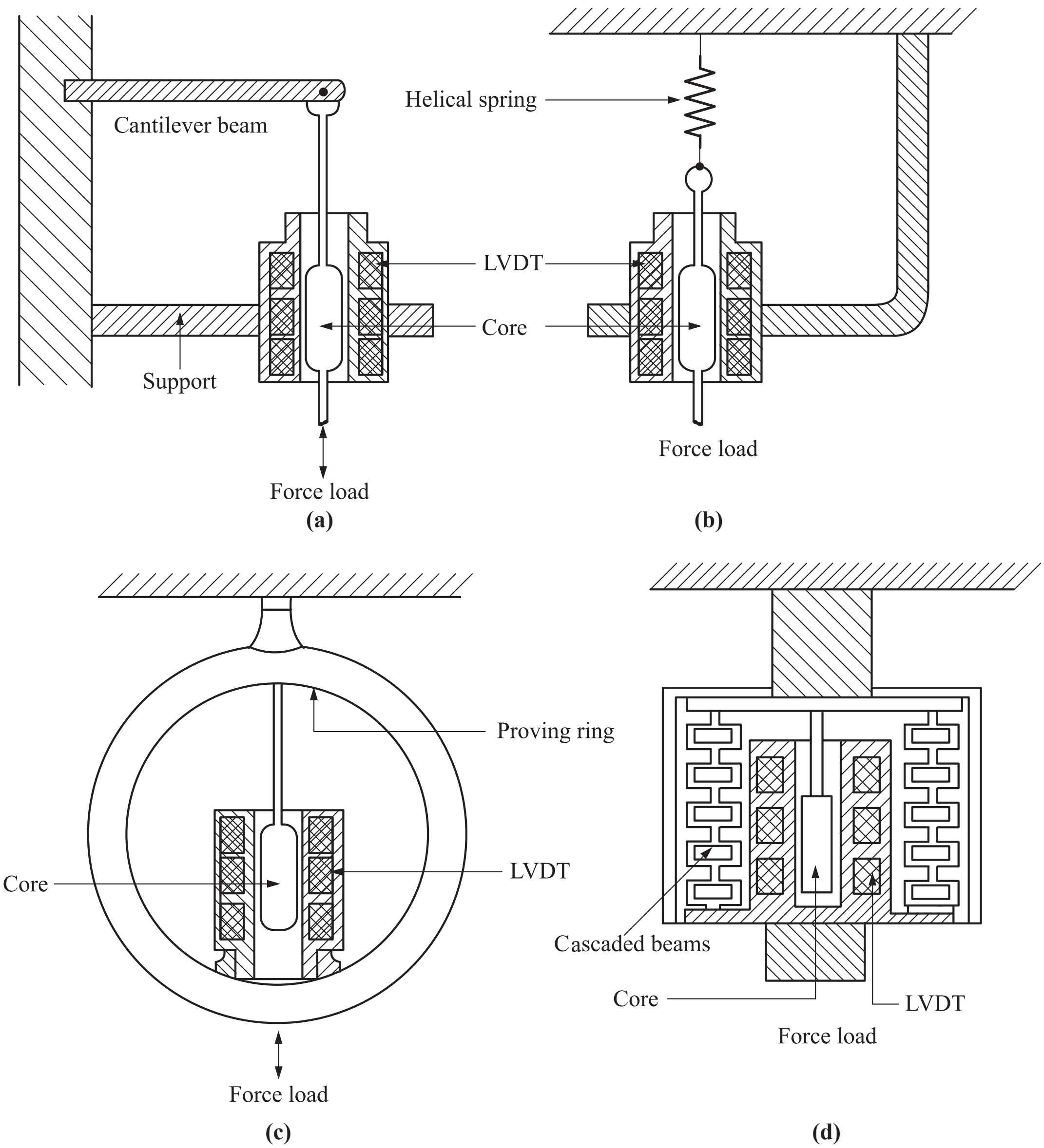


Figure 8.24 LVDT with elastic members

depends on the rotor position with respect to that winding. For the rotor position shown in Fig. 8.28, the voltages in the stator windings are given by the equations:

$$e_1 = E_{max} \sin \omega t \cos \theta \tag{8.1}$$

$$e_2 = E_{max} \sin \omega t \cos (120 - \theta) \tag{8.2}$$

$$e_3 = E_{max} \sin \omega t \cos (120 + \theta) \tag{8.3}$$

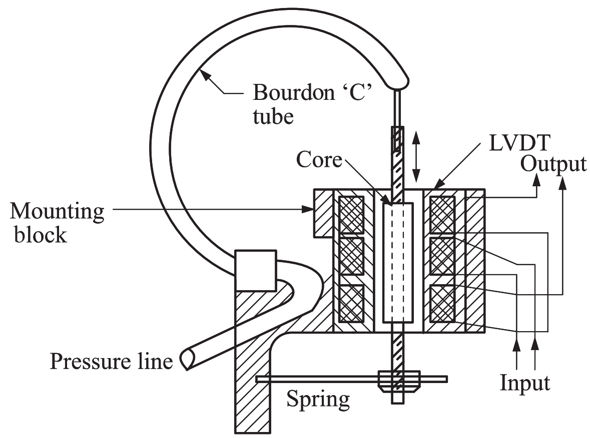


Figure 8.25 LVDT with C tube pressure transducer

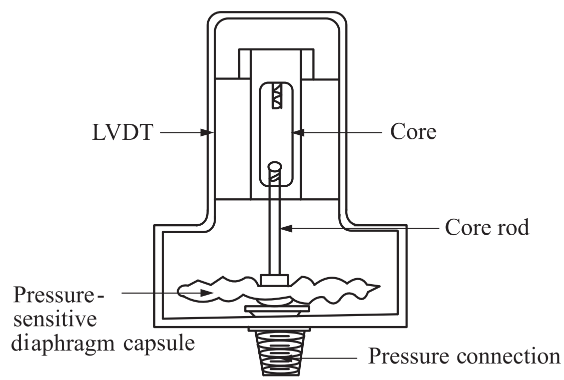


Figure 8.26 Capsule-type pressure transducer with LVDT

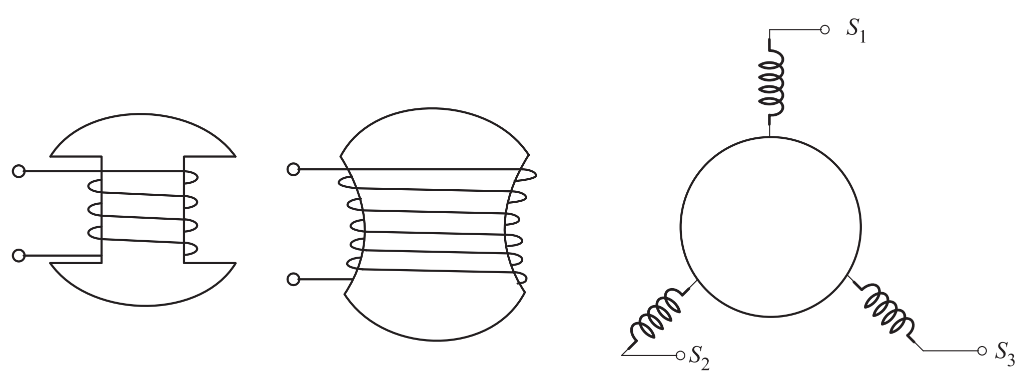


Figure 8.27 Types of synchro rotors (a) dumb-bell shaped (b) cylindrical and (c) circle

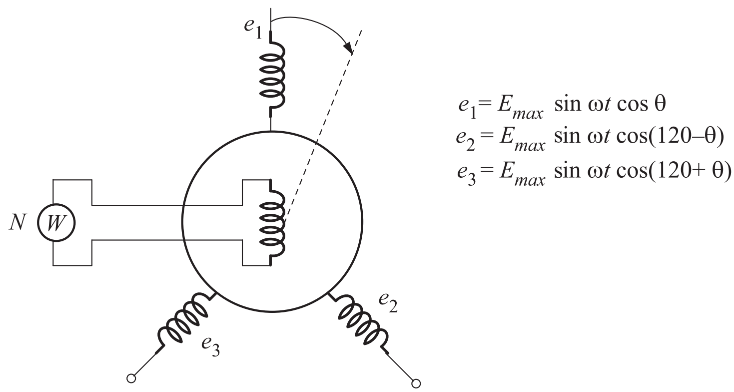


Figure 8.28 Rotor position of a synchro

The frequency of the stator winding voltage is the same as the frequency of the voltage applied to the rotor. $E_{\max}$ is the peak voltage induced in the coil. It cannot be that all the three voltages given by equations (8.1) – (8.3) are in the same phase. In addition, for a given set of stator voltages, there is only one corresponding rotor position. The frequency of supply to the rotor is either 50 or 400 Hz. The 400-Hz supply is used in aircraft systems for reducing the size of the synchros.

A pair of synchros can be used for transducing angular position to a voltage signal. Consider the system shown in Fig. 8.29.

The stator coils are connected as shown (S_1 to S_1, S_2 to S_2, and S_3 to S_3). The rotor of the synchro generator (*SG*) is kept at $\theta = 0$ position. This position is known as the *mechanical zero position*. This rotor is supplied with AC voltage. The rotor of the second unit, which is known as the control transformer (*CT*), is kept at 90° to the mechanical zero position of the rotor of the *SG*. This rotor is cylindrically shaped in construction to reduce the variation of reluctance with position (Fig. 8.27(b)). The rotor of the *SG* is normally dumb bell shaped as shown in Fig. 8.27(a). Now when $\theta = 0$ the flux is aligned with the winding of *SG* and induces a maximum voltage in winding *I*. In other windings also voltages are induced but of different magnitudes. These voltages circulate a current in windings S_1, S_2, and S_3 of the *CT* and produce a flux in the stator space of the *CT*. This flux pattern will be exactly the same in

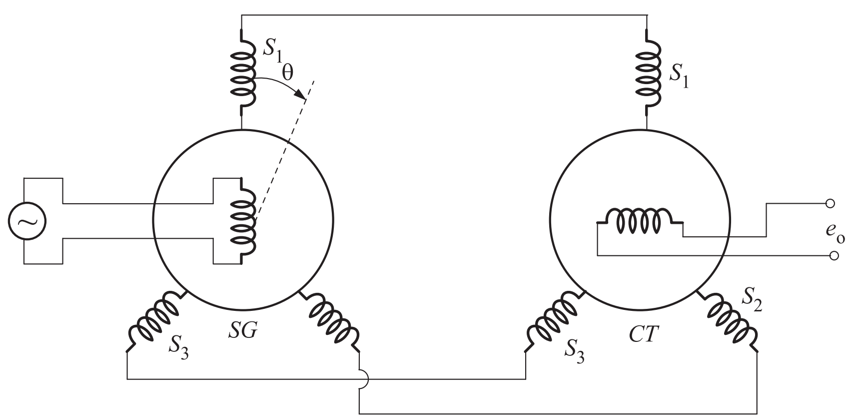


Figure 8.29 Transmission of an angular position using a synchro

every direction. This can be shown by considering current flowing through the stator windings of the *CT*. Therefore, the flux in the *CT* is aligned with winding *I* of the *CT* stator. As the rotor winding of the *CT* is 90° to the winding *I* the voltage in the *CT* rotor is zero. This position of the rotor is known as the *electrical zero position*. Thus, when θ is equal to 0°, e_o is also equal to zero. Now a 10° displacement is given to the rotor of the *SG*. The flux in the *CT* rotor space is also displaced by the same 10° from the winding *I*. The angle between this flux and the *CT* rotor winding is 80° and hence a voltage is induced in the winding. This voltage will be maximum if the angle of displacement of the *SG* rotor is 90°. Note that at any time the rotor of the *CT* is not moved at all. If a graph is plotted between the angular displacement and the output voltage, it will be θ sinusoidal variation as shown in Fig. 8.30. The negative half is purposely drawn for θ values ranging from 180° to 360° to stress the point that the phase angle between the supply voltage and the output voltage is 180°. This graph gives only the magnitude of the voltage. The frequency will be the same as that of the supply voltage for linear operation. The initial straight line portion of the graph is utilised. In this region the equation is $e_o = K_s\theta$, where K_s is the synchro constant. If the rotor of *SG* is given a sinusoidal motion with 10 oscillations per second, then the actual output will be a modulated wave as shown in Fig. 8.31.

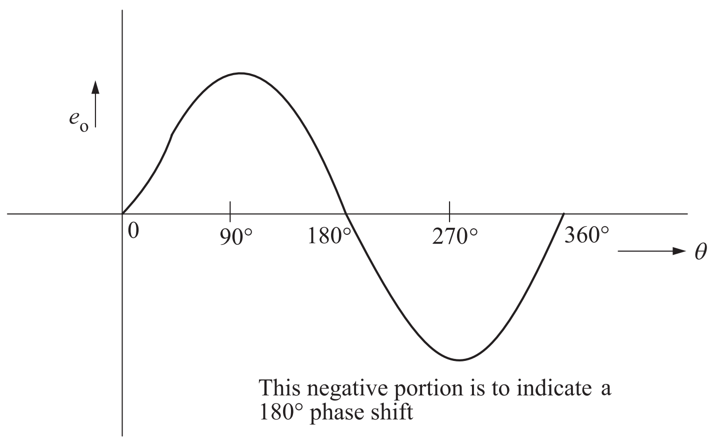


Figure 8.30 Output waveform with angular position

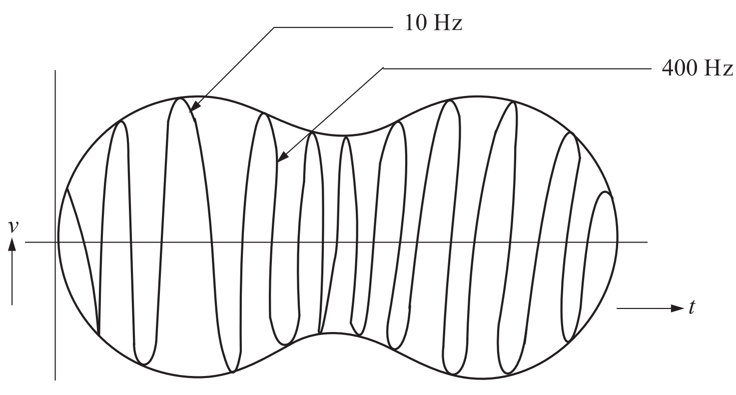


Figure 8.31 Modulated output waveforms

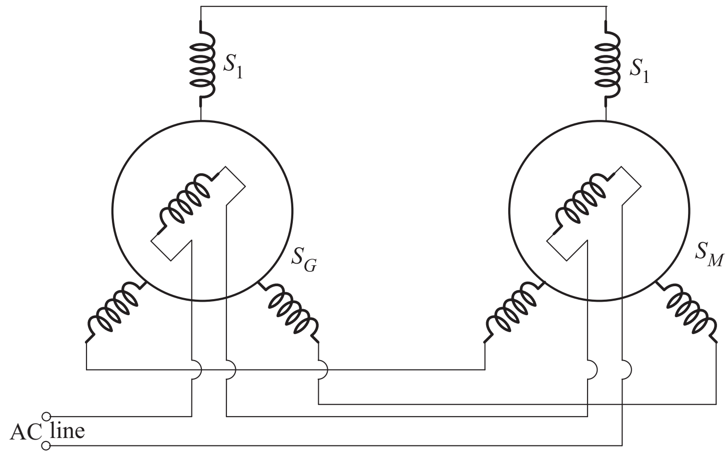


Figure 8.32 Synchros for telemetering angular position

This arrangement is used as an error-sensing device in position control systems. A pair of synchros can be used for telemetering the angular position to a remote location. The connections for this are shown in Fig. 8.32. The first unit is a SG and the second unit is known as the motor. The motor rotor is equipped with some additional damping units. The angular displacement of the motor follows that of the generator. The *SG* can be in one room and *SM* can be placed in a remote place. There will be five wires connecting them. There are other synchro units known as synchro differential units. These units have three windings in the rotor also. These are used to transmit an intelligence corresponding to the sum of two angular displacements over a distance.

In an *induction-potentiometer* there is one winding on the rotor and one on the stator (Fig. 8.33). The primary is excited with AC and voltage is induced into the secondary. The amplitude of this output voltage varies with the mutual inductance between the two coils and this varies with the angle of rotation. For a single-turn coil, the variation with the angle would be sinusoidal and only a small linear range around null would be obtained. By distributing other rotor and stator windings, a linear relation upto ±90° rotation may be obtained.

8.8 VARIABLE RELUCTANCE ACCELEROMETER

Figure 8.34 shows another common version of the variable reluctance principle. This is an accelerometer for the measurement of acceleration in the range $\pm 4g$, where 'g' is the acceleration due to gravity. Since the force required to accelerate a mass is proportional to the acceleration, the springs supporting the mass deflect in proportion to the acceleration. The mass is made of iron and thus serves both as an inertial element for transducing acceleration to force and also as a magnetic circuit element for transducing motion to reluctance. Normally, such an instrument would be constructed so that the iron core would be halfway between the two *E* frames when the acceleration was zero, thus giving a zero output voltage for zero acceleration. However, to detect motion on both sides of zero, a fairly involved phase-sensitive demodulator would be required. To eliminate the demodulator, the iron core and springs are adjusted so that the core is offset to one side by an amount equal to the spring deflection corresponding to $4g$ acceleration. Thus, with no acceleration applied, the output accelerates to 5.0 V

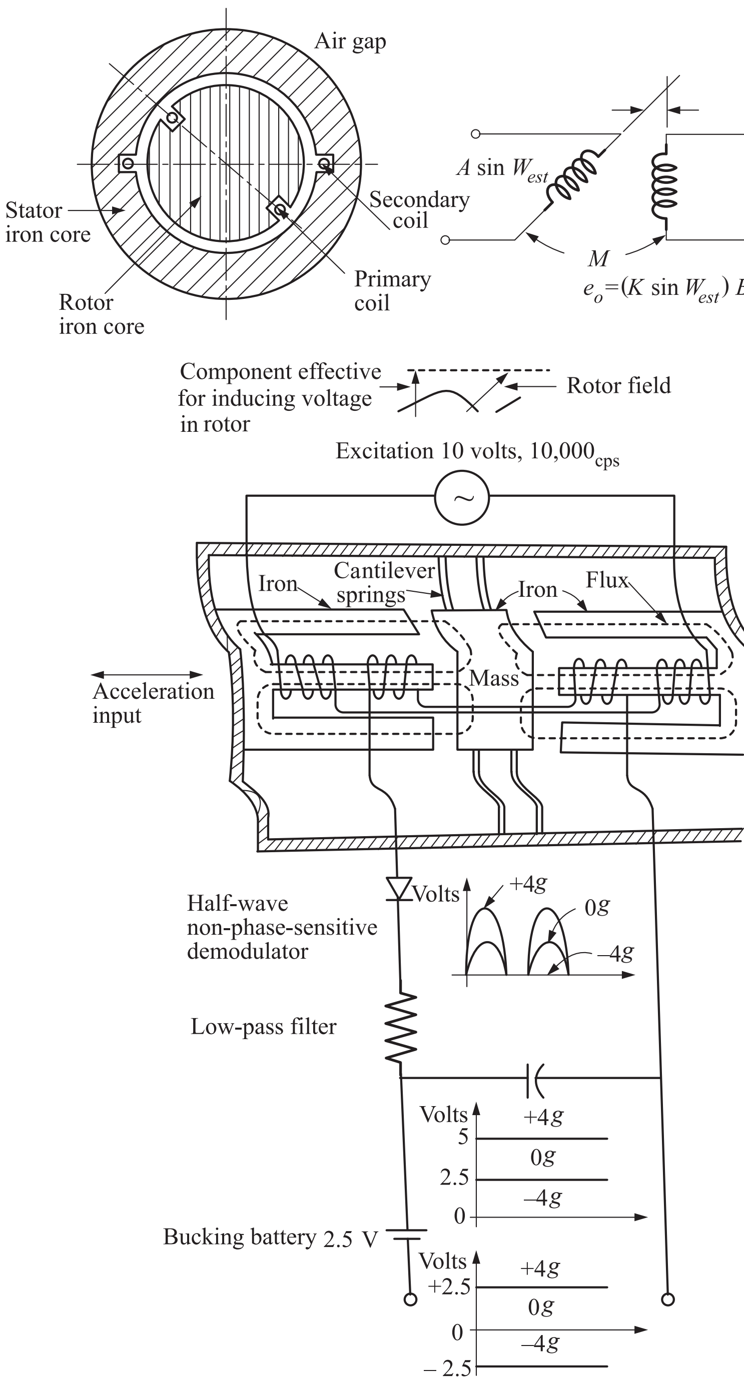


Figure 8.33 Induction potentiometer synchro

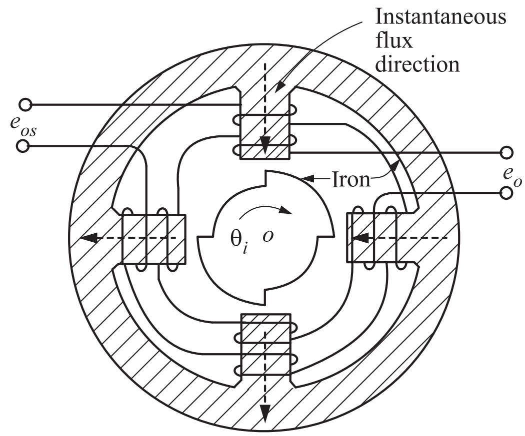


Figure 8.34 Variable reluctance accelerometer

and when –4*g* is applied the output will be zero. With the help of a bucking battery, a simple diode, and a filtering circuit, a phase-sensitive output proportional to the acceleration is obtained.

The actual full-scale motion of the mass in this particular instrument is just a few thousandth of a centimeter giving a displacement sensitivity for the variable reluctance element of almost 1000 V/cm.

8.8.1 Microsyn

Another variable reluctance element is the *Microsyn,* a rotary component shown in Fig. 8.34, which is widely used in sensitive gyroscopic instruments. The figure shows the instrument in the null position where the voltage induced in coils 1 and 3 is just balanced by those of coils 2 and 4. The null motion of the input shaft position increases the reluctance of coils 1 and 3 and decreases the reluctance of coils 2 and 4, thus giving a net output voltage e_o. Motion in the opposite direction causes a similar effect except that the output voltage has a 180° phase shift. If a direction-sensitive DC output is required, a phase-sensitive demodulator is necessary:

The excitation voltage is 5–50 V at 60–5000 Hz.
Sensitivity is of the order 0.2–5 V/degree.
Non-linearity is about 0.5% full scale for ±7° rotation and 1.0% for ±10°.

8.9 TEMPERATURE MEASUREMENT

Temperature measurement is among the most common and important measurements made in controlling industrial processes. The precise control of temperature is the key factor in many operations.

Temperature measurement can be made in many ways. They have been divided into two general classifications, those that are primarily electrical or electronic in nature and those that are not.

8.9.1 Electric Methods

1. Change in volume of a liquid when its temperature is changed.
2. Change in pressure of a gas when its temperature is changed.
3. Changes in vapour pressure when the temperature is changed.
4. Change in dimensions of a solid when its temperature is changed.

8.9.2 Electrical Methods

1. Electromotive forces generated by thermocouples.
2. Change in resistance of materials as their temperature is changed.
3. Temperature measurement by ascertaining the energy received by radiation.
4. Temperature measurement by comparing the colours of a controllable filament and the object whose temperature is sought. Figure 8.35 denotes the various methods.

It will be noted that several methods may be indicated for a given temperature span. However, all will not be equally well suited to any given temperature measurement. Therefore, the selection must be based not only on the range and span but also on factors such as life speed of response accuracy and means of mounting the sensing element. In some situations one may not be able to get the required temperature.

8.10 THERMOCOUPLES

For high and low-temperature measurements, thermocouples are most important. In 1821, Seebeck discovered that when two dissimilar metal wires are twisted together and heated, a sensitive meter connected to the other end of the pair will indicate a voltage (often called an electromotive force, *emf*) which is almost directly proportional to the difference in temperature between the heated or hot junction and the other end, which is called the cold junction. The theory of how exactly this *emf* is generated is not well understood even today.

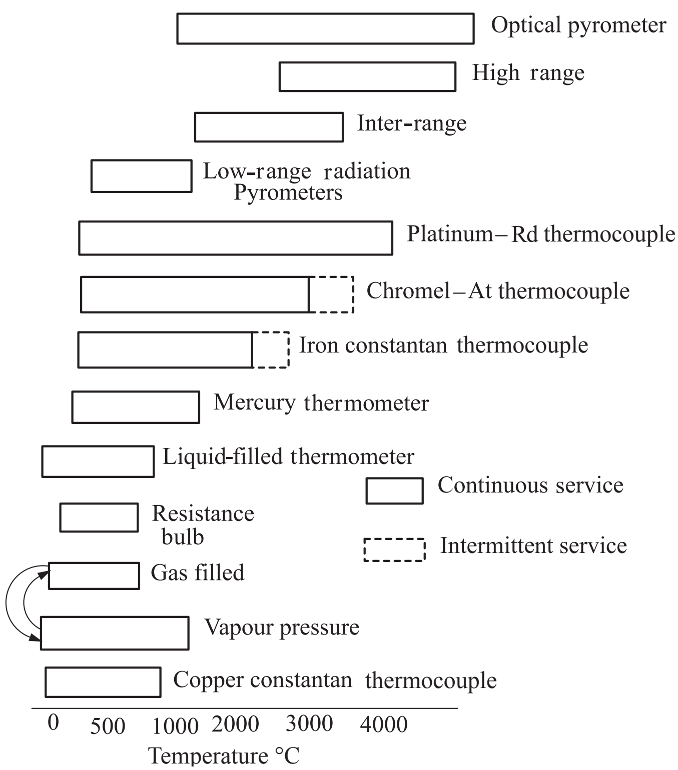


Figure 8.35 Temperature ranges of different types of thermocouples

For the analysis of most practical thermocouple circuits, the laws of thermocouple behaviour may be stated as follows:

1. The thermal *emf* of a thermocouple with junctions at T_1 and T_2 is totally unaffected by temperature elsewhere in the circuit if the two metals used are homogeneous (Fig. 8.36(a)).
2. If a third homogenous metal C is inserted into either A or B, as long as the two new thermo junctions are at like temperatures, the net *emf* of the circuit is unchanged, irrespective of the temperature of C away from the junctions (Fig. 8.36(b)).
3. If metal C is inserted between A and B at one of the junctions, the temperature of C at any point away from the AC and BC junctions is immaterial. So long as the junctions AC and BC are both temperature T_1, the net *emf* is the same as if C is not there (Fig. 8.36(c)).
4. If the thermal emf of metals A and C is E_{AC} and that of metals B and C is E_{CB}, the thermal *emf* of metals A and B is $E_{AC} + E_{CB}$ (Fig. 8.36(d)).
5. If a thermocouple produces *emf* E_1 when its junctions are at T_1 and T_2, and E_2 at T_2 and T_3, it will produce $(E_1 + E_2)$ when the junctions are at T_1 and T_3 (Fig. 8.36 (e)).

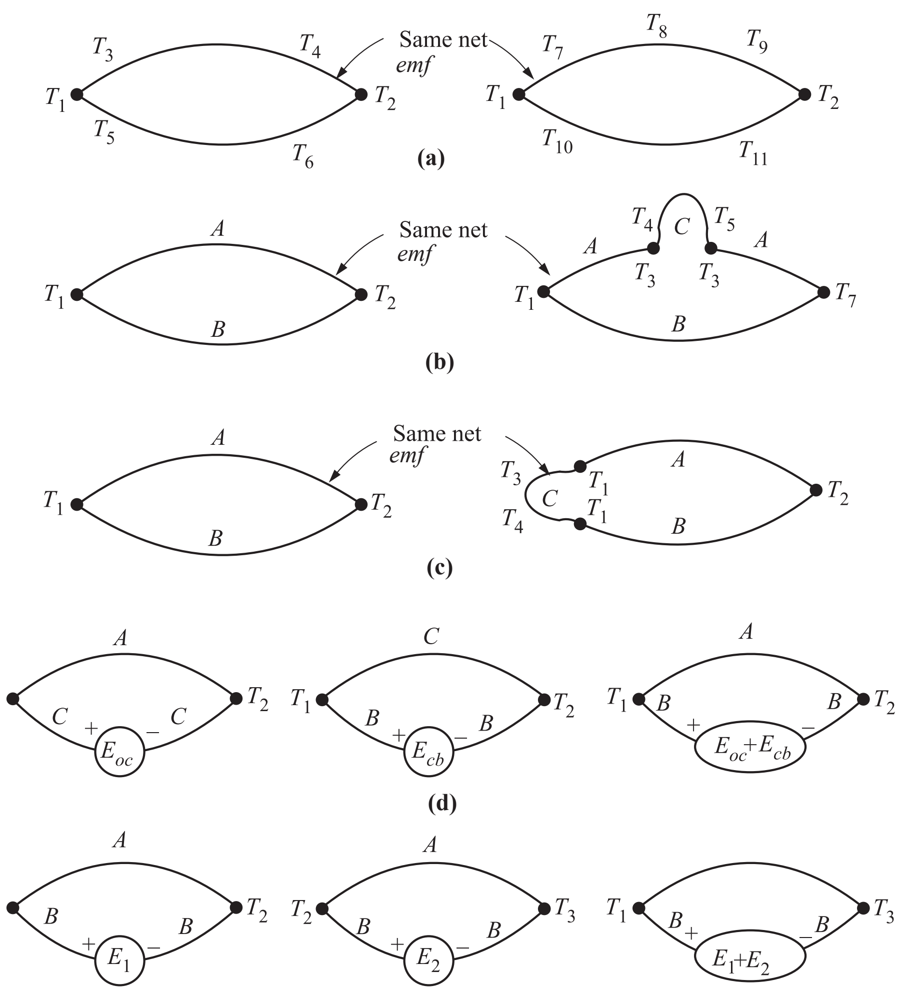


Figure 8.36 Laws of thermocouples—diagrams

8.10.1 Thermocouple Junction

The measuring junctions of a thermocouple may be formed by any method providing the necessary strength and electric contact. Wires may either be twisted together before soldering or welding to increase junction strength, or they may be welded. The cut wires will be usually curved from having been coiled and should be carefully straightened. Hammering, excessive twisting, and bending of the coil will change the wire, altering the thermal *emf* or may damage the surface and contribute to short life. Junctions of base metal thermocouples are often made with a solder having a melting point above any temperature at which the thermocouple is to be used. Acid fluxes should not be used because of their corrosive effect on the wires.

Welded junctions can be used at higher temperatures and are usually stronger. About one-half to one inch of each wire in a base metal junction should be cleaned with abrasive paper. The cleaned ends are then twisted together approximately one and half turns for gas and electric arc welding, or are brought into longitudinal contact for about a half-inch for electrical resistance welding. No surface preparation is required in this welding of noble metal wires, but extra care in handling is needed to avoid cold working and contamination by dirt, oil, perspiration, etc. Noble metals are best welded in a two-electrode arc, although gas welding is occasionally used.

Many thermocouple junctions are now welded with inert gas-shielded arc-welding equipment. This method should be used for easily oxidised materials like tungsten.

8.10.2 Thermocouple Insulation

There must be no electric connection between thermocouple wires except at the measuring instrument and at the measuring junction. For high-temperature service, ceramic insulators threaded over the wires are commonly used. Almost any ceramic material of suitable temperature rating may be used for base metal thermocouples. For thermocouples made of noble metals, low silicon insulators are used to prevent silicon poisoning of platinum wires in reducing atmospheres. Where temperature and other service conditions permit, insulated thermcouple wire often replaces ceramic-insulated construction especially if wires are too long.

8.10.3 Soldered, Drawn, or Rolled Sheathed Thermocouple Construction

These thermocouples are similar to electric heating elements in construction. Thermocouple wires in compacted magnesia, beryllia, or alumina insulation are surrounded by an integral metallic sheath. Junctions may be of the forms shown in Fig. 8.37. Sheathed construction available in 1.25 cm. and smaller diameters with one or more pairs of thermocouple wires provides exceptional physical and chemical protection to the wires, and the ability to bend the thermocouple in complex forms. It also provides insulating ability to weld or solder the sheath to the supporting structure or the surface to be measured and in exposed junction designs. Such a construction also gives a high speed of response.

8.10.4 Disposable-Tip Thermocouples

These are marketed for spot checking temperature of molten metal baths especially steel in open hearth and oxygen furnaces. Details are shown in Fig. 8.38. The tip fits on the end of a lance that has a connector, which fits well with the contact wire or ribbons of the thermocouple tip. Usually the cardboard-enclosed tip fits into a longer cardboard tube, which protects the lower part of the lance.

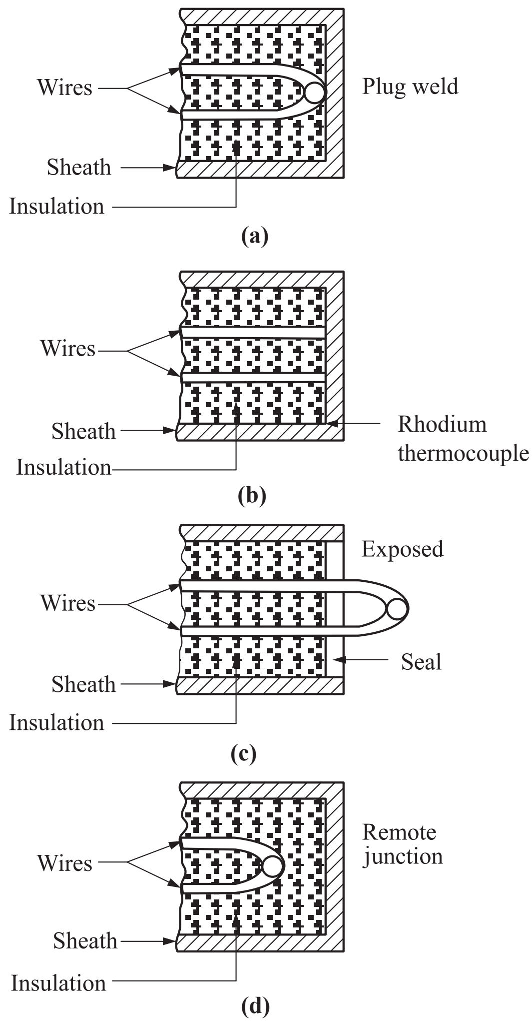


Figure 8.37 Thermocouple junctions—construction details

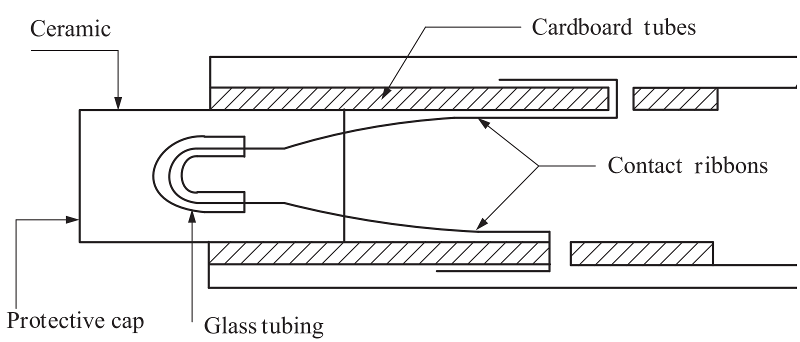


Figure 8.38 Disposable-tip thermocouples

The protective cap, usually metal, protects the thermocouple and glass tubing until the assembly has passed through the slag. It melts and shortly thereafter, the glass tubing and the thermocouple reach the temperature of the molten metal.

8.10.5 Homogeneity of Thermocouple Wires

Two additional factors involved in the use of thermocouples are the homogeneity of the wire and the annealing treatment given to it. Homogeneities may be either chemical or mechanical. Annealing relieves mechanical strains. The circuit shown in Fig. 8.39 may be used to check inhomogeneity. Move the dry ice slowly from one location to the next in order to establish a temperature difference equal to approximately 100 °C over a short length of the wire. A and A′, are assumed to be at room temperature; so there are two local temperature gradients. In accordance with the law of homogeneous circuits, any *emf* detected will be an indication of inhomogeneity in the wire in these regions. The annealing of the wire is normally done by the manufacturer.

8.10.6 Installation of Thermocouples

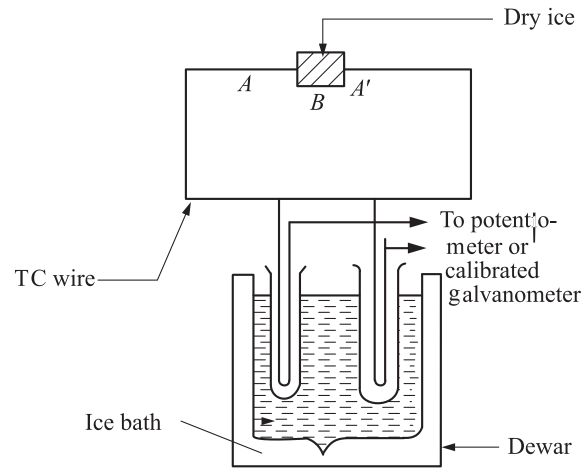
Locate the thermocouple so that the junction is immersed sufficiently far into the region in which the temperature measurement is desired. The temperature of a measuring junction may usually be brought to within a few hundredths of a degree of that of a liquid by immersing the bore wires into the liquid for a distance of 10 wire diameters. Immersion of 10 diameters of a protecting tube would be required for approximately the same accuracy. In most cases the measuring junction should be inserted as far as practicable into the heated medium.

Good thermal contact between the thermocouple wires adjacent to the junction and a solid whose temperature is to be measured will often reduce conduction errors. When measuring the hotness of gases at high temperatures, the cooling of the junction by radiation to the surrounding cool walls should be avoided by providing a radiation shield around the thermocouple. Thermocouple characteristics are shown in Fig. 8.39(b).

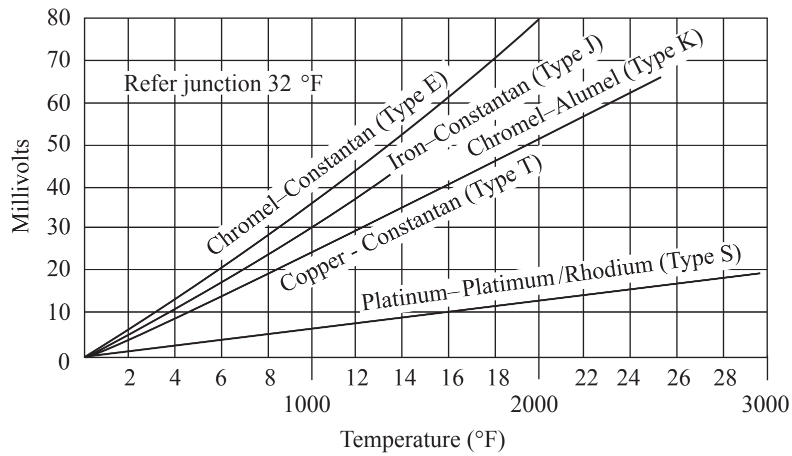
8.10.7 Cold Junction Compensation

The *emf* developed by a thermocouple depends on the temperature of both the measuring and reference junctions. The reference tables are usually based on a reference junction temperature of 0°C (32°F). In any application, it is not practical to maintain the reference junction at the ice point. The observed *emf* is added to the *emf* that the thermocouple would develop, if the reference junctions are at 0°C and the measuring junctions are at the actual temperature of the reference junction. The temperature of the reference junction can be maintained at a value higher than room temperature by a small temperature-controlled oven. A constant correction to the measurement gives the true temperature.

In other instruments, the temperature of the reference junction is at the ambient temperature, with correction being provided by temperature-sensitive devices. Some galvanometer-actuated instruments use a bimetallic spiral attached to hair spring assembly, which changes the pointer position as ambient temperature changes, so that the true temperature is read in the scale. The reference junction is located so that the reference junction and the spiral are at the same temperature. In most self-balancing potentiometers, a temperature-sensitive resistor automatically corrects the instrument calibration as ambient temperature changes. The reference junctions are located so that they are at the same temperature as the resistor.



(a)



(b)

Figure 8.39 (a) Circuit to check homogeneities in thermocouple tips and (b) thermocouple output voltage as a function of temperature for various thermocouple materials

8.10.8 Resistance Thermometry

Resistance thermometry utilises the characteristic relationship of electrical resistance with temperature to measure temperature. For pure metals, this relationship may be expressed by

$$R_t = R_o\,(1 + aT + bT^2 + cT^3 + \cdots)$$

where R_o is the resistance in Ω at reference temperature (usually at ice pt 0°C), R_t the resistance at temperature T, Ω, a the temperature coefficient of resistance Ω/Ω°C, and b and c the coefficients calculated on the basis of two or more known resistance–temperature (calibration) points.

Most elements constructed from metal conductors generally display positive temperature coefficients with increase in temperature resulting in increased resistance, whereas most semiconductors display a characteristic negative temperature coefficient of resistance.

8.11 PLATINUM RESISTANCE THERMOMETERS

Only a few of the pure metals have a characteristic relationship suitable for the fabrication of sensing elements used in resistance thermometers. Of all the materials, platinum has the optimum characteristics for service over a wide range of temperatures. Platinum with a temperature coefficient of resistance equal to 0. 00385 Ω/Ω°C (from 0 to 100°C) is being used as a standard for industrial thermometers. It has a high melting point, and does not volatilise appreciably at temperatures below 1200°C.

The resistance–temperature relationship for platinum element is given by the equation

$$T = \frac{100\left(R_t - R_o\right)}{R_{100} - R_o} + \delta\left(\frac{T}{100} - 1\right)\frac{T}{100}$$

where T is the temperature, °C, R_t the resistance at temperature t, Ω, R_o the resistance at 0°C, Ω, R_{100} the resistance at 100°C, Ω, and δ the calendar constant (approximately 1.50).

The thermometer's actual performance depends on the function of its critical design features. Since the sensitive winding of a resistance element usually comprises a small diameter wire, which must be protected from the medium being measured, generally it is mounted in a metal protective housing in such a manner that the thermal resistance path from the outer surface of the housing to the wire is held to a minimum. It is imperative that the resistance winding be well insulated electrically from its housing since shunt resistance will affect the thermometer's performance.

Modern application requirements for high sensitivity, fast response, and extreme ruggedness at a relatively low cost have led to the most versatile, reliable, and widely used basic platinum resistance thermometer design—the fully encapsulated Resistance Thermometer Device (Fig. 8.40). The platinum wire, usually of 0.00254 cm outer diameter or less, is wound into a coil and inserted into a multibore high-purity ceramic tube. The most commonly used ceramic material is aluminum oxide. The winding is completely embedded and fused within the ceramic tube utilising extremely fine-grained powder. The intimate contact permits a rapid speed of response. Advanced manufacturing techniques can produce platinum elements with a base resistance of 100 Ω and physical dimensions as small as 0.2 cm in outer dia and 2 cm long. Internal lead wires, usually a base metal alloy with high temperature capabilities, such as inconel are welded to the sensitive elements—noble metal lead wires in close proximity to the element with three or four lead circuits the most commonly utilised. A multibore ceramic tube serves to insulate the internal lead wires throughout the entire length of the assembly. The thermometer's external extension lead wires, usually teflon-insulated copper, are attached to the internal lead wires just above the ceramic insulators. The entire internal assembly is inserted into a stainless steel or inconel sheath, which has been treated to remove all forms of contamination. High purity, fine-grained ceramic powder is then packed into the assembly using vibration techniques after which the entire unit is hermitically sealed at the lead end with high-temperature epoxy materials.

For industrial temperature measurements in the range –70 to ±150°C, resistance thermometers with nickel-sensing element, and for range –200 to +150°C, copper resistance thermometers are being successfully used.

8.12 SPECIAL RESISTANCE THERMOMETER

Many special resistance thermometers have utilised the basic wafer design, having the characteristics of small mass combined with good thermal contact, resulting in an extremely fast time response. A fine insulated wire of copper, nickel, or platinum is usually sandwiched between two protecting sheets of insulating material and sealed as shown in Fig. 8.41.

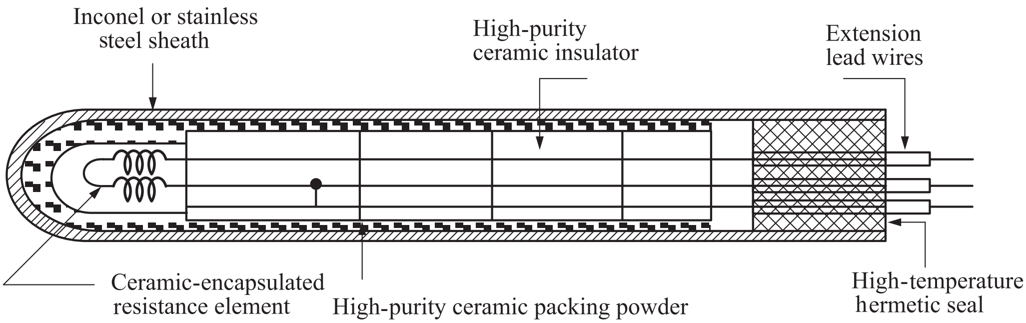


Figure 8.40 Thermocouple construction

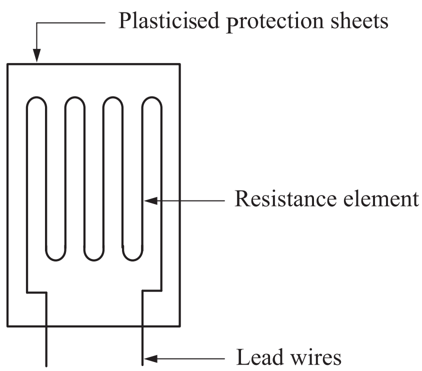


Figure 8.41 Special-type resistance thermometer

8.12.1 Performance and Testing Procedure

In addition to the temperature range, temperature coefficient, and interchangeability, the following performance characteristics are of primary concern:

1. *Accuracy:* The accuracy of calibration is defined as the ability of a thermometer to conform to its predetermined resistance–temperature relationship and is commonly expressed in terms of the percent of actual temperature reading. The testing is done either by comparison with a standard thermometer or by a fixed point technique by measuring known temperatures like freezing of pure metals.
2. *Stability:* It is the ability of the thermometer to maintain and reproduce its specified resistance–temperature characteristics for long periods of time with its specified temperature range of operation. The degree of stability is expressed as *drift*. The stability test method requires that the thermometer maintain suitability under cyclic test conditions within a usable temperature range.
3. *Time response:* Time response is that time required for the thermometer to react to a step change in temperature and reach the resistance corresponding to 63. 2 % (one time constant) of the total temperature change. In order to assume a repeatable step change in temperature,

test methods call for a reference point, usually ice point and an elevated fixed point, that is, constant temperature-stirred water bath. The test thermometer should be properly connected to a high-speed recorder.

4. *Self-heating:* The flow of current through the element produces heat, which can be a source of error. The well-designed resistance thermometer will properly dissipate self developed heat into the measured medium resulting in extremely small errors. Standard test methods to determine self-heating effects require the test thermometer to be immersed in a uniform heat transfer medium and supply current in increments of 1–2 mA within the specified range of excitation. A resistance versus current curve is drawn and extrapolated for values less than 1 mA.

8.12.2 Insulation Resistance

Of all the factors contributing to insulation breakdown, moisture effects are most common, particularly in fully encapsulated designs where ceramic packing powder readily absorbs moisture. Therefore, insulation resistance is a function of temperature and calls for tests to be carried out at different temperatures within the operating range.

8.12.3 Vibration Resistance

Test apparatus includes a vibration shaker capable of producing the sinusoidal vibration level over the specified frequency range while the test thermometer is mounted on the shaker head.

8.13 THERMISTORS

Thermistors are included in the class of solids known as semiconductors, having electronic conductivities between those of conductors and insulators. The name thermistor is derived from thermally sensitive resistors, since the resistance of a thermistor varies as a function of temperature.

A thermistor is an electrical device made of a solid semiconductor with a high temperature coefficient of resistivity, which would exhibit a linear voltage current characteristic if its temperature is held constant. When a thermistor is used as a temperature-sensing element, the relationship between resistance and temperature is of primary concern. The approximate relationship that applies to most thermistors is

$$R_T = R_o \exp \beta \left(\frac{1}{T} - \frac{1}{T_o} \right) \tag{8.4}$$

where R_o is the resistance value at reference temperature T °K in Ω, R_T the resistance at temperature T °K, Ω, and β the constant over temperature range, dependent on manufacturing process and construction characteristics (specified by the supplier):

$$\beta = \frac{E}{K} \tag{8.5}$$

where E is the electrovolt energy level in eV and K Boltzmann's constant (8.625×10^{-5} eV/K).

A second form of approximate resistance–temperature relationship is expressed in the form

$$R_T = R_\infty e^{\beta/T} \tag{8.6}$$

where R is the thermistor resistance in Ω as temperature approaches infinity.

Both equations (8.4) and (8.5) are only best approximations and, therefore, are of limited use in making highly accurate temperature measurements. However, they do serve to compare thermistor characteristics and thermistor types.

The temperature coefficient is usually expressed as a percent change in resistance per degree of temperature change and is approximately related to β by the equation

$$a = \frac{dR}{dT}\left(\frac{1}{R}\right) = \frac{-\beta}{T_0^2} \tag{8.7}$$

where T_o is in °K. It should be noted that the resistance of the thermometer is solely a function of its absolute temperature. Furthermore, it is apparent that the thermistor's resistance–temperature function has a characteristic high negative coefficient as well as a high degree of non-linearity. The value of the coefficient for a common commercial thermistor is in the order 2 – 6 % per °K at room temperature. This value is approximately 10 times that of metals used in the manufacture of resistance thermometers.

Resultant considerations due to the high coefficient characteristic of thermistors include inherent high sensitivity and a high level of output, eliminating the need for extremely sensitive read-out devices and the lead wire matching techniques, respectively. However, limitations on interchangeability (particularly over wide temperature ranges), calibration, and stability—also inherent to thermistors—are quite restrictive. The high degree of non-linearity in the resistance–temperature function usually limits the range of read-out instrumentation. In many applications, special prelinearisation circuits must be used before interfacing with related system instrumentation. The negative temperature coefficient may also require an inversion (to positive form) when interfacing with some analog and/or digital instrumentation.

8.13.1 Thermistor Construction Techniques

A number of metal oxides and their mixtures, including the oxides of cobalt, copper, iron, magnesium, manganese, nickel, tin, titanium, uranium, and zinc are among the most common semiconducting materials used in the construction of thermistors. Usually compressed into the desired shape from the specially formulated powder, the oxides are then recrystallised by heat treatment, resulting in a dense ceramic body. The lead wires are then attached while electric contact is maintained, and the finished thermistor is then encapsulated.

There are many mechanical configurations for the thermistors as a primary temperature element. Beads are made (Fig. 8.42(a)) by forming small ellipsoids of material suspended on two fine lead wires approximately 0.10 in apart. The material is sintered at elevated temperature and the lead wires become tightly embedded within the bead, making electric contact with the thermistor material. For more rugged applications, the bead thermistor is sealed into the tip of a glass, ceramic, or suitable metal sheath.

Disc thermistor configurations (Fig. 8.42(b)) are manufactured by pressing the semiconductor material into a round dye to produce a flat circular probe. These pieces are sintered and then silvered on the two flat surfaces. Thermistor discs range from 0.25 cm to 2.5 cm diameter and 0.05 cm to 1.25 cm in thickness. Disc thermistors are commonly applied where a moderate degree of power dissipation is required.

Washer-type thermistors (Fig. 8.42(c)) are manufactured like disc thermistors, except that a hole is formed in the centre of the sensor to provide for bolt mounting. Normal washer configurations are approximately 1.9 cm in diameter and are applied where higher power dissipation is a primary requirement.

Rod-type thermistors (Fig. 8.42(d)) are extruded through dies, resulting in long cylindrical probes commonly varying from 0.0125 cm to 0.265 cm in diameter and from 0.625 cm to 5 cm length, with lead wires usually attached at the ends of the rod. Rod configurations are generally of high terminal resistances and are generally applied where power dissipation is not a principal concern.

8.13.2 Thermistor Performance Characteristics

Evaluation of thermistor performance characteristics is in many cases similar to that of resistance thermometers.

Figure 8.43 demonstrates the logarithm of the specific resistance versus temperature relationship for three typical thermistor materials as compared to platinum metal. The specific resistance of the thermistor represented by the curve decreases by a factor of 50 as the temperature is increased from 0 to 100°C. Over the same temperature range, the resistivity of platinum will increase by an approximate factor of 1.39.

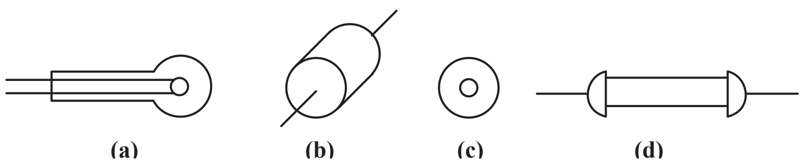


Figure 8.42 Different types of thermistors

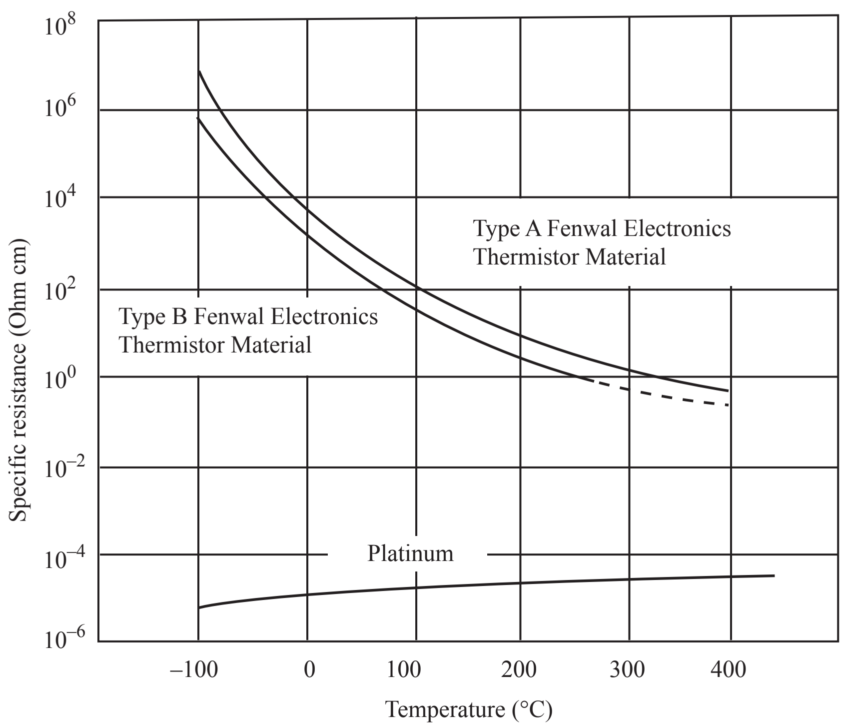


Figure 8.43 Thermistor characteristics

In thermistors, terminal resistance at room temperature range from about 1 Ω to the order of 10 Ω depending upon the composition, the shape, and the size. For a given type, they commonly vary from 10 to 30% in resistance from nominal value at the reference temperature. Some types may be specially manufactured and selected to bring resistance values within closer limits.

The range in power-dissipation factor varies from approximately 10^{-5} W to several watts per degree Celsius of resultant rise in temperature. The value varies inversely with the degree of thermal isolation of the thermistor element.

The time constant varies from few tenths of a second to several minutes among general-purpose thermistors. Naturally, the shorter times are associated with the smaller bead configurations, and the longer times with the longer rods and discs as well as larger beads. The time constant varies directly with the thermal capacity of the thermistor and inversely with the dissipation factor. The stability with time of the resistance-temperature function depends upon thermistor construction as well as the applications. Slight changes may occur in the resistivity of the semiconductor or in the contact medium and its relation with the semiconductor. Resistivity may change through chemical changes in composition, caused by decomposition or diffusion processes, which are generally accelerated at higher temperatures. Certain oxide compositions are inherently more stable than others. Thermistors enclosed in glass coverings with sintered-in noble metal lead wires usually exhibit relatively good stability with time. If subjected to cycling at relatively high temperatures, some types of thermistors may exhibit a behaviour characteristic known as *relaxation*. Resulting resistance changes may be as much as several percent following the cyclic temperature change, with a recovery time of few hours. When thermistors are held at a fixed temperature with negligible current flow, resistance changes in the order of 1% per year (Fig. 8.44).

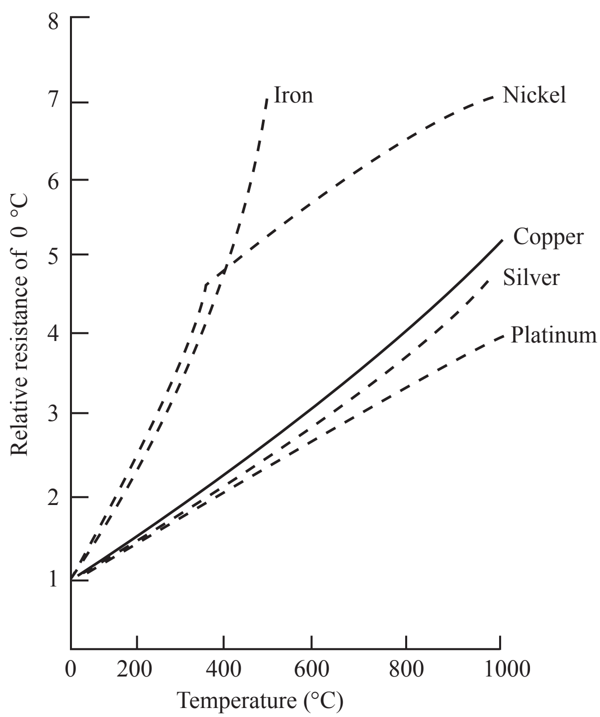


Figure 8.44 Relative resistance (R_t/R_{ref}) versus temperature graph for some pure metals

The upper operating temperature limit of thermistors is set by physical changes of the semiconductor material or by calibration considerations. This limit will depend upon the construction of the thermistor and the degree of precision required for a given application. Typical temperature ranges are in the order of 100–400 °C. The low temperature operating limit is usually determined by the resistance, reaching such a high value so as to make convenient measurements difficult. For this application, only thermistors of relatively low coefficient and very low room temperature resistance are practical.

8.13.3 Applications of Thermistor

The application of thermistors as primary temperature elements follows the usual principle of resistance thermometry. Conventional bridge or other resistance-measuring circuits as well as constant current circuits are employed. Special application considerations must be given to the negative and highly non-linear resistance-temperature relationship, as previously noted.

8.13.4 Sensistors

Sensistors are heavily doped semiconductors, having almost the same properties as that of metals. They have a positive temperature coefficient of resistance. The materials used are oxides of Manganese and Cobalt.

8.14 DIGITAL TEMPERATURE-SENSING SYSTEM

The analog outputs of the various temperature sensors can be transformed to a digital form by using any of the available analog to digital conversion techniques.

Electronic oscillators using piezoelectric quartz crystal as the resonant element that establishes the frequency of oscillation have been widely used for many years. For the most critical application, it has been necessary to place the crystal in a temperature-controlled oven, since the natural frequency of the crystal varies with temperature, causing drifts in oscillatory frequency. This difficulty is turned to good advantage in the quartz thermometer where the crystal is placed in a probe, which serves as a temperature-sensing element. Changes in probe temperature cause a frequency change in proportion. By applying the oscillator voltage to an electronic counter for a definite time interval, a direct digital reading of temperature is obtained.

8.15 MISCELLANEOUS TRANSDUCERS

8.15.1 Flow Measurements

Principle: Movement of the fluid stream that flows through the primary element of the rate meter is directly or indirectly used to actuate a secondary device, and the rate of flow is inferred from known physical laws or from empirical relation.

Head meters are those that operate by the measurement of the pressure differential or head across a suitable restriction to flow in the pipe line.

Primary element: Some form a restriction in the flow line to induce the head.

Secondary element: Connected to the differential head and measures it as a means of determining the flow rate.

Principle of Head Flow Meters: It is assumed that the fluid is flowing though an inclined pipe line as shown in Fig. 8.45. The relation between pressure differential (head) and velocity can be derived from Bernoulli's equation. For ideal incompressible fluids, this takes the form

$$Z_1 + \frac{p_1}{\gamma} + \frac{V_1^2}{2g} = Z_2 + \frac{p_2}{\gamma} + \frac{V_2^2}{2g}$$

where

Z = elevation of the centre line,
p = static pressure, absolute,
γ = specific weight,
V = average stream velocity,
g = acceleration due to gravity.

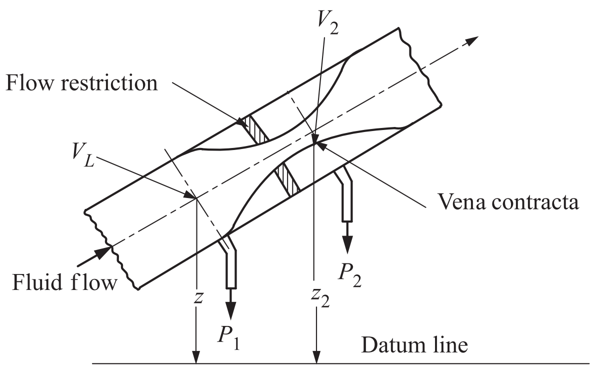


Figure 8.45 Flow meter

For the upstream, suffix 1 is used and for the downstream, suffix 2 is used. If the pipe is horizontal, then

$$\frac{V_2^2 - V_1^2}{2g} = \frac{p_1 - p_2}{\gamma} = \text{ln} = \text{effective head}$$

since $Q = aV_2 = AV_1$

$$V_1 = V_2 \frac{a}{\text{A}}$$

where a is the area of restriction and A the area of the pipe.

$$V_2^2 = \frac{2gh}{1 - a^2/\text{A}^2} = \frac{2\,gh}{1 - \left(\frac{d}{D}\right)^4}$$

$$V_2 = \sqrt{\frac{2gh}{1-(d/D)^2}}$$

$$q_t = \text{theoretical volume of flow}$$

$$= V_2 a = a\sqrt{\frac{2gh}{1-\beta^4}}$$

where

$$\beta = \text{Ratio of diameters} = \frac{d}{D}$$

In terms of weight,

$$W_t = q_t = a\gamma\sqrt{\frac{2gh}{1-\beta^4}}$$

In order to correct the equations for q_t and W_t from theoretical to actual flow rates based upon experimental data, a correction factor is incorporated in the equation as

$$C = \frac{\text{actual rate of flow}}{\text{theoretical rate of flow}} = \frac{q}{q_t} = \frac{W}{W_t}$$

Furthermore, let

$$K = \frac{C}{\sqrt{1-\beta^4}}$$

where K is the discharge coefficient. Including velocity of approach, the term $\frac{1}{\sqrt{1-\beta^4}}$ is commonly known as the approach factor

$$q = K_a\sqrt{2gh}$$

$$W = K_a\gamma\sqrt{2gh}$$

Flow-measuring devices generally fall into one of the categories, namely primary devices or quantity meters and secondary devices known as rate meters. The distinction between the two is based on the character of the sensing element that interacts with the fluid flow. Quantity measurements, by mass or volume, are usually accomplished by counting successive isolated portions, whereas rate measurement is inferred from effects of flow rates on pressure, force, heat transfer, flow area, etc. Quantity meters are generally used for the calibration of rate meters. All these devices are basically of the mechanical type.

However, there are certain flow-measuring devices that cannot be strictly categorised as fluid mechanical types of primary flow-measuring devices or the indirect type of rate flow meters. In these devices, the principle of operation is generally non-fluid mechanical in nature. For example, the ultrasonic time delay technique, the electromagnetic effects of fluid flow, and heat transfer from heated wires, Doppler frequency shift of scattered light, etc. are some of the principles used in the measurement of fluid flow parameters.

8.15.2 Ultrasonic Flow Meters

The principle of this type of instrument is based on the apparent change in the velocity of propagation of sound pressure pulses in a fluid with a change in velocity of the fluid flow. In practice, we employ short bursts of sinusoidal pressure tubes whose frequency is above the audio frequency range (i.e., above 20 kHz). A typical frequency may be of the order of 10 MHz.

$$t_o = \frac{l}{V_s}$$

where l is the distance between transmitter and receiver and V_s the velocity of sound in fluid

Now in a fluid moving with a velocity 'V', the transit time t becomes

$$t = \frac{Q}{V_s + V} \frac{l(V_s - V)}{V_s^2 - V^2}$$

For flow velocities of liquids encountered, in practice, the error involved would be negligible if we assume $V << V_s$. Therefore, the above equation becomes

$$\text{t} \approx \frac{l}{V_s}\left[1 - \frac{V}{V_s}\right] = t_o\left[1 - \frac{V}{V_s}\right] \tag{8.8}$$

If we define $\Delta t = (t_0 - t)$ as the travel time difference, then Δt becomes

$$\Delta t = t_0 \frac{V}{V_s} = \frac{lV}{V_s^2} \tag{8.9}$$

Since the measurement of t_0 is generally not provided for, in the present arrangement, it is preferable to have an additional set of the transmitter–receiver system along with the present system (Fig. 8.46(b)) to determine the transit time, against the direction of flow. If t_1 is the transit time along the flow and t_2 the transit time against the flow, then we get the travel time difference Δt as

$$\begin{aligned}\Delta t &= t_2 - t_1 \\ &= \frac{l}{V_s - V} - \frac{l}{V_s + V} = \frac{2lv}{V_s^2 - V^2} \\ &= \frac{2lv}{V_s^2} \qquad (\because V << V_s)\end{aligned}$$

Thus, the output signal proportional to Δt is linear in V for constant V_s. However, the calibration constant is strongly dependent on the temperature and the pressure of the flow field.

Another approach in the technique is the oscillating loop system (Fig. 8.46(c)), also called the frequency difference method. In this system, the effect of sonic velocity is eliminated by arranging each pair of transducers in an oscillating loop. A pulse is emitted by the transmitting transducer T_1 (pulse generator) and is received by the receiving transducer R_1 after time t_1. This pulse is amplified and instantaneously fed back to the transmitting transducer for retransmission. This generates a train of pulses in each path whose time period equals the accoustical travel time. Now the repetition frequencies along and against the fluid flow are

$$f_1 = 1/t_1 = \frac{V_s + V}{l} \quad \text{and} \quad f_2 = 1/t_2 = \frac{V_s - V}{l}$$

Now the frequency difference or beat frequency becomes

$$\Delta f = f_1 - f_2 = \frac{2V}{l}$$

Thus, the frequency difference is directly proportional to 'V', which affords a useful linear relationship. In addition, the expression is independent of the value of sonic velocity, V_s. However, this method has the drawback that the frequency difference is usually very small and continued interval is needed in actual practice resulting in a selectively large response time for reasonably good resolution.

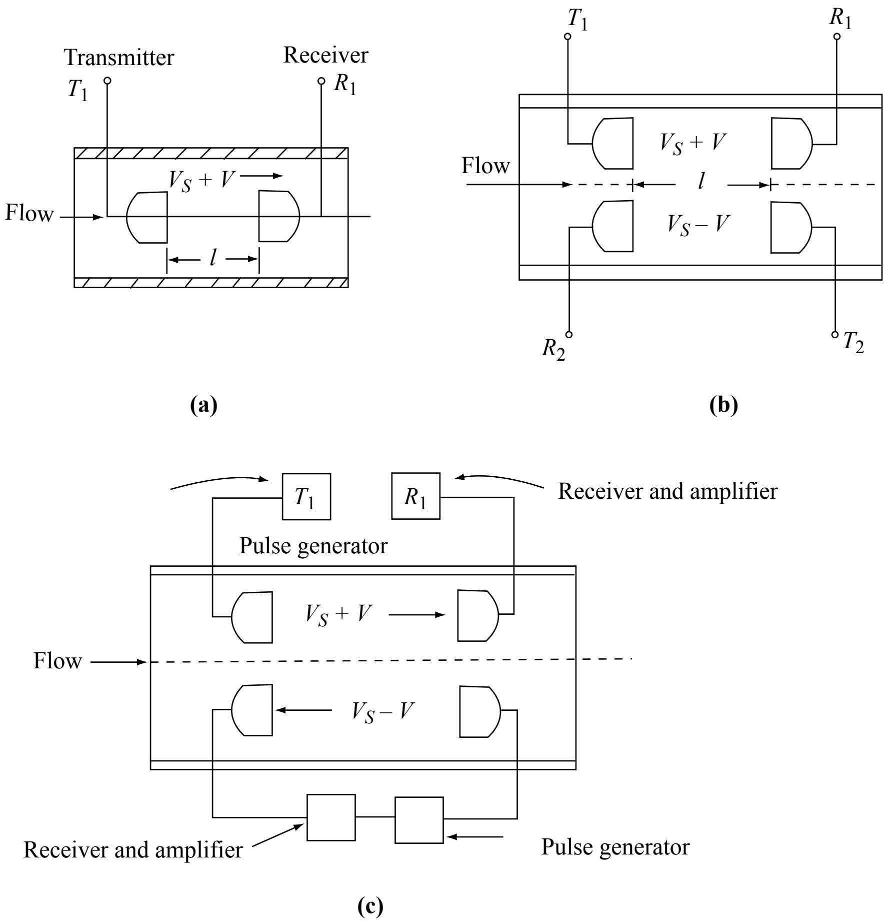


Figure 8.46 Ultrasonic flow meters: (a) travel difference method (single–transmitter receiver system) (b) travel time difference method (twin transmitter–receiver system) and (c) oscillating loop method

The following are the advantages of an ultrasonic flow meter:

1. Offers negligible resistance to the metered fluid system.
2. Has reasonably good accuracy, of the order of ± 2% of the full-scale value.
3. Has a linear relationship between the velocity and the output.
4. Is suitable for both liquids and gases.
5. Since the output is electrical, the readout can easily be either analog or digital.

Because of these *advantages*, the device finds special applications, namely:

(1) Measurement of ocean currents.
(2) Vessel speeds.
(3) Water flows in large conduits.
(4) Flow or various bio-medical and industrial fluids.

The main *disadvantage* of the instrument is its relatively high cost due to close tolerances and high accuracies involved in the design and manufacture of mechanical, acoustical, and electronic portions of the system. This has somewhat limited its wide use in industrial applications presently.

8.15.3 Electromagnetic Flow Meter

The rate of flow may be determined, for electrically conducting fluids, by measuring the *emf* induced across the fluid stream when it passes through the magnetic field. However, it gives quite satisfactory results for fluids with as low conductivity as water.

The principle of operation of the unit is directly analogous to Faraday's law of electromagnetic induction for solid conductors. The law states that whenever a conductor cuts lines of magnetic field, an induced *emf* is generated and the magnitude of this *emf* is proportional to the rate at which these lines are cut and the *emf* is perpendicular to the plane of conductor and the magnetic field. The direction of the induced *emf* is given by Fleming's right-hand rule (Fig. 8.47(a)).

The construction of the electromagnetic flow meter is shown in Fig. 8.47(b). It consists of the following:

1. A permanent magnet or an electromagnet that may be either AC or DC around a non-conducting pipe.
2. Two electrodes placed at right angles to the magnetic field for picking up the induced *emf*.
3. Fluid flow in the pipe, which is at right angles to the plane of magnetic flux lines and the induced *emf* direction, which is along the line joining the electrodes.

Now for the conducting fluid flows, the induced *emf* E_o according to Faraday's law is:

$$E_0 = Blv \times 10^{-8}\ \text{V}$$

where B is the magnetic flux density in Tesla (wb/cm^2), l the length of the conductor in cm, and v the velocity of the conductor in cm/s.

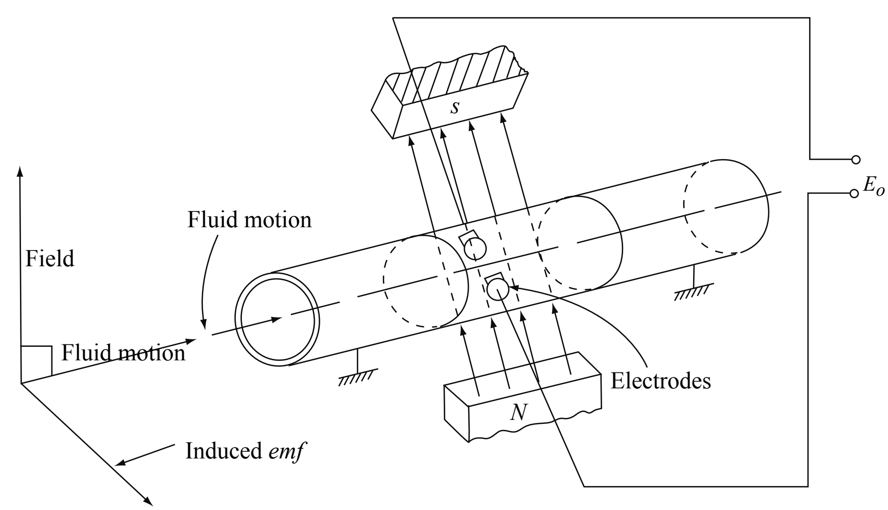


Figure 8.47 Fleming's right-hand rule

The effective length of the conductor corresponds to the inner diameter of the pipe, and the velocity of the conductor is proportional to the mean flow velocity. The volume flow rate for the circular pipe is given by $Q = (TT/4)d^{2v}$. Thus, the above equation can be modified as

$$E_0 = \frac{4B}{d} Q \times 10^{-8}\ \mathrm{V}$$

This shows that the volume flow rate Q is directly proportional to the induced *emf* E_0 as long as the flux density remains constant.

The *emf* is measured by means of two electrodes built into the non-magnetic length of the pipe. Further, these electrodes are made completely flush with the inner surface so that they do not obstruct the flow and at the same time are in direct contact with the flowing liquid. Unfortunately, the *emf* generated by the instrument is very small, i.e., typically of the order of 1 mV, and the resistance of the fluid is often very high. Therefore, the output of the instrument is generally amplified suitably.

In the practical equipment, an alternating magnetic field is preferred. The main reason is that it prevents polarization of the electrodes (i.e., collection of gas bubbles on the electrodes due to electrolytic action, which forms insulating pockets in the vicinity of the electrodes).

The main advantage of the electromagnetic flow meter is that it causes no obstruction in the flow line of the metered fluid. This makes it particularly suitable for fluids containing solid matter. Further, the device is quite accurate and has a wider linear range with good transient response. That is why this instrument is widely used for metering corrosive acids, cement slurry, sewage, paper pulp, detergents, greasy and sticky fluids, etc. However, these meters are usually expensive and their use is limited to fluids having a conductivity at least of the order of 1×10^{-6} /cm.

8.15.4 Theory of Head Flow Meters

Head flow meters operate on the principle of converting energy from one form to another. In the case of liquid, two forms of energy are considered, namely kinetic and pressure energies. In case of gas or vapour, a third form of energy is involved in the interchange, namely the internal energy of gas or vapour. An orifice plate and a Venturi meter deal with the conversion of pressure energy into kinetic energy of the fluid. In the case of a Pitot tube, conversion is from kinetic energy to pressure energy at the impact tube.

8.15.5 Various Tap Connections

Vena contracta taps: Vena contracta taps are those located at one pipe diameter upstream and at the point of minimum pressure, a short distance downstream from the orifice.

Pressure taps: At 2½ diameter upstream and 8 pipe diameters downstream.

Flange taps: These taps are located 2.5 cm from the surface of the orifice plate.

Corner tap: This is similar to the flange tap, except that pressure is measured at the corner between the orifice plate and the pipe wall.

8.15.6 Advantages and Limitations

Vena contracta: This is the optimum location. A tap location too far downstream in the unstable area may result in inconsistent measurement. It should also be noted that the tap location at points of extreme pressure instability should be avoided.

Pressure taps: The coefficients are affected by the condition and roughness of the long downstream length of the pipe. Further, the tap holes are at a location where they are difficult to inspect.

Flange taps: These taps have the following advantages:

1. They are located close to the face of the flange where they are accessible for inspection.
2. They are symmetrical and adapt themselves to measurement of flow in either direction.

Corner taps: They are common for all pipe sizes. The small clearances of the passages are a possible source of trouble. These taps are found to be susceptible to dirt, freezing, and clogging with hydrates when measuring natural gas and are more dependent on the thickness of the plate and gasket than with other tap connections. They are more affected by upstream disturbances.

8.15.7 Characteristics of Head Flow Meters

Orifice plate: It is easy to install and replace the orifice plate inside the pipe. The economy and ease of manufacture are further advantages. Therefore, head flow metres are used in many applications.

Eccentric and segmental orifice plates: If granular solids are present in a flowing fluid, condensate in steam, vapour, or gas in the liquid, the projecting rim of the concentrate forms a dam and these foreign materials build up in the approach pipe at the plate, causing a change in the distribution of flow.

Venturi meter: In the Venturi tube, the ultimate pressure recovery is attained by a long cone exit. This implies a taper of 10:1 or 20:1 slope of the profile of the cone with respect to the axis of the Venturi (Fig. 8.48).

Since most of the pressure is recovered in the region of high velocity, a large part of the possible recovery can be attained by using a truncated recovery cone having an outlet diameter somewhat smaller than the inlet to the Venturi. The shape or the length of the outlet section has no effect on the coefficient of the discharge of the tube.

In industrial flow measurement, pressure recovery characteristics are of less importance, since any pressure that is recovered is dissipated in some form of regulation or control.

When power costs are not involved the most common reason for the choice of the Venturi tube compared to the orifice plate is the better performance of solids in suspension in the flowing fluid.

Flow nozzle: This consists of a bell-shaped approach section of elliptical profile followed by a cylindrical throat tangent to the ellipse. The axes of long radius nozzle are d and 2/3 d for low diameters and $D/2$ and 1/2 (D-d) for high-diameter ratios (D is the pipe diameter and d the throat diameter). It is found that the throat lengths exceeding 60% of the throat diameter affects the coefficient. Pipe wall taps located approximately one pipe diameter upstream and 1/2 pipe diameter downstream from the inlet face of the nozzle give best results (Fig. 8.49, 8.50).

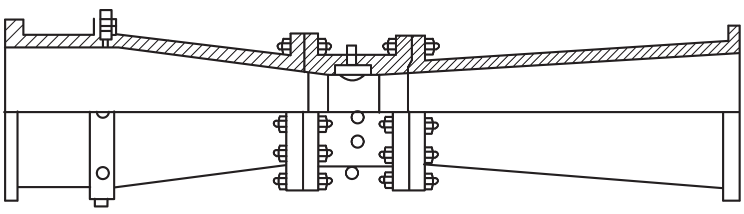


Figure 8.48 Venturi meter

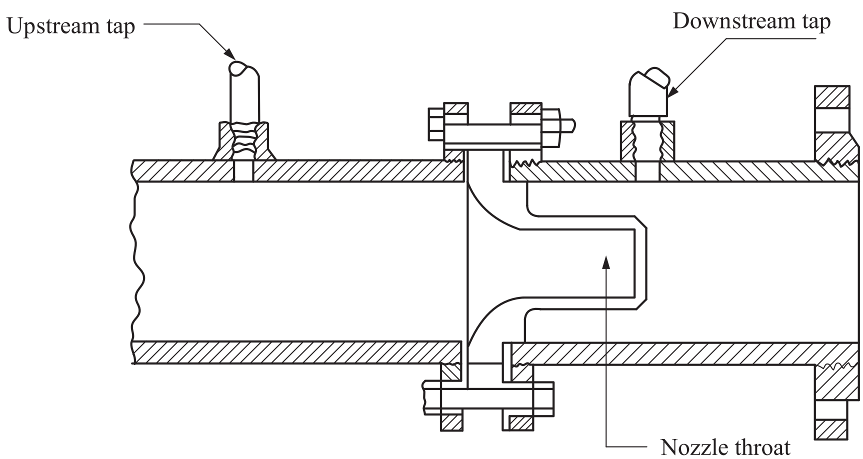


Figure 8.49 Flow nozzle with pipe line taps

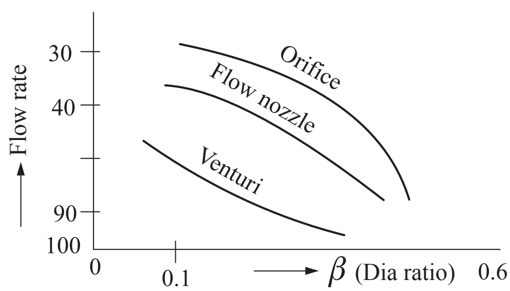


Figure 8.50 Characteristics of head flow metres

Pitot tube: It is a very effective tool for spot checking, but its tendency to plug when the flowing fluid contains small amounts of solid matter, its velocity range limitations when used with standard commercial instruments, and its sensitivity to absorb velocity distribution affect the limitation of this tube (Fig. 8.51).

8.16 AREA FLOW METERS

8.16.1 Rotameters

A rotameter consists of a weighted plummet contained in an upright tapered tube with the large end up. It is lifted to the position of equilibrium between the downward force of the plummet and the upward force of the fluid, flowing past the plummet through the annular orifice. In smaller sizes, the tube is made of glass, which is graduated so that the flow can be read directly by observing the position of the plummet (Fig. 8.52).

8.16.2 Piston-Type Area Meter

The usual piston-type meter comprises a sleeve or cylinder held rigidly in a cast body and a well-fitted piston or metering plug. Orifices, usually rectangular, are cut into the sleeve. These orifices are uncovered by the piston or plug until sufficient area is opened to permit passage of the flow being measured. The metering edges consist of the port edges and the bottom edge of the plug. The position of the piston or metering plug provides a direct indication of the orifice and consequently, the rate of flow. Distribution affects the limitation of this tube (Fig. 8.53).

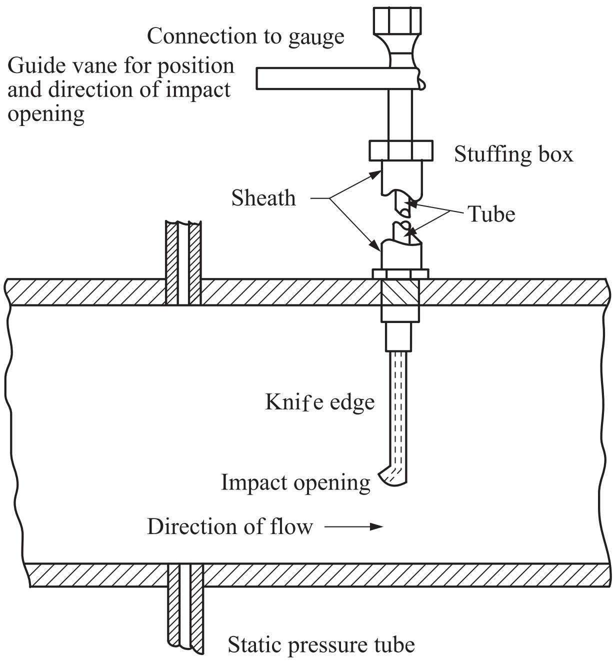


Figure 8.51 Single opening or elementary type of a Pitot tube

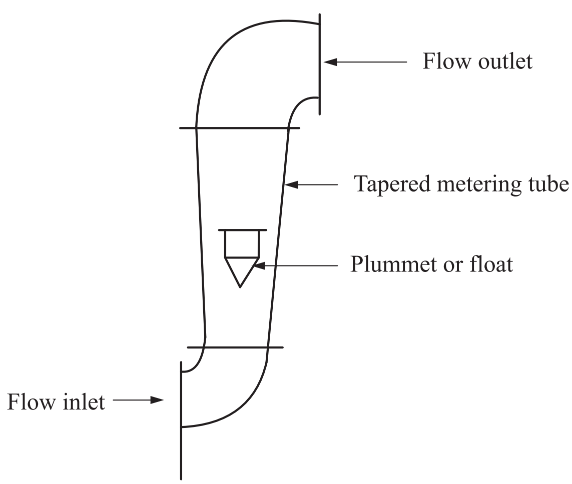


Figure 8.52 Basic components of a rotameter

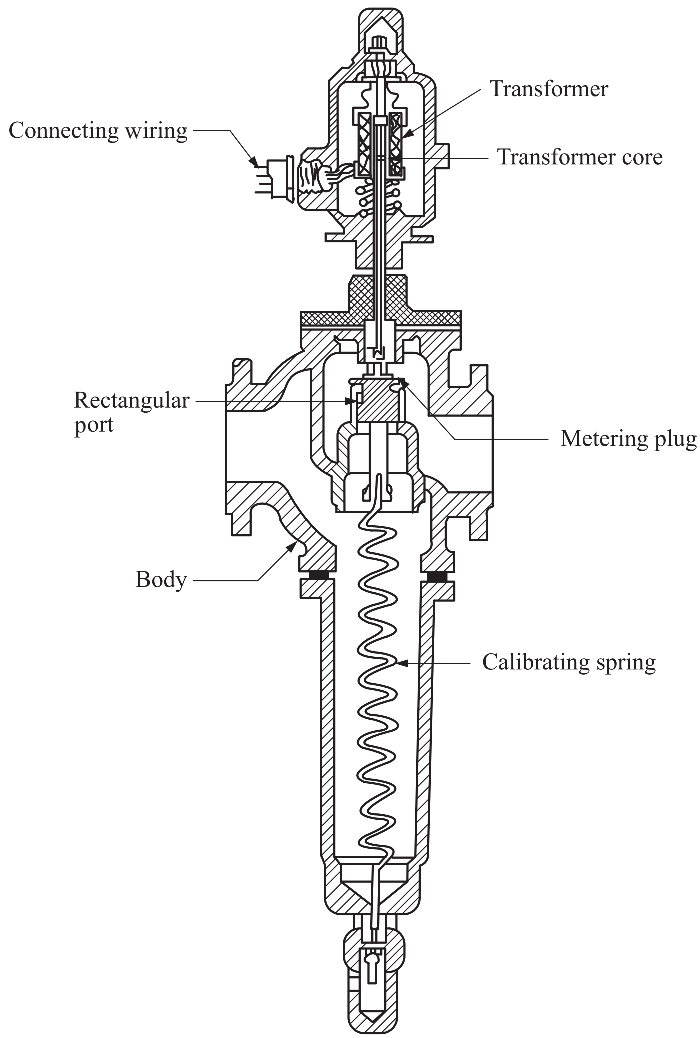


Figure 8.53 Cross-section of a spring-loaded piston-type area meter

8.16.3 Laser Doppler Anemometer (LDA)

The most recent advancement in the area of flow measurement is the laser Doppler anemometer (LDA), which is also known as the optical type of velocity meter. This instrument measures the instantaneous velocities of gas or liquids flowing in a transparent channel.

The operating principle of this device is based on the Doppler shift in frequency of the light scattered by an object moving relative to the radiating source. The technique basically consists of focusing laser beams at a point in the fluid where velocity is to be measured. At this focal point, the laser light scattered from the fluid or fluid particles entrained in the fluid is sensed by a photodetector. The output of the detector after processing gives the magnitude of the Doppler frequency shift, which is directly proportional to the instantaneous velocity of the flow. The use of a laser beam instead of a monochromatic light beam is necessitated because of the following:

1. Laser provides much higher quality of monochromatic (single frequency) light source.
2. Laser beam is coherent, i.e., it stays in phase with itself over long distances.
3. Its frequency is very stable and is precisely known to about 1 part in 10^7. This enables one to accurately detect the relative Doppler shift frequency, i.e., $\Delta f/f$, which is of the order of 10^{-7} for flow velocities of the order of 45 km/hr.
4. Unlike a monochromatic beam, the wavelength of the laser beam is deflected minimally by changes in the ambient pressure, temperature, or humidity.

There are several optical arrangements for the LDA. The most commonly employed LDA is the dual beam or the fringe mode shown in Fig. 8.54. In this the laser source employed is usually the helium–neon laser of 5–15 mW power. When the measured flow is at a distance of 10 to 20 cm, this laser operates at a wavelength of 632.8 mm (5×10^{14} Hz). This laser beam is split into two equal parts by means of a beam splitter in the form of either a rotating optical grating or an optical prism or a half silvered mirror pieced 45° to the beam. The focussing lens is placed in front of the fluid to be measured. At the total point, the two beams interfere with each other to form an interference fringe pattern that consists of alternate regions of high and low intensity. If the tiny tracer particles (dust or dirt particles present in water or air flows) pass through the region of high intensity, they would scatter light and cause a Doppler shift in the frequency of the scattered light. Thus, the light received by the photodetector will show a varying electrical signal, the frequency of which is proportional to the rate at which the particles cross the interference fringes.

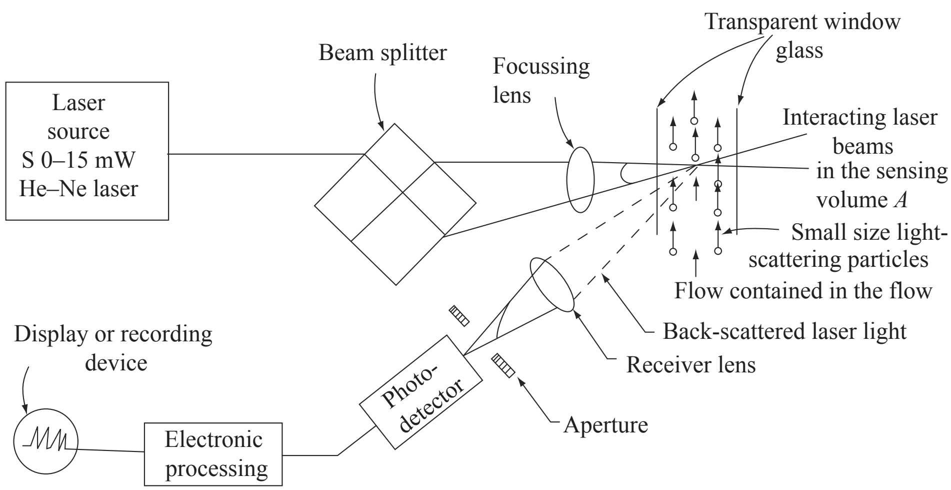


Figure 8.54 Doppler anemometer in dual beam or fringe mode

The spacing between the fringe is given by the expression $\bar{x} = x/2 \sin\left(\dfrac{\theta}{2}\right)$, where θ is the angle between the two converging beams and λ the wave length of the laser beam.

If the tracer particle (assumed to have a velocity equal to that of the fluid) passes across the fringes with a velocity v in a direction perpendicular to the fringes, the signal would experience a Doppler shift in the frequency given by $\Delta f = \dfrac{2v}{\lambda}$ $(\theta/2)$, where λ is the wavelength of the laser beam in the fluid (which may be different from the vacuum wavelength by a factor equal to the index of refraction). The above equation may be expressed alternatively as $\Delta f = \frac{2nv}{\lambda_o} \sin(\theta/2)$, where n is the index of refraction of the fluid and λ_o the wavelength of the laser beam in vacuum.

The LDA has a number of significant advantages over the hot wire velocity metering techniques. They are as follows:

1. There is no transfer function involvement, i.e., the output voltage of the instrument is directly proportional to velocity. Therefore, the instantaneous velocity rather than its inference from pressure difference or heat transfer phenomenon can be measured accurately
2. Non-contact type of measurements, i.e., no physical objects, are inserted in the flow field and thus the flow is undisturbed by the measurement.
3. Very high-frequency response of the order of MHz is possible. Therefore, its accuracy in the determination of fluctuating velocities is far superior to that of a hot wire anemometer because there is practically no time lag in the measurement.
4. Very high accuracies of the order of ±0.2%.
5. Measurements are possible in a miniature size of volume of the order of a 0.2-mm size cube.
6. Suitable for measurements in both gas and liquid flows.
7. The instrument has established proven superiority over the other methods of measurement in the following areas:
   - Investigation of boundary layers and shock wave interaction phenomenon for both laminar and turbulent flows.
   - Determination of three-dimensional wing tip vortices near the tips of the aircraft wings.
   - Measurements of flow between the blades of turbines.
   - Combustion and flame phenomenon in gas turbines and jet propulsion systems.
   - In vivo measurement of blood flows.
   - Remote sensing of wind velocities

However, the instrument has the following disadvantages:

1. It involves the need for a transparent channel.
2. The measurement technique is not suitable for clean flows. For such flows, the tracer particles have to be seeded in the flow for scattering the incident light.
3. The instrument is quite expensive and requires a high degree of experience and skill in operation.

8.17 POSITIVE DISPLACEMENT METERS

8.17.1 Basic Requirements

1. Simplicity of design is required to make maintenance possible without specially trained personnel.
2. Accuracy within stated limits.
3. Availability in a wide variety.
4. Availability in different materials and calibrations for the measurement of widely different materials.
5. Reasonably low pressure loss.

8.17.2 Nutating Piston Meters

A nutating piston meter is also known as a disk meter. Each cycle of the measuring piston or disc displaces a fixed volume of liquid. There is only one moving part, namely the piston. The liquid enters through the inlet port and fills the spaces above and below the piston, which fits closely in the measuring chamber (Fig. 8.55).

The advancing volume of liquid moves the piston in a nutating motion until the liquid discharges from the outlet port.

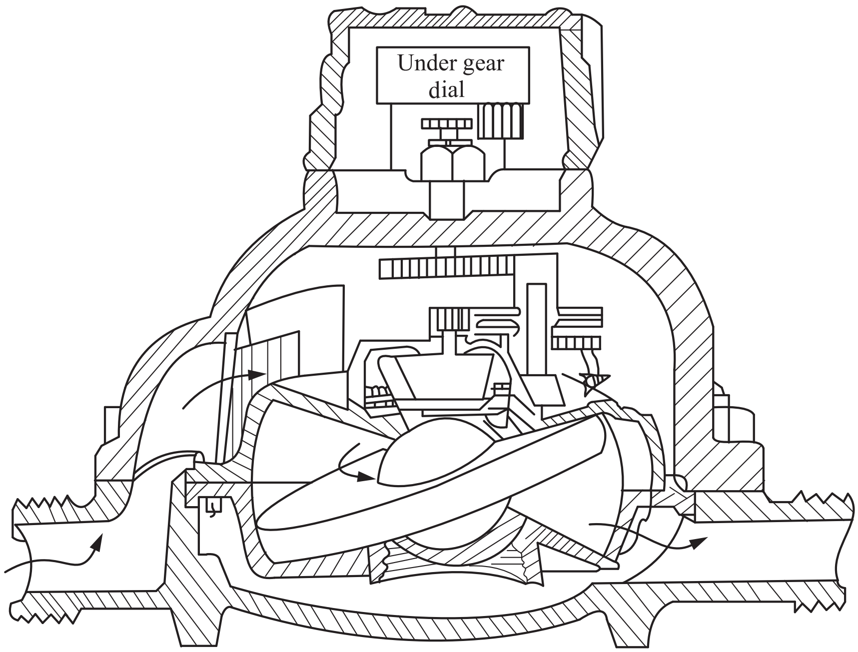


Figure 8.55 Sectional view of a typical nutating piston meter

8.17.3 Rotating Meters

They are currently known as current meters or velocity meters. They operate on the turbine principle, that is, the volume is measured by the movement of a wheel or turbine type of impeller, which is actuated by the velocity of the liquid flowing through it. They are used to measure continuous high flow rates with minimum pressure loss.

8.17.4 Oscillating Piston Meters

In this type, the measuring chamber is cylindrical with a central abutment and division plate separating the inlet port A on one side and the outlet port B on the other. The balanced piston is also cylindrical, but it has a horizontal web carrying a post in the centre and is slotted to clear the division plate. The post, extending below the web, guides the piston within the measuring chamber around a roller in a lower abutment well. The post also extends above the web to drive an arm by which its motion is transmitted through a set of reducing gears to a register, which records the quantity of water (Fig. 8.56).

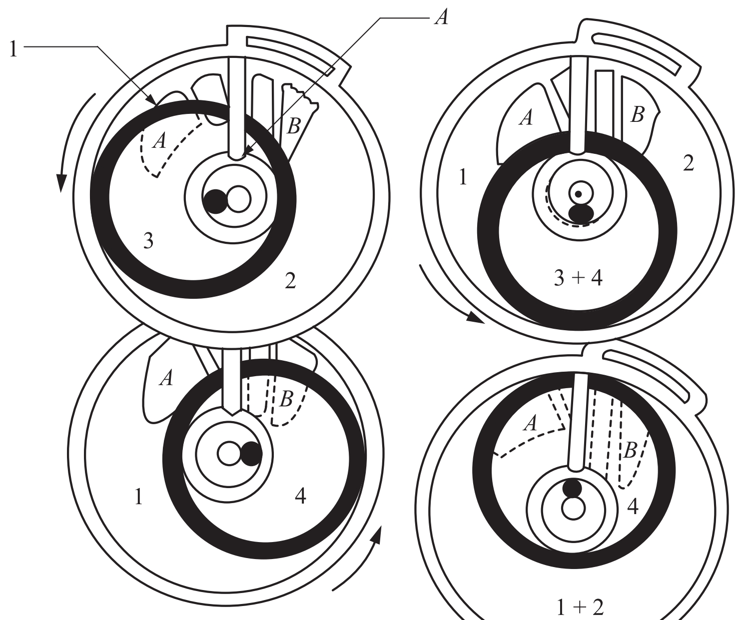


Figure 8.56 Oscillating piston meter

8.18 MAGNETIC FLOW METER

This meter is used to measure the rate of flow of fluids, which present extremely difficult handling problems such as corrosive acids, rayon, viscose, sewage, rock, and acid slurries. This is based on Faraday's law of electromagnetic induction:

$$E = chdv$$

where E is the voltage induced, c the constant, d the length of conductor, h the magnetic field, and v the velocity of the conductor (Fig. 8.57).

The fluid to be measured is the moving conductor. It passes through the tube and the alternating magnetic field h. The conductive fluid is analogous to a continuous series of fluid disk, the diameter of which equals the inside diameter of the tube. The disks flowing through the tube are conductors of length d, where d is equal to the diameter of the disk.

8.18.1 Flow Meter Requirements

1. Wide range.
2. Obstructions.
3. Linear.
4. Immune to viscosity and temperature.
5. For a large variety of pipe sizes.

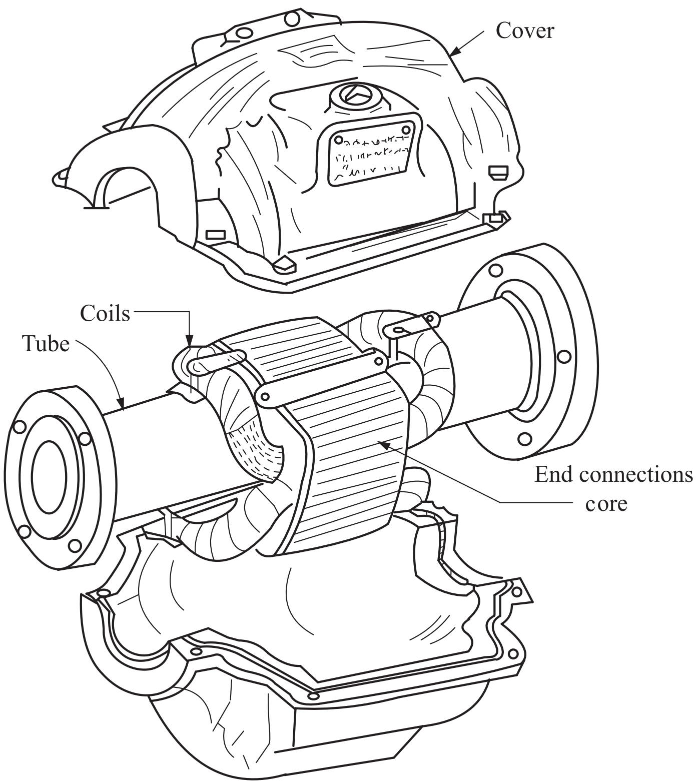


Figure 8.57 Enlarged view of a magnetic flow meter

8.19 VARIABLE CAPACITANCE TRANSDUCERS

One of the basic three parameters of electrical circuits is capacitance. This capacitance is the function of area between two plates, the separation between them and the dielectric medium in between.

C for parallel plate condensers is

$$C = \frac{A\varepsilon}{x}$$

where A is the area of the plate, ε the permittivity of the medium, and x the distance between the area and permittivity.

Therefore, if one of the above three factors change, the capacitance changes. This change in capacitance can be suitably transduced into a voltage or frequency signal. For a voltage signal, output bridge circuits and frequency variation oscillator circuits are employed. The capacitance is connected to the voltage and the charge by the equation $Q = VC$, where Q is the charge in coulombs, V is voltage in volts, and C is capacitance in farads.

$$\frac{dQ}{dt} = V\frac{dC}{dt} + C\frac{dC}{dt}$$

Any variable or parameter that affects the capacitance can be transduced into a voltage or current variation. For example, linear displacement, rotary displacement, velocity, acceleration, pressure level, and thickness of material are some of the parameters that can be transduced. The most popular form of a variable capacitor used in a motion measurement is the parallel plate capacitor with a variable air gap. The main problem in this case is the high output impedance.

For example, let us consider a capacitor of area 4 cm^2 and 0.02-cm gap. The capacitance of this is

$$C = \frac{A\varepsilon}{x} = \frac{4\times 10^{-4}}{0.02\times 10^{-2}} \times 8.854 \times 10^{-2}\ (\text{F/m}) = 17.78\ \text{pF}$$

The impedance of this capacitor at a frequency of 10,000 Hz is

$$\frac{1}{2} = \frac{1}{2\pi \times 10^{4} \times 17.78 \times 10^{-12}} = 89{,}000\ \Omega = 89\ \text{k}\Omega$$

Because of this high impedance level, the noise level is high resulting in complex electronic circuitry. The other problem is that the variation in capacitance with respect to x is non-linear

$$C = \frac{8.854A}{x}\ \text{pF}$$

$$\frac{dC}{dx} = \frac{-8.854A}{x^2}$$

The sensitivity increases as x decreases in a parabolic manner. However, the percentage change in C is equal to the percentage change in x

$$\frac{dC}{dx} = \frac{-C}{x}; \quad \frac{dC}{C} = \frac{-dx}{x}$$

8.19.1 Advantages of Capacitance Transducers

1. Its force requirements are very small.
2. As the moving plates have very little mass, design of transducers with highly desirable response characteristics is possible.
3. There is no necessity for a physical contact.
4. The capacitance does not depend on the conductivity of the metal electrodes or its variation with temperature.
5. The capacitance transducer can be shielded easily against the effect of external electric stray fields.

8.19.2 Practical Capacitor Pickups

Equibar differential pressure transducer: Differential pressure can be transduced by a three-terminal capacitor as shown in Fig. 8.58.

Spherical depressions of a depth of about 0.00254 cm are ground into the glass disks. These depressions are gold coated for fixed plates of a differential capacitor. A thin stainless steel diaphragm is clamped between the disks and serves as the movable plate. With equal pressure applied to both parts, the diaphragm is in a neutral position, the bridge is balanced, and $e_o = 0$.

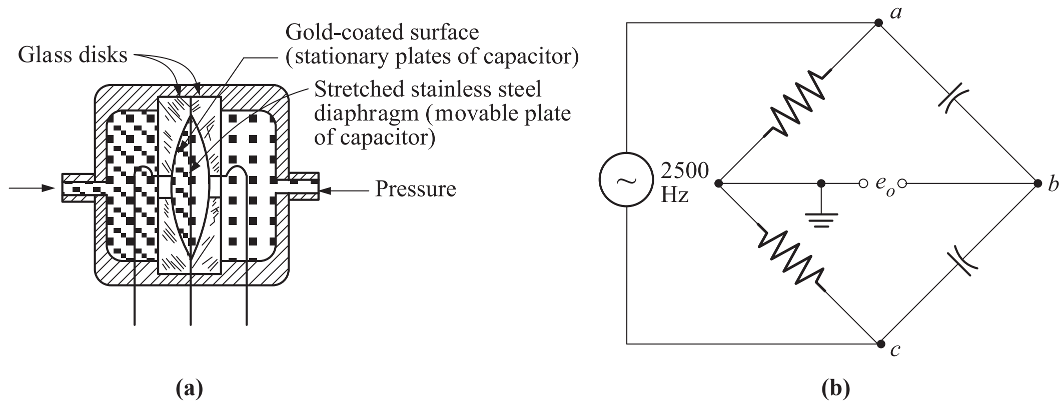


Figure 8.58 (a) Cross-section of a capacitance transducer and (b) bridge circuit

If one pressure is greater than the other, the diaphragm deflects, giving an output e_o in proportion to the differential pressure. For the opposite pressure difference, e_o exhibits a 180° phase change. The high-impedance level requires a cathode follower amplifier at e_o. A direction-sensitive DC output can be obtained by conventional phase-sensitive demodulation and filtering. Balance resistors are not shown in Fig. 8.58.

8.19.3 Feedback-Type Capacitance Pickup

This circuit in Fig. 8.59 gives a very good linearity. This makes use of a high-gain operational amplifier.

This amplifier has got the following characteristics:

1. High-input impedance.
2. Gain is of the order 10^5–10^8.
3. Low-output impedance.

The equation governing its operations as the follows. Applying Kirchhoff's current law,

$$i_f - i_g + i_x = 0$$

$$i_f = C_r \frac{de_{ex}}{dt} \text{ as } G \text{ is the ground potential}$$

$$i_x = C_x \frac{de_o}{dt} \text{ as } G \text{ is the ground potential}$$

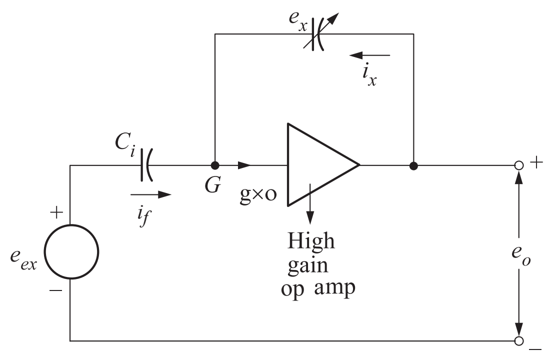
$$i_r = -i_x$$

$$C_f \frac{de_{ex}}{dt} = -C_x \frac{de_o}{dt}$$

$$e_o = -\frac{C_f}{C_x} e_{ex} = \frac{-C_f x}{C_x} e_{ex} = K_x$$

Now the output voltage is directly proportional to the plate separation x. Linearity is thus achieved for both large and small motions. In commercial instruments using op amps, e_{ex} is a 50 kHz. Since wave of fixed amplitude. The output e_o is also a 50 kC sinewave, which is rectified and applied to a

DC voltmeter calibrated directly in distance units. For vibratory displacement, e_o will be an amplitude-modulated wave as shown in Fig. 8.60.



C_f is a fixed capacitor

C_r is the displacement-varying capacitor

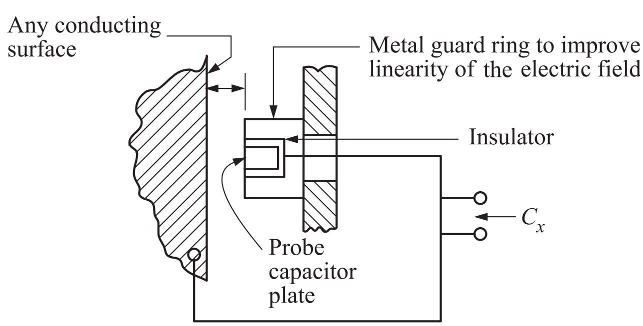


Figure 8.59 Feedback-type capacitance pickup

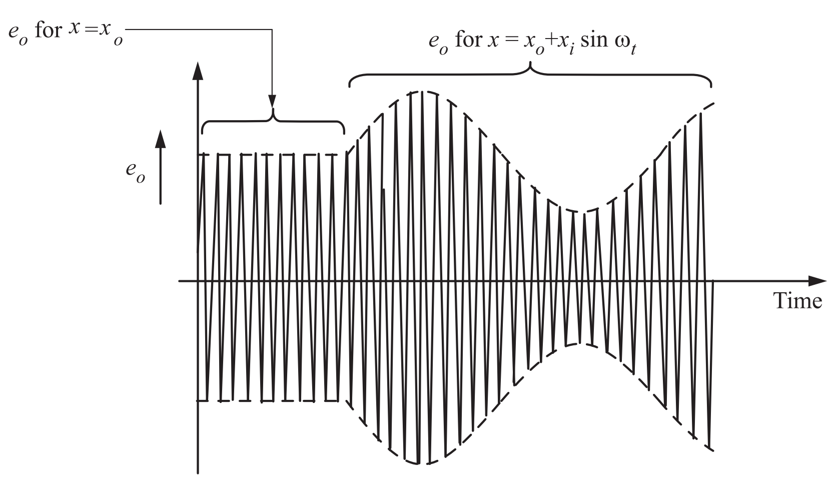


Figure 8.60 Output waveforms

The average value of this wave after rectification is the mean separation of the plates and can still be read by the same meter as used for static displacements. The vibration amplitude around this mean position is extracted by applying the e_0 signal to the demodulator and a low-pass filter with the cutoff at 10 kHz. The output of this filter is applied to a peak-to-peak voltmeter directly calibrated in connection to an oscilloscope for viewing the vibration waveform. The instrument is provided with six different probes ranging from 0.11 to 2.54 cm in diameter of the capacitance plate and covering the full-scale displacement ranges of 0.0025, 0.0125, 0.025, 1.25 cm, respectively. The overall system accuracy is of the order of 2% of the full scale. Any flat conducting surface may serve as the second plate of the variable capacitor. Thus, in vibrating machine parts, the parts themselves may often perform this function. The resolution of 0.5% of the full scale indicated will be 0.0025 cm. The full-scale probe is capable of detecting motion as small as 12.5 μcm.

The capacitance will be maximum for the position shown in Fig. 8.61 and minimum for a shaft position 180° to the position shown in Fig. 8.62.

Here the shaft has to make an electrical contact through the bearing. The connection through the bearing is unreliable. Therefore, a split stator capacitor is used, which does not require electrical contact with the rotor plate. Many a transducer structure or dielectric configuration can be used. Some of them are shown in Fig. 8.62.

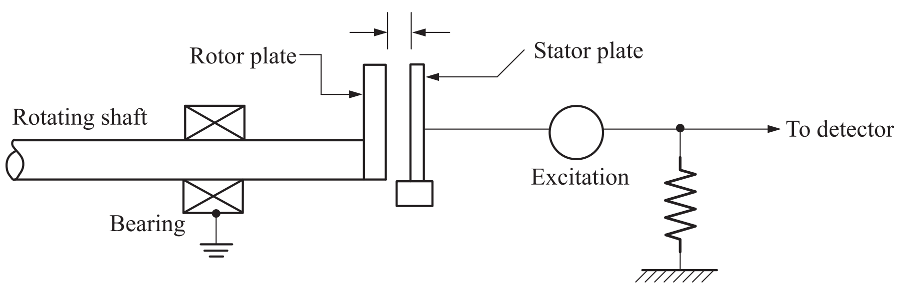


Figure 8.61 Position of transducer plate for maximum capacitance

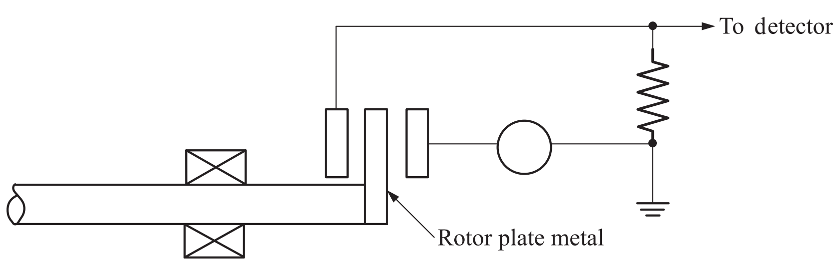


Figure 8.62 Position of transducer plate for minimum capacitance

8.19.4 Carbon Microphone

Figure 8.63 is a simplified version of a typical capacitor microphone. The pressure response is found by assuming a uniform pressure p_1 to exist all around the microphone at any given time. This is actually the case of sufficiently low sound frequencies, but reflection and diffraction effects distort this uniform field at higher frequencies. The diaphragm is generally a very thin metal membrane, which is stretched by suitable clamping arrangement. The diaphragm thickness ranges from about 0.000254 to 0.0051 cm. The diaphragm is deflected by the sound pressure and acts as a moving plate of a capacitance displacement transducer. The other plate of the capacitor is stationary and may contain properly designed damping holes. The damping effect is used to control the resonant peak of the diaphragm response. A capillary air leak is provided to give equalisation of steady pressure on both sides of the diaphragm to prevent the diaphragm from bursting. The variable capacitor is connected to a simple series circuit with a high resistance R and polarised with a DC voltage of about 200 V. This polarising voltage acts as circuit excitation and also determines the neutral diaphragm position.

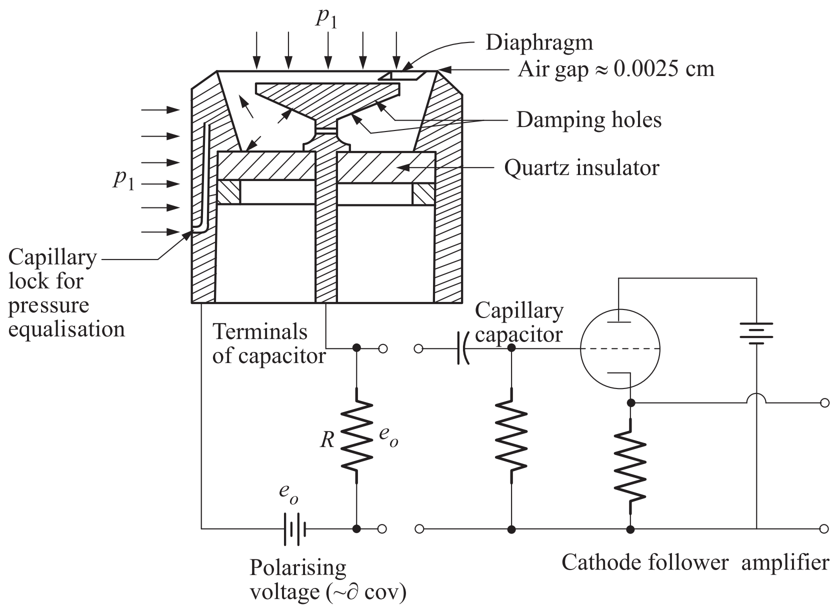


Figure 8.63 Carbon transducer and circuit

8.19.5 Circuitry for Capacitance Transducers

Capacitance transducers being passive require external power to convert the capacitance into voltage, which can be pre-calibrated in terms of the measurand (Figs. 8.64, 8.65). The various types of circuits utilised for the purpose can be classified into four categories:

1. Frequency-modulating oscillator circuit.
2. Circuit employing DC excitation.
3. AC bridges for amplitude modulation.
4. Pulsewidth-modulating circuit.

8.19.6 Frequency-Modulating Oscillator Circuit

For a low-frequency variation of the measurand, either the Wien bridge oscillator or the phase shift oscillator is used. The Wien bridge oscillator (Fig. 8.66) in view of many of its good features has become a universal circuit. The advantages are:

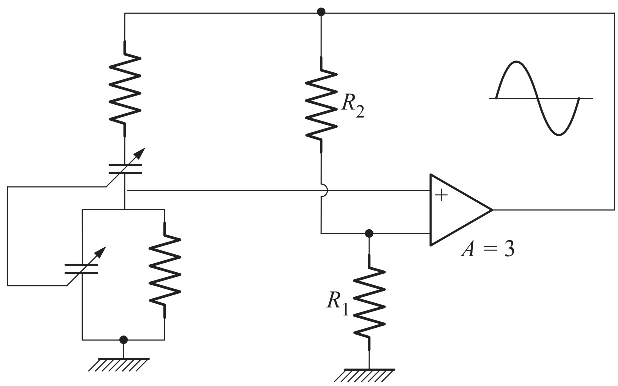


Figure 8.64 Frequency-modulating oscillator circuit for a carbon microphone transducer

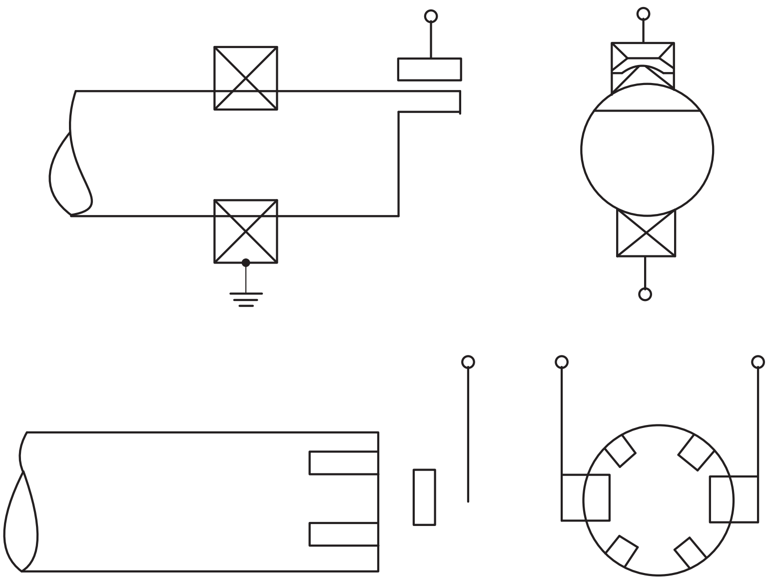


Figure 8.65 Carbon microphones

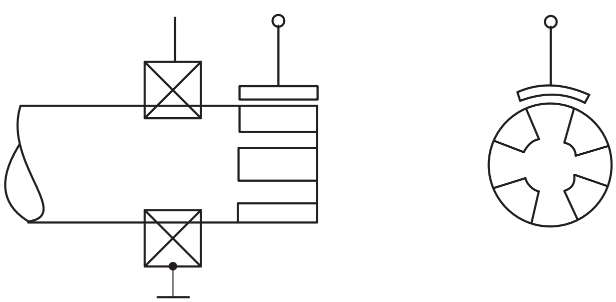


Figure 8.66 Cross-section of a transducer

(i) Low distortion, less than 1%.
(ii) Amplitude stability against aging, temperature variation, and supply variation using a thermistor in the negative feedback path.
(iii) Low output impedance.

The disadvantage is that a gauged dual capacitor is necessary. Therefore, if a single capacitor is to be employed, the Hartley oscillator configuration is to be used as shown in Fig. 8.67. The lowest centre frequency is of the order of tens of kHz but 1 MHz is a typical value. This is not suitable for measurement, and a frequency less than 10 Hz as the frequency stability of the low price oscillator cannot be better than 100 parts per million.

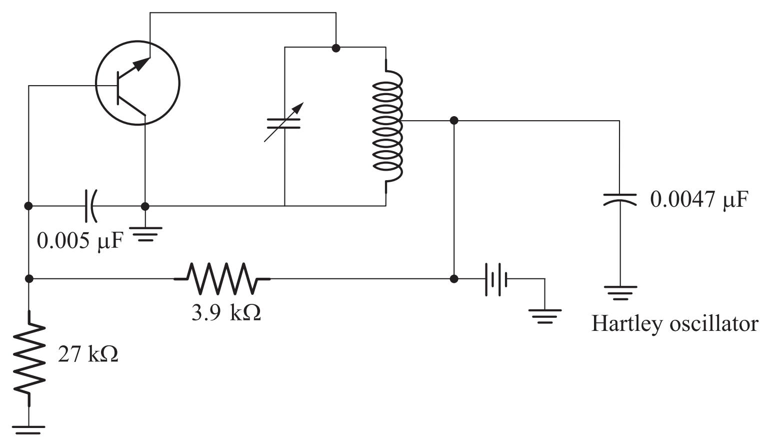


Figure 8.67 Hartley oscillator configuration for a carbon microphone transducer

8.19.7 Circuits Using DC Excitation

With reference to Fig. 8.68, the following equation can be written:

$$\frac{E-V}{R} = i = \frac{dQ}{dt} = \frac{d}{dt}(VC) = C\,\frac{dv}{dt} + V\,\frac{dc}{dt}$$

$$\frac{E}{R}-\frac{V}{R} = C_o\,\frac{dv}{dt} + E\,\frac{dc}{dt} = \frac{A\in}{x_o}\,\frac{dv}{dt} - E\,\frac{A\in}{x^2}\,\frac{dx}{dt}$$

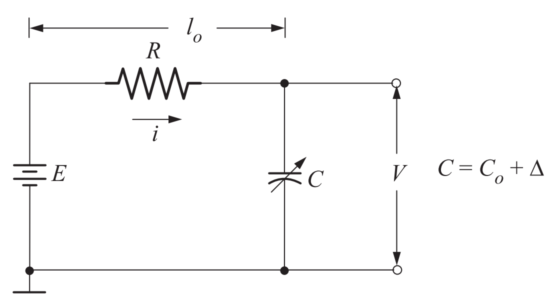


Figure 8.68 DC excitation circuit for transducers

Let $e_o = E + V$:

i.e.,
$$\frac{e_o}{R} + \frac{RA\in}{x_o}\frac{de_o}{dt} = \frac{EA\in}{x_0^2}\frac{dx}{dt}$$

$$e_o(s)\left(1+\frac{RA\in S}{x_o}\right) = \frac{E}{z_o}\frac{RA\in}{x_o} \quad Sx(s) \text{ where } S \text{ is operator } \frac{d}{dt}$$

$$\frac{e_o(s)}{x(s)} = \frac{K\tau S}{1+\tau S}$$

where

$$K = \frac{E}{x_o}$$

$$\tau = \frac{RA\in}{x_o}$$

The frequency response of this circuit is given in Fig. 8.69, which shows that static displacement cannot be measured with this circuit.

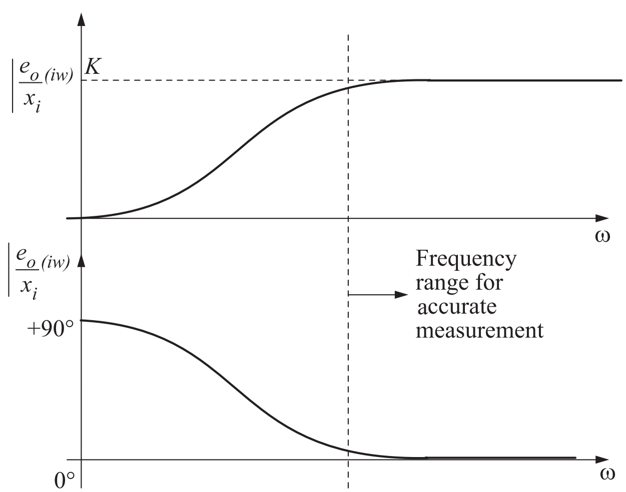


Figure 8.69 Output waveforms

8.19.8 AC Bridges for Amplitude Modulation

Figure 8.70 shows a simple arrangement for converting capacitance variation into voltage. The detector may be calibrated to read the displacement directly.

A twin-network shown in Fig. 8.71 is used more often than the other capacitance circuits. This network has the following advantages:

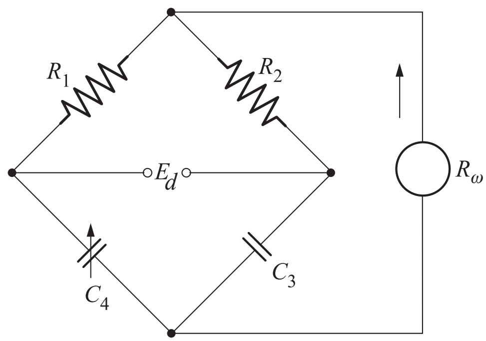


Figure 8.70 Circuit for converting capacitance variation into electrical voltage

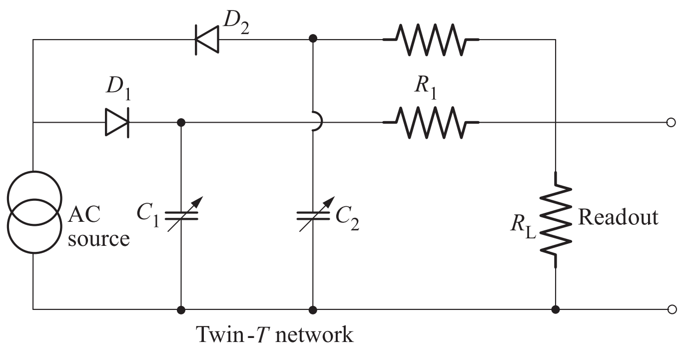


Figure 8.71 Twin *T*-network for a capacitance transducer

1. The source *S*, the transducer capacitor C_1, the comparison capacitor C_2, and the output circuit are grounded.
2. The diodes D_1 and D_2 are operated at a high level so that they operate in the linear region of their characteristics.
3. The output voltage is remarkably high. It has been observed that when operated with a sinusoidal input voltage E_1 of 46 V rms at a frequency of 1.3 MHzs, a variation in the capacitance difference from 0 to +7 pF causes a change in the output voltage from –5 to +5 V DC (which is 10 V swing) into a 1 MΩ load output voltage from –16 to +16 V.
4. The noise level in the output measured in a bandwidth of 7 Hz is of the order of 10 μV related to a typical output voltage of 10 V. The signal to noise ratio is 106 (S/N = 120 dB).
5. The rise in time of the output signal depends on the load resistance. For a load resistance of 1000 Ω the rise in time is of the order of 20 μs so that fast mechanical movements can be measured.

The working of this bridge is as follows: During positive half cycle of the applied AC, current flows through diodes D_1 and C_1, thus charging C_1. During the other half of AC, Capacitor C_2 is charged and at the same time C_1 discharges through R_1 and R_L. And C_2 discharges through R_L and R_2 when C_1 is charged.

If C_1, D_1, and R_1 are identical to C_2, D_2 and R_2, the average current through R_1 is zero. This happens when there is no pressure difference, there is no difference between C_1 and C_2. A change in pressure will change C_1 from C_2 and hence a net current will flow through R_L.

This network is capable of producing around ± 5 V.

Diode pump. A more versatile circuit to transduce frequency or capacitance variation into a voltage is the diode pump. Figure 8.72 shows the details of this circuit used to transduce the capacitance variation into voltage variation. The operation of the diode pump circuit can be explained as follows. The transducer is C_1, which is very small compared to C_2. The time constant RC_2 should be sufficiently small compared to the period of the pulse input. When the pulse goes positive, C_1 charges to the peak magnitude V. When the pulse goes to zero, D_2 conducts, so the output voltage is $V C_1/(C_1 + C_2)$. When the pulse goes positive next, D_2 is not conducting and C_2 discharges through R. The waveforms are shown in Fig. 8.73.

The average value $V_{dc} = V_R C_2$. Therefore, if V and R are constants, the output is directly proportional to C_2. Typical values for the circuit shown in Fig. 8.71 is C_1 = 10 μF, R = 10 kΩ, C_2 = 20–40 pF, and the output is 1 mV/pF.

8.19.9 Pulsewidth-Modulating Circuit

A differential capacitor variation is transformed into a pulsewidth by the circuit shown in Fig. 8.74.

The circuit shown in Fig. 8.75 is a versatile IC version of the above principle for transducing the capacitance variation. The output pulsewidth is given by the equation. When the trigger is applied, the output goes high and the AND gate is abled and the counter starts counting.

The counting is inhibited after a time T = 1.1 RC. For example, if we use a 4-digit counter and the clock is 1 kHz, 100 pF can be displayed as 100. 0, if T = 1 s

$$T = 1.1R \times C$$

$$R = \frac{1 \text{ sec}}{1.1 \times 100 \times 10^{-12}} = \frac{1 \times 10^{12}}{1.1 \times 100} = 910 \text{ M}\Omega$$

$$R = 910 \text{ M}\Omega$$

8.20 PIEZOELECTRIC TRANSDUCER

When certain solid materials such as quartz, lithium sulphate, and barium titanate are deformed, electric charge is generated within themselves. This effect is reversible if a charge is applied and the material will mechanically deform in response. These actions are given the name piezoelectric effect (piezo means force).

The mechanical input/electrical output direction is the basis of many instruments for measuring acceleration, force, and pressure. It can also be used as a means of generating high-voltage low-current electrical power such as that used in spark-ignition engines and electrostatic dust filters. The electrical input/mechanical output direction is applied in small vibration shakers, sonar systems for acoustic detection, and location of underwater objects and industrial ultrasonic non-destructive test equipments.

8.20.1 Materials

Twenty of the thirty-two crystallographic classes exhibit piezoelectric properties, but only a few materials are practical for piezoelectric transducers, primarily quartz, Rochelle salt, ammonium dihydrogen

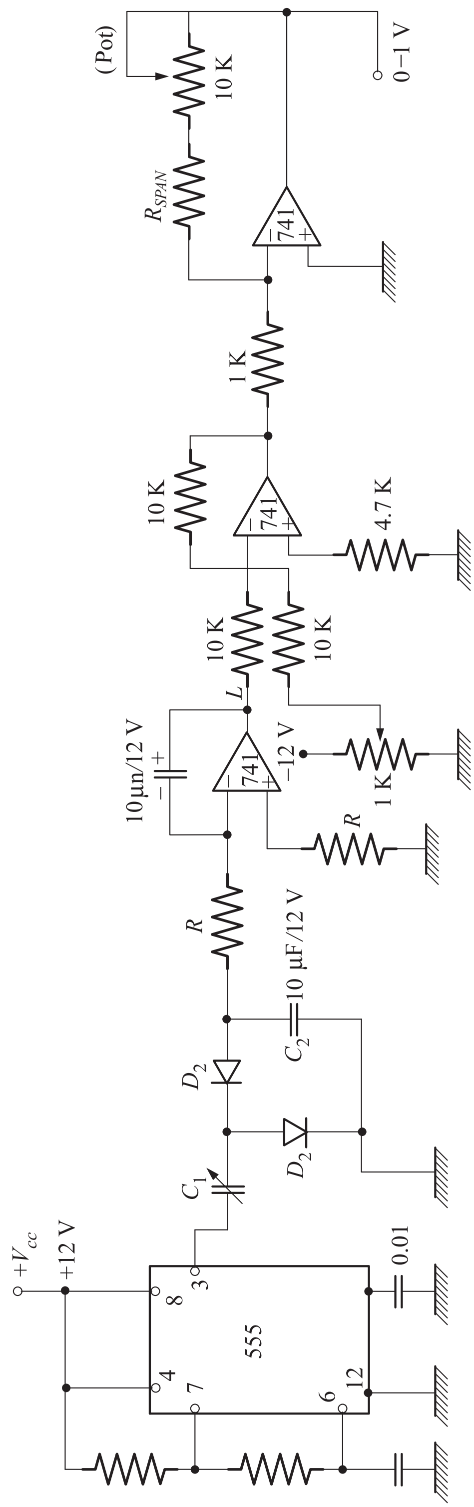


Figure 8.72 Diode pump circuit to transform capacitance or frequency variations into voltage

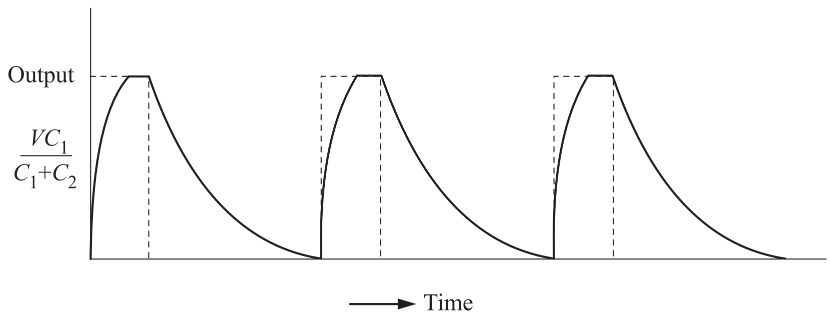


Figure 8.73 Output waveforms of a diode pump circuit

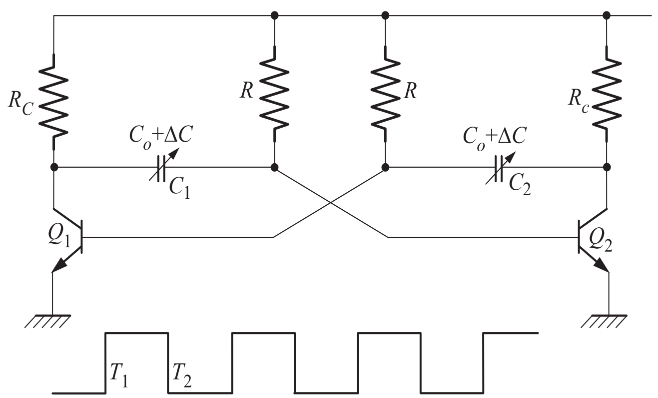


Figure 8.74 Circuit to transform differential capacitor variation into pulsewidth

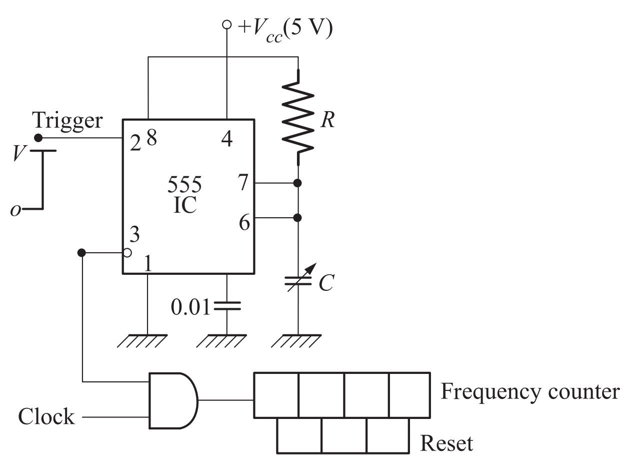


Figure 8.75 IC circuit to generate pulses for capacitance variation

phosphate (ADP), and ceramics made with barium titanate, dipotassium tartrate, potassium dihydrogen phosphate, and lithium sulphate.

Quartz (SiO_2): is mechanically and thermally the most stable among the piezoelectric materials. Its internal electric losses are small, and its volume resistivity is higher than 10^{14} Ωcm. Because of its relatively small piezoelectric effects, its application is restricted to such uses where high tensile strength, high mechanical stability, or operation at elevated temperature is essential (the material can be operated safely up to 550°C). Quartz plates are applicable primarily for thickness expansion and transverse compression operation.

Rochelle salt ($KNaC_4H_4O_6.4H_2O$): is stable at room temperature between 35 and 85% RH. At higher humidity it deliquesces, and at lower humidity it gets dehydrated. Coating with wax is recommended to restore the humidity effects. At a temperature of about 55°C the crystal will disintegrate.

The dielectric constant in the direction of the *Y*-axis of the crystal varies within wide limits with temperature and also the resonance frequency of the crystal changes with temperature. The *X*-cut of the Rochelle salt crystal is primarily for face shear motion. Crystals can also be cut for transverse (longitudinal) expansion and the cuts can be combined to form bender or twister bimorph systems. The volume resistivity is of the order of 10^{12} Ωcm or more. The mechanical strength is considerably lower than that of quartz.

Ammonium dihydrogen phosphate ADP-($NH_4H_2PO_4$): has a higher thermal stability than Rochelle salt. It is stable up to a temperature of 180°C, but for technical reasons, is used with an upper limit of about 125°C. It operates satisfactorily in a range from 0 to 44% RH, if condensation of water on surface (causing surface leakage) is avoided. The volume resistivity is considerably lower than that of the other substances described above. It is about 10^{10} Ω-cm and decreases with increasing temperature and content of impurities.

The material (primarily the *Z*-cut of crystal) is applicable for face shear and longitudinal expansion operation and for twister bimorphs. The sensitivity is high. For a load of 1 dyne/cm^2, a 1 cm thick crystal gives an output voltage of 1.78×10^{-4} V.

Barium titanate ceramics ($BaTiO_3$): The single barium titanate crystal is ferroelectric, that is, if exposed to an electric field it becomes polarised and maintains its polarisation even after the field is removed. Application of the field in the reverse direction causes a reverse polarisation. The process follows a hysteresis curve.

In the ceramic polycrystalline form, the individual crystals normally are randomly oriented. By the application of a polarised electric field (of the order of 1000 to several 1000 V/mm for about 1 hr at room temperature), the ferroelectric domains are lined up and the materials become piezoelectric. The materials lose some of the piezoelectric activity in the first few days and remain fairly constant. The aging process can be accelerated by subjecting the material to a higher temperature (80°C). The addition of about 4% lead titanate extends the long time stability.

Barium titanate is applicable for thickness expansion and for transverse expansion. They can be combined to form bender bimorphs but they do not operate in shear modes and consequently, they cannot be combined to form twister bimorphs.

8.20.2 Equivalent Circuit

Consider Fig. 8.76. The charge generated by the crystal can be expressed as $Q = K_q K_i$, where K_q is in Coulombs/cm and K_I is in cm. The resistances and capacitances can be converted into a current generator:

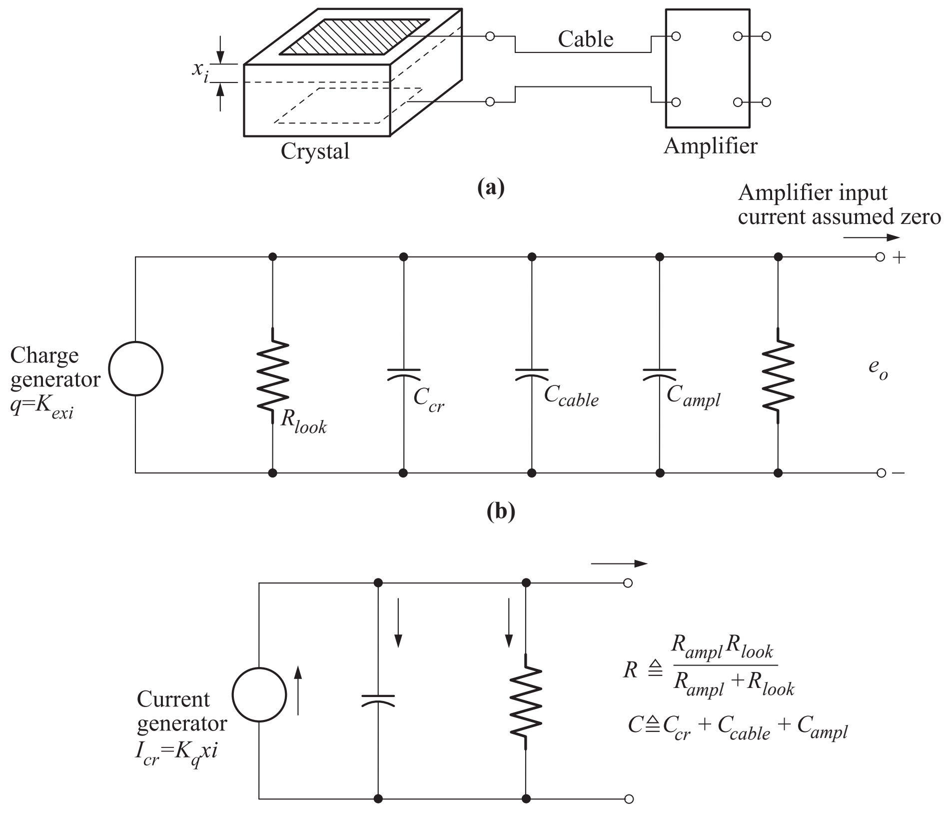


Figure 8.76 (a) Piezoelectric transducer (b) equivalent circuit and (c) simplified equivalent circuit

$$i_{cr} = \frac{dq}{dt} = K_q \frac{dx_i}{dt}$$

Then

$$i_{cr} = i_C + i_R$$

i.e.,

$$k \frac{dx_i}{qdt} = C\frac{de_o}{dt} + \frac{e_o}{R}$$

Taking Laplace transforms on both sides,

$$K_q S_{X1}(S) = CSE_o(S) + \frac{E_o(S)}{R}$$

Then

$$\frac{E_o(S)}{X_1(S)} = \frac{k\tau D}{\tau D+1}; \quad \left[D = \frac{d}{dt}\right]$$

where $k = k_q /C$ is the sensitivity in V/cm and $T = RC$ the time constant in sec.

This has a steady-state response to a constant, X_i is zero just like the capacitance pickup. Therefore, static displacements cannot be measured. For a flat amplitude response within, say 5%, the frequency must exceed ω, where

$$(0.95)^2 = \frac{(\omega\tau)^2}{(\omega\tau)^2 + 1}$$

and

$$\omega_1 = \frac{3.02}{T}$$

To get a better low-frequency response, the value of T should be larger. We can also show that larger the value of T, better will be the pulse response and time response of this device.

8.20.3 Piezoelectric Coefficients

A brief discussion of the definitions and relations of the coefficients used to describe piezoelectric properties is appropriate. Double-subscripted notation is used to represent the constant. The convention is that the first subscript refers to the direction of the electrical effect and the second to that of the mechanical effect using the axis numbering systems in Fig. 8.77.

(a) *d*-Coefficient: This is the fundamental piezoelectric coefficient. It is related to the direction of the applied force with respect to the crystals axes, indicated by two subscripts, which are derived from the representation of stress and strain in crystals. Because of crystal symmetry in quartz, only the following coefficients are significant:

$$d_{11} = 2.26 \times 10^{11} \text{ Coulombs/kg}$$
$$= 2.30 \times 10^{-12} \text{ Coulombs/Newton}$$
$$d_{14} = -0.655 \times 10^{-11} \text{ Coulombs/kg}$$
$$= -0.670 \times 10^{-12} \text{ Coulombs/Newton}$$

also,

$$d_{12} = -d_{11}$$
$$d_{25} = -d_{14}$$
$$d_{26} = -2d_{11}$$

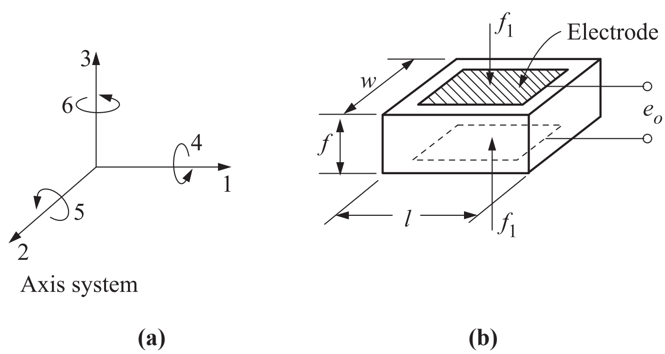


Figure 8.77 Piezoelectric transducer

The d-coefficient gives the charge output per unit force input (or charge density per unit pressure) under short circuit conditions, that is, if the charge is being led away on developing so that it has no second-order effect on the crystal deformation. In the MKS system it is measured in

$$\frac{\text{Coulombs}}{\text{Newton}} = \frac{\text{Coulombs/m}^2}{\text{Newton/m}^2} \quad \text{(direct effect)}$$

In a system of dimensions in which the mechanical and electrical unit of energy is the same, such as the MKS system, the same numerical coefficient serves the direct and the inverse piezoelectric effect. In the latter case the d-coefficient measures the deformation obtained per unit applied voltage under no-load conditions, that is,

$$\frac{m}{V} = \frac{m/m}{v/m} \quad \text{(inverse effect)}$$

(b) *g*-Coefficient: Dividing the d-coefficient by the absolute dielectric constant,

$$\varepsilon\varepsilon_o = \varepsilon \times 8.85 \times 10^{-12} \quad \text{(MKS)}$$

which yields the g-coefficient representing the generated *emf* gradient per unit pressure input. It is a most convenient coefficient for computing the output voltage of piezoelectric transducers, if additional shunt capacity is not considered. Its dimension is

$$\frac{V/m}{\text{Newton/m}^2}$$

(c) *h*-Coefficient: While the d and g-coefficients are related to the applied forces, the h-coefficient n is derived from the deformation of the crystal. It is obtained by multiplying the g-coefficient by Young's modulus valid for the appropriate crystal orientation of the material, and thus measures the *emf* gradient per unit mechanical deformation or

$$\frac{V/m}{m/m}$$

(d) Coupling coefficient, *k*: The coupling coefficient can be computed by taking the square root of the product of h and d. It represents the square root of the ratio of the mechanical energy stored in the crystal to the electrical energy absorbed by the crystal, or vice versa, and is thus a measure of the efficiency of the crystal as an energy converter. It is mainly used in connection with piezoelectric generator-type transducers where the transducer is employed as an electromechanical vibrator and in crystal-type filters. In principle, a comparison of the coupling factors of different materials permits the assessment of their relative sensitivities without referring to crystal dimensions. When considering generator-type transducers, however, one must realise that they are usually loaded mechanically whilst the coupling coefficient refers to short circuit output conditions, that is, without mechanical restraint. The required correction, however, is small with materials of low coupling coefficients, such as quartz or even barium titanate, but may approach a value of 2 for Rochelle salt. In an instrument transducer the coupling coefficient is probably less important than in acoustic and ultrasonic applications.

The piezoelectric sensitivity of quartz varies with temperature. Between 20 and 200°C the temperature coefficient of d_{11} is about –0.016% per degree C. At lower temperatures, the variation is less pronounced. Above the melting point quartz loses its piezoelectric character permanently. Although dielectric constant changes very little at temperatures up to about 250°C, there is an immediate and continuously increasing drop of resistivity, amounting to about six orders between the room and Curie-point temperature. Quasi-static measurements with quartz-type transducers at high environmental temperatures are therefore impracticable.

8.20.4 Modes of Deformation

There are a number of possible modes of deformation by which a piezoelectric transducer may suffer under load. They are portrayed in Fig. 8.78.

8.20.5 Multiple Arrangements (Stacks)

The elements may be connected in series leading to a higher output voltage for the same force. If they are connected in parallel, then a lower output impedance results (Fig. 8.79).

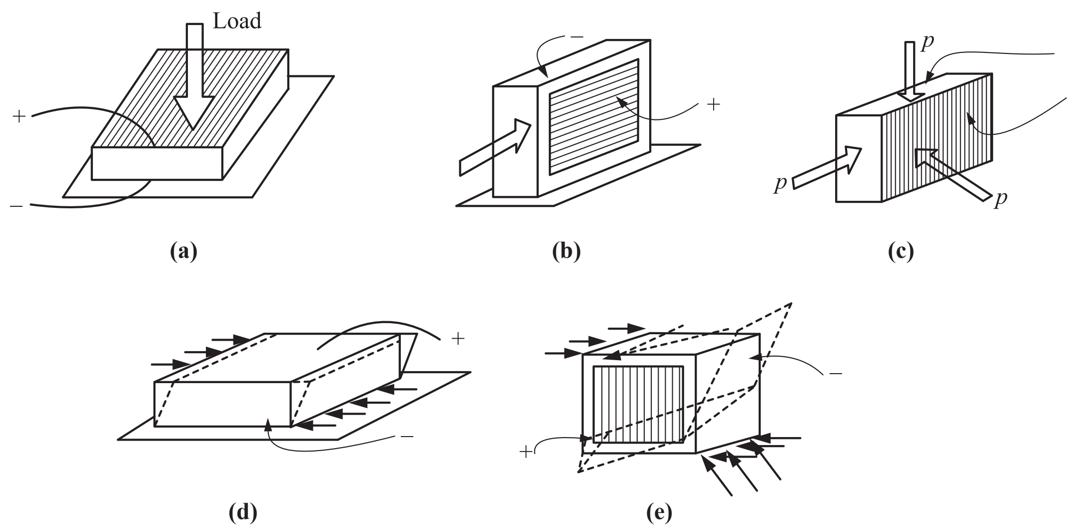


Figure 8.78 Modes of deformation in piezoelectric transducers: (a) thickness expander (TE) (b) length expander (LE) (c) volume expander (VE) (d) thickness shear (TS) and (e) face shear (FS)

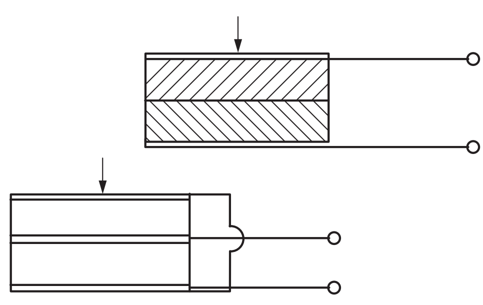


Figure 8.79 Multiple arrangements

8.20.6 Bimorphs

(Trade name of Brush Electronics Company.) Bimorph bender consists of two transverse expanding plates connected together in such a manner that one plate contracts and the other expands when a voltage is applied. If the element is free to move, then it will bend as shown (exaggerated) in Fig. 8.80.

This bimorph element has got a higher sensitivity and permits larger deflection than a single solid one (Fig. 8.81).

8.20.7 Bimorph Twisters

Two face shear plates are cemented together to have a series connection so that their expanding diagonals are perpendicular. If a voltage is applied and if both plates are free to move then it will bend as shown in Fig. 8.82.

8.20.8 General Form of Piezoelectric Transducers

A typical construction of a piezoelectric accelerometer is as shown in Fig. 8.83. The crystal is pre-loaded to about 1000 psi stress by screwing down the cap on the hemispherical spring. The pre-stressing puts the piezoelectric material at a more linear part of its stress charge curve. It allows measurement of acceleration in both directions without the crystal going into tension. When the pre-load is applied, a voltage of certain polarity is developed but this soon leaks off to zero if any further load due to acceleration gives a plus or minus charge depending on the direction of the motion.

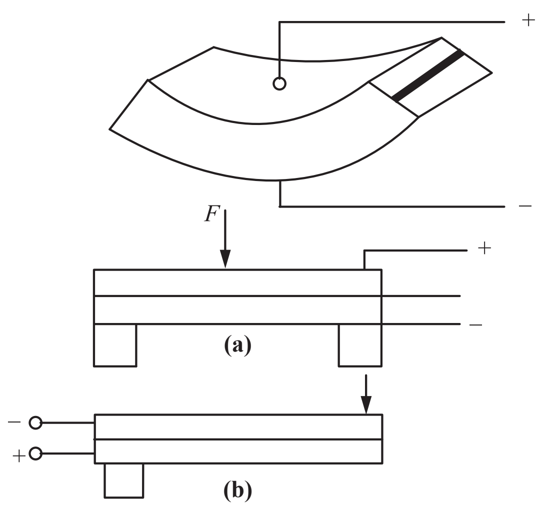
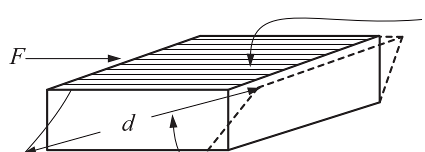


Figure 8.80 Biomorph benders



d—Expanding diagonal

Figure 8.81 Biomorph element

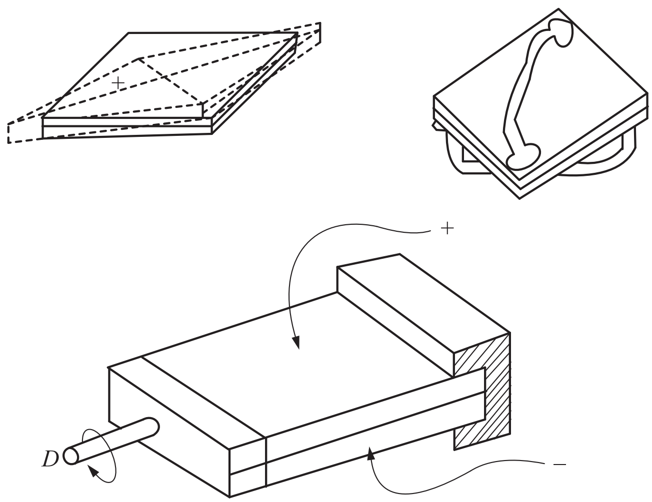


Figure 8.82 Bimorph twisters

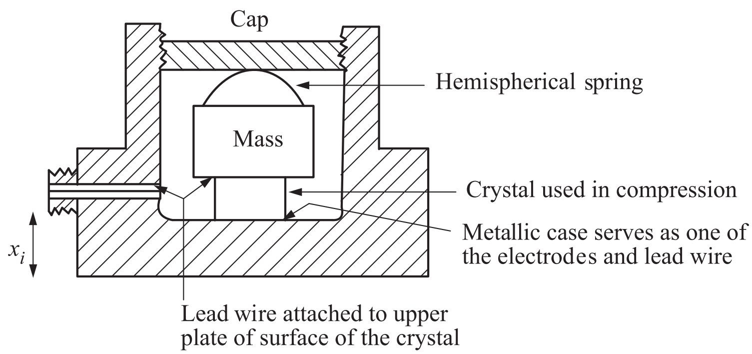


Figure 8.83 Piezoelectric transducer arrangement

8.20.9 Available Typical Piezoelectric Accelerometers

A single low –*g* instrument has a sensitivity of 50 mV/g and will measure acceleration from 03 to 1000*g* with a non-linearity of 1% of full scale. It has a natural frequency of 20 kHz and frequency response flat ±5% from 20 to 4000 when used with 100 MΩ input impedance, and a capacitance of 600 pF with a 1 m cable (Fig. 8.84).

Cross-axis sensitivity is 5%.
Sensitivity drift is ±10% from –30 to +230°.
Size 1 in cube weight 2 oz.

8.20.10 Shock Accelerometer

1. Sensitivity: 5 V/g.
2. Range: 0–1000 g with 1% non-linearity.
3. Natural frequency: 35 kHz.
4. Frequency response: flat with ±5% from 0. 1 to 7000 Hz with a charger amplifier.

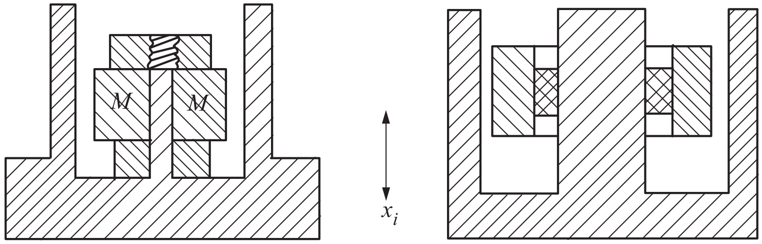


Figure 8.84 Accelerometers reduce cross-sensitivity

5. Pulse response ±5% for pulses shorter than 0. 66 sec; no ringing for pulses longer then 0.15 msec.
6. Capacitance: 100 pF.
7. Cross-axis sensitivity: 5%.
8. Sensitivity drift: ±1% from –100 to 350 °C.
9. Size 7 in cube weight 1 oz.

8.20.11 Environmental Effects

8.20.11.1 Effect of temperature. The temperature dependence of a piezoelectric material is far too complex. Very little is known about the temperature dependence of charge sensitivity though it is known from experience that it roughly follows the tendency of permittivity/temperature characteristics.

8.20.11.2 Effect of humidity. Humidity may affect piezoelectric materials in two ways.

1. Water-soluble crystals, if not protected, can stand only a limited range of environmental humidity. Air of lower humidity extracts the water of crystallization and makes it amorphous.
2. All piezoelectric materials suffer from loss of insulation resistance when water vapour condenses on their faces.

Therefore, a water proof coating is a must. Silicone grease is filled into cavities of transducers and connectors.

8.20.11.3 Transverse sensitivity. If a linear acceleration transducer is exposed to a sinusoidal linear acceleration in a radial direction perpendicular to the main axis (direction of polarization), the sensitivity assumes a two-lobed shape as shown schematically in Figs. 8.85 and 8.86.

8.20.11.4 Pressure sensitivity. This problem is encountered in piezoelectric acceleration transducers due to airborne high-intensity noise. Materials with a VE mode can produce a signal when exposed to sound pressure. However, if a transducer is designed to prevent admittance of sealed housing, then net pressure sensitivity may not be experienced.

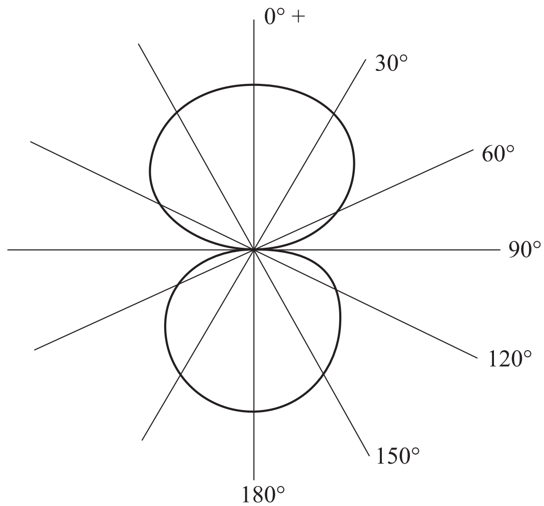


Figure 8.85 Two-lobed shape of sensitivity variation

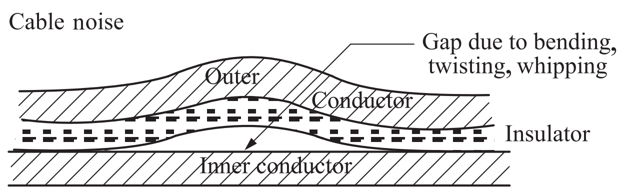


Figure 8.86 Sensitivity variation of acceleration transducer

8.21 MAGNETOSTRICTIVE TRANSDUCERS

Ferromagnetic materials like iron, nickel, 68 permalloy, etc. change their magnetic permeability under mechanical stress. This is known as the Villari effect.

The permeability can increase or decrease depending upon the material, the type of stress (compression, tension, or torsion), and the magnetic flux density in the sample. The ferromagnetic elements that have the above property are known as magnetostrictive or magnetoelastic materials. This property can be made use of in constructing a transducer to convert a stress to a variation in induction (Fig. 8.87).

8.21.1 Magnetostrictive Materials

The magnetostrictive materials are as follows:

1. Nickel.
2. Permalloy—nickel alloy with 68% nickel.
3. Ferroxcube B—highly brittle; therefore, not used much.

The BH curves of nickel and permalloy are shown in Fig. 8.88. We can have transducers to measure pressure or load-vibrating stress-acceleration.

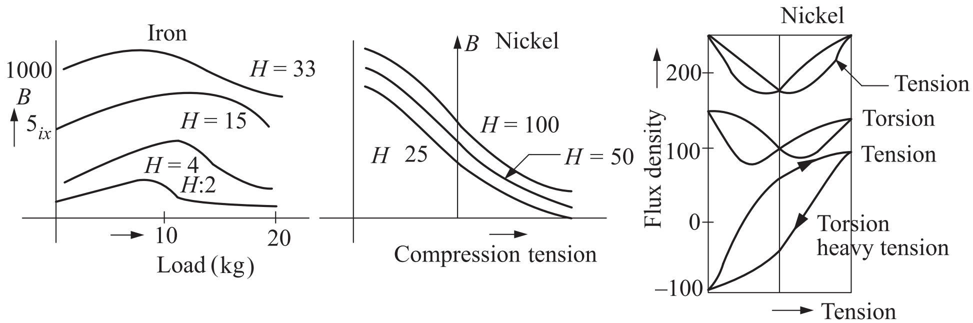


Figure 8.87 BH curves

8.21.2 Practical Forms of Transducers

Depending upon the cross-sectional area of the probe, forces from several grams up to several tons can be measured directly. The displacement at the input of the transducers is very small of the order of several microcms (microns).

The yoke *Y* surrounding the magnetoelastic probe *P* (Fig. 8.89) increases the sensitivity by increasing the magnetic flux and acts as a shield against external stray magnetic fields. The yoke as well as the core should be laminated to reduce eddy currents. A magnetostrictive phonograph pickup is illustrated in Fig. 8.90. Two pickup coils are wound on the two halves of the torsion wire. Hysteresis is reduced and linear response is obtained by giving the wire an initial twist.

A stylus displacement will decrease the stress on one side and increases it on the other side by an equal and opposite amount. This increases the magnetic flux in one half of the wire and decreases it in the other half. The frequency response is flat over 150–15,000 Hz.

The highest frequency with which the transducer will respond is determined by the mechanical resonance frequency of the magnetoelastic probe. This can be as high as several 10,000 Hzs. Eddy currents are likely to introduce limitation at a frequency higher than 5000 Hz.

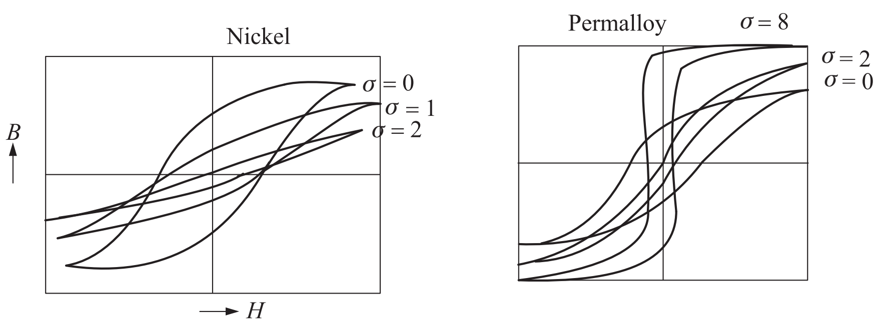


Figure 8. 88 BH curves of nickel and permalloy

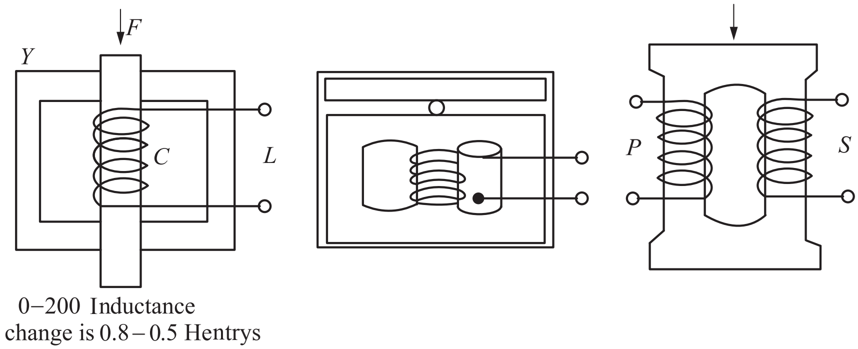


Figure 8.89 Magnetoelastic probe

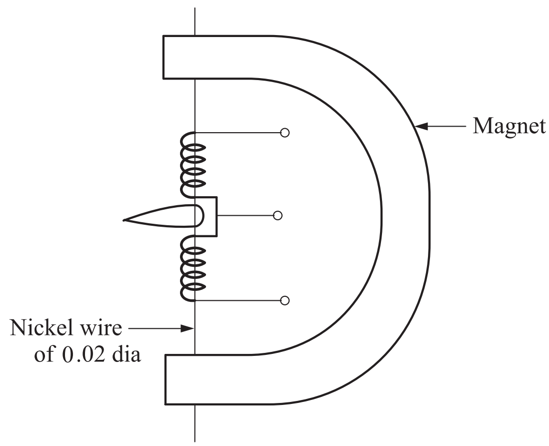


Figure 8.90 Magnetostrictive phonograph pickup

8.21.3 Magnetostrictive Torque Transducers

Thin-wall tubes of iron and nickel are magnetised in a direction bisecting the principal planes of stress (either axially or tangentially) with zero-applied torque. With zero-applied torque, no voltage is present in the coil V_s —application of torque generates tensile and compressive stresses of equal magnitude with the ion shaft as shown in Fig. 8.91.

If the material is magnetostrictive, this causes a rotation of M towards the direction of tensile stress in which magnetostriction is positive. Now the component $M \sin \delta$, a voltage in the coil V_s, is proportional to the torque.

8.21.4 Errors

Hysteresis error can be large, and it can be reduced by aging, pre-stressing, and using the transducer over small ranges of stress variation.

8.21.5 Temperature

1. Increase in temperature causes an increase in inductance at zero stress.
2. Increase in temperature changes the stress sensitivity of the transducer by about 0.2% for a variation of 1°C.

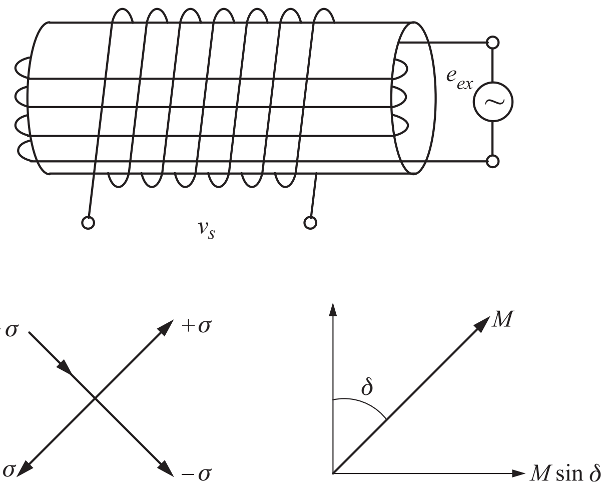


Figure 8.91 Generation of stresses on application of torque

8.21.6 Variation of Supply Voltage

This causes relatively small variation because inductance variation depends upon the magnetic field strength in the probe. This increases first with increased field strength and then decreases beyond a certain operating point. Therefore, there exists an operating point where small supply-voltage variation has no influence on the transducer output.

8.21.7 Eddy Currents

The effects of eddy currents can be reduced by choosing the core material with a high electrical resistivity.

8.21.8 Input Impedance

The mechanical input impedance can be made large by choosing the core with a high tensile strength. This type of transducer can be built to measure large forces up to several tons and for fast transient phenomena where frequency is of the order of several 1000 cycles/sec.

The accelerometers can be built to measure several thousand '*g*' (gravity). These can be very rugged also. These instruments need individual calibration as their characteristics depend upon temperature. The characteristics and their calibration should be checked periodically as they can be changed by environmental influences. The other important transducers are hall effect transducers, ionisation transducers, elastic transducers, and digital transducers.

8.22 LIQUID-LEVEL MEASUREMENT

Usually in industries vast quantities of liquids such as water, solvents, chemicals, etc. are used in a number of industrial processes. Liquid-level measurements are made to ascertain the quantity of liquid held in a container or vessel. The liquid level affects both pressure and rate of flow in a container and therefore its measurement and/or control becomes quite important in a variety of processes encountered in modern manufacturing plants. Liquid-level measurements can be broadly classified as:

1. Direct methods.
2. Indirect methods.

8.22.1 Direct Methods

Some methods by which the actual liquid level is directly indicated are by means of a simple mechanical type of device. Here we discuss a float-operated scheme with electrical output. In this, the float acts as a primary transducer that converts liquid-level variation into a suitable displacement. This displacement is sensed by the secondary transducer such as resistance type of potentiometric device, inductive type of LVDT, etc. Figure 8.92 shows the schematic of the float-actuated rheostatic (resistive) device. The float displacement actuates the arm that causes the slider to move over the resisitive element of a rheostat. The circuit resistance changes and this resistance change is directly proportional to the liquid level in the tank.

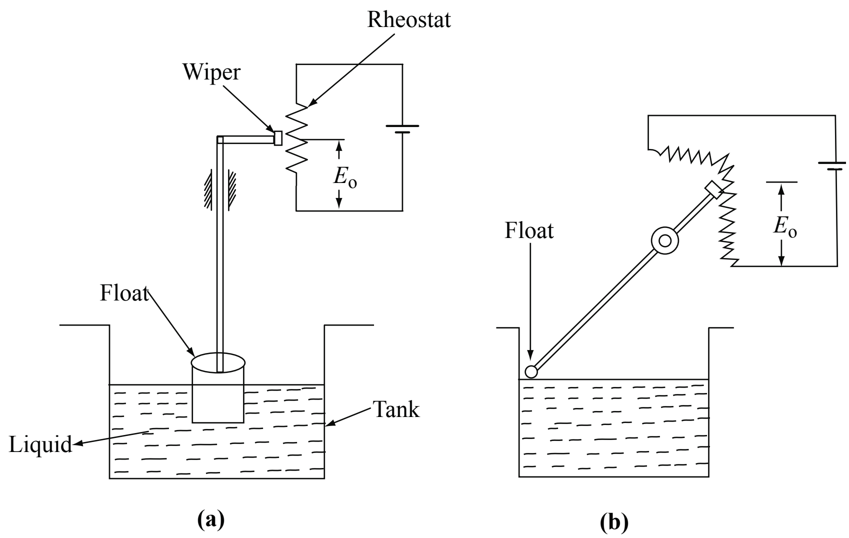
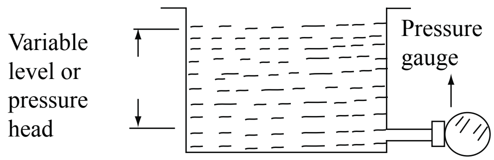


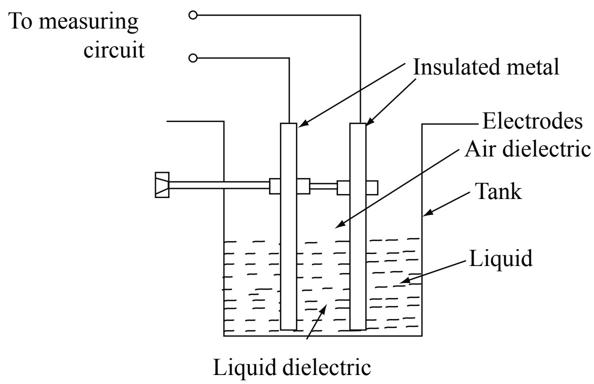
Figure 8.92 Typical float-operated rheostatic liquid level gauge: (a) linear potentiometer liquid-level gauge and (b) rotary potentiometer liquid-level gauge

8.22.2 Indirect Liquid-Level Measurement

The hydrostatic pressure created by a liquid is directly related to the height of the liquid column ($p = \rho gh$). Therefore, a pressure gauge is installed at the bottom or on the side of a tank containing the liquid (Fig. 8.93(a) and (b)). The rise and fall of the liquid level causes a corresponding increase or decrease in the pressure, which is directly proportional to the liquid level h. These gauges function smoothly when the liquids are clean and non-corrosive.



(a)



(b)

Figure 8.93 (a) Hydrostatic pressure-type-level measuring device and (b) dielectric liquid-level gauge

8.22.3 Capacitance-Level Gauge

A simple capacitor consists of two electrode plates separated by a small thickness of insulator (which can be solid, liquid, gas, or vacuum) called the dielectric. The change in the liquid level causes a variation in the dielectric between the two plates, which in turn causes a corresponding change in the value of the capacitance of the condenser. Therefore such a gauge is also termed a *dielectric level gauge*.

The magnitude of capacitance depends on the nature of the dielectric, and varies directly with the area of the plate and inversely with the distance between them. The capacitance can be changed by any of these factors. The capacitance is given as C (in μf) = $0.0885k/d$ (k = 1 for air). The capacitance would be at a minimum when the liquid fills the entire space between the electrodes.

The change in capacitance can be measured by a suitable measuring unit such as a capacitive Wheatstone bridge by either manual null balancing or automatic null balancing using the null detecting circuit with a servomotor that indicates the level reading.

For the measurement of level in the case of non-conducting liquids, the four-probe arrangement may be satisfactory since the liquid resistance is sufficiently high. For conducting liquids, the probe plates are insulated using a thin coating of glass or plastic.

The capacitance-type level gauge is relatively inexpensive, versatile, reliable, and requires minimal maintenance. These units have no moving parts, are easy to install, and are adaptable to large and small vessels. Further, such devices have a good range of liquid-level measurement, i.e., from a few cms to more than 100 m. These gauges also find wide use in other important applications such as determining the level of powdered or granule solids, liquid metals (high temperatures), liquefied gases (low temperature), corrosive materials (such as hydrofluoric acid) and in very high-pressure industrial processes.

8.23 ULTRASONIC-LEVEL GAUGE

Sound waves are directed towards the free surface of the liquid under test from an ultrasound transmitter. These waves get reflected from the surface of the liquid and are received by the receiver. In this technique, liquid-level variations are quite accurately determined by detecting the total time taken by the wave to travel to the liquid surface and then back to the receiver. The longer the time interval, the farther away is the liquid surface, which in turn is a measure/indication of the liquid level.

It may be noted that the operating principle of this instrument is quite simple. However, the actual instrument is expensive and requires a high degree of experience and skill in operation. However, its main advantage is a wide range of applications in level measurement for different types of liquid and solid substances.

8.24 MEASUREMENT OF HUMIDITY AND MOISTURE

The amount of water vapour content in the atmosphere, i.e., humidity, is an important process, which is variable in a number of industrial processes. This is because humidity/moisture content affects the behaviour of many commercial materials such as paper, textiles, paint, tobacco, soap powder, paints, fertilisers, leather, wood products, etc. Humidity measurement and control are necessary during many industrial processes as well as in heating and airconditioning systems. For example, environmental heat and cold can be endured without much discomfort if the humidity of air is carefully controlled so that it establishes human comfort conditions.

Humidity of the atmosphere represents the measure of water vapour present in the air. Alternatively, it could be considered as the degree of dampness of the air. Humidity is generally expressed in terms of '*absolute humidity*' or '*relative humidity*'. The absolute humidity of a gas is defined as the mass of water vapour present in a unit volume of gas and is usually expressed in *grams per cubic meter*. The other term, relative humidity (RH), compares the humidity of air with the humidity of saturated air at the same temperature and pressure.

RH is defined as the ratio of the mass of water vapour present in a given volume of gas to the mass of water vapour necessary to saturate the same volume of gas at the same temperature.

8.25 PHOTOCONDUCTIVE CELLS

The photoconductive or photoresistive cell and the photodiode are variable resistance transducers. These cells are much smaller, more sensitive, and more rugged than any of the photoemissive types.

The photoconductive cell varies its electrical conductivity in accordance with the light intensity it receives. It consists of a thin coating of selenium, germanium, lead sulphide, or other metallic oxides between two electrodes on a glass plate and may be used in an electrical circuit as a variable resistance. To produce a voltage output, an external source of power is used. In strong light, electrical resistance is reduced in the order of 5:1 for selenium and 80:1 for lead sulphide as compared with their resistance in the dark. Photoconductivity can be produced by light of longer wavelength (near infrared) than required for photoemission.

Other photoconductive transducers are the photodiode and the phototransistor. These semiconductor devices operate on the principle that light gives enough energy to the valence-band electrons to raise them to the conduction band. This action increases the supply electrons and holes, which act as current carriers and decreases resistivity. Efficiency decreases rapidly in the infrared region beyond 2μ and beyond blue or ultraviolet. The phototransistor is quite small in size and is adaptable in applications where a large number of individual, light-sensitive elements are required in a small area.

Figure 8.94(a) shows a simple resistance bridge with a photocell in one arm and the transistor in a position so as to amplify any unbalance that occurs when resistance of the photocell changes because of illumination. The slider is initially adjusted to balance the bridge when the photo cell is in the dark. Then the transistor is in cutoff modes, and no current flows in the circuit except the bleeder current, which is about 1.0 mA. When the cell is illuminated upto 1 mA or more, collector current will flow depending upon the DC gain of the transistor being used. This circuit has a current sensitivity of approximately 1.0 A per lumen. The circuit in Fig. 8.94 is an improved version of that in Fig. 8.94(a), which increases the sensitivity by taking the advantage of the complementary nature of NPN and PNP transistors used in the bridge circuit. Current from the NPN transistor flows through the base to the emitter of the PNP transistor and is amplified by the current gain of this second transistor. The resulting sensitivity is 10 A per lumen:

1. The clearance between the pickup and the actuating medium.
2. The rate of movement of the actuating medium.
3. The size of the actuating medium.

8.26 PHOTO PULSE PICKUP

A photo tube may be used to detect rotation and other motions either by interruption of the light beam or by the detection of translucent and opaque portions of moving components. The methods used include sensing the reflection of a white spot on a rotating shaft (Fig. 8.95) and sensing light beam interruptions caused by a perforated wheel. The series of discrete pulses thus produced are counted over a precise time internal. From this the shaft velocity can be estimated.

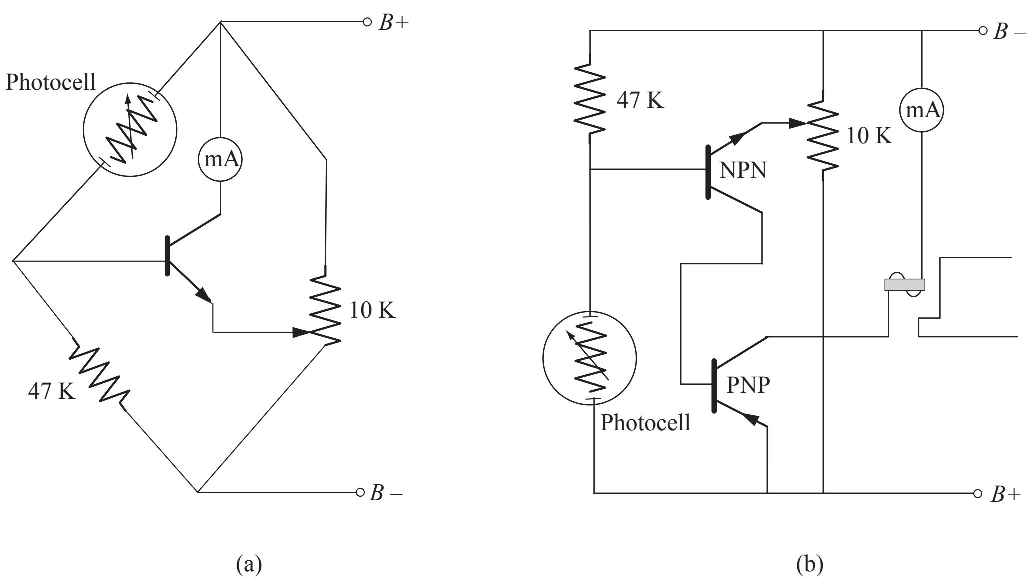


Figure 8.94 Photocell circuits

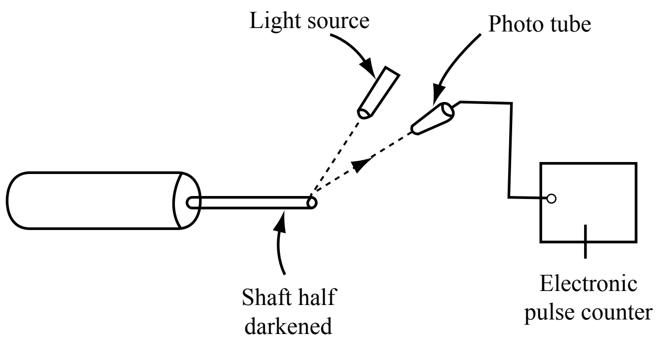


Figure 8.95 Photo pulse pickup

8.27 DIGITAL ENCODERS AND ENCODER TRANSDUCERS

Digital encoders and encoder transducers produce a distant coded output signal in response to physical variable changes. As the output of these devices is a true digital output, it is suitable for direct entry into digital computers or data-handling systems without further conversion. Digital encoders may be read out continuously or may have to be stopped momentarily for readout, depending upon the design and the code used. They may also be used as part of other transducers, especially when small in size.

8.27.1 Shaft Position Encoders

The shaft position encoder was the first of all digital transducers and is still the most efficient. It consists of a shaft attached to a disk mask or a drum with a digital coded scale. The scale may be formed either with a combination of conducting and non-conducting areas or a combination of translucent and opaque surfaces. There exists a definite coded form for each discrete position of rotation of the disk. Resolutions are from 100 points to 50,000 (with 25 cm diameter disk) unique positions per 360° turn. They are read out by a series of brushes or photocell arrangement. Each brush or photocell produces an input to a separate channel. The sequence and order of positive indications represent the shaft position in coded form. They are manufactured in a variety of sizes from 5 cm to over 25 cm diameter.

Rectilinear encoders use the same principles as the rotary type except that the scale is made over a linear motion. The typical advantage of the linear encoder is that they avoid the precision gearing, which is necessary to measure linear motion with the more common shaft position encoders.

8.27.2 Encoder Transducers

Some physical parameters may be directly digitalised into a coded pulse output utilising the rotary and linear encoders. For instance, the Bourdon tube has been used to position a shaft position directly by means of suitable linkages in a digital pressure gauge. Encoders are also used for wind direction measurement. Mechanical displacement or the extension of pressure-actuated bellows may be similarly used as a direct coded digital device.

They are not adequate when small motions or low energy input parameters are to be measured. In such cases it is possible to construct a shaft position servosystem that will respond to the output of a conventional analog transducer. Here, a voltage is taken from the analog transducer and fed to the input of the self-balancing mechanism, a coded digital output signal is provided. Many forms of this combination exist ranging from the standard 50 cm by 60 cm null balance recorder or indicator to a miniaturised version 5 cm diameter and 20 cm long. Wholly electronic analog to digital converters, coupled with an analog transducer and mounted in the same case are now possible with the use of advanced semi-conductor techniques.

8.28 FIBRE OPTIC DISPLACEMENT TRANSDUCER

This type of transducer can be used to measure displacement in the range of microcentimeters to tenths of centimeters. The probe (Fig. 8.96), 0.05 cm–0.75 cm in diameter and 8 cm long, consists of a bundle of several hundred optical fibres, each a few thousands of cm in diameter. The fibre bundle extends for 30 to 90 cm to the electronic chassis where it is divided into two equal groups of fibres. One group (transmitting fibres) is exposed to a light source and thus carries light to the probe tip, where it is emitted and reflected by the target surface. While targets of high reflectivity give greater output, even rather dull surfaces can be measured. The reflected light is picked up by the other (receiving) group of fibres, and transmitted to the photodetector whose output produces a DC voltage related to the probe target gap.

Probes of this general type display a front slope and a back slope with useful measuring ranges. At zero gap, no light can escape from the transmitting fibres and the output is zero. As the gap opens up more of the target surface is illuminated and reflection increases, giving a very sensitive and nearly linear range of measurement (front slope). As the gap increases finally the entire target is illuminated, giving a peak output. Motion beyond this point causes reduction in response, since

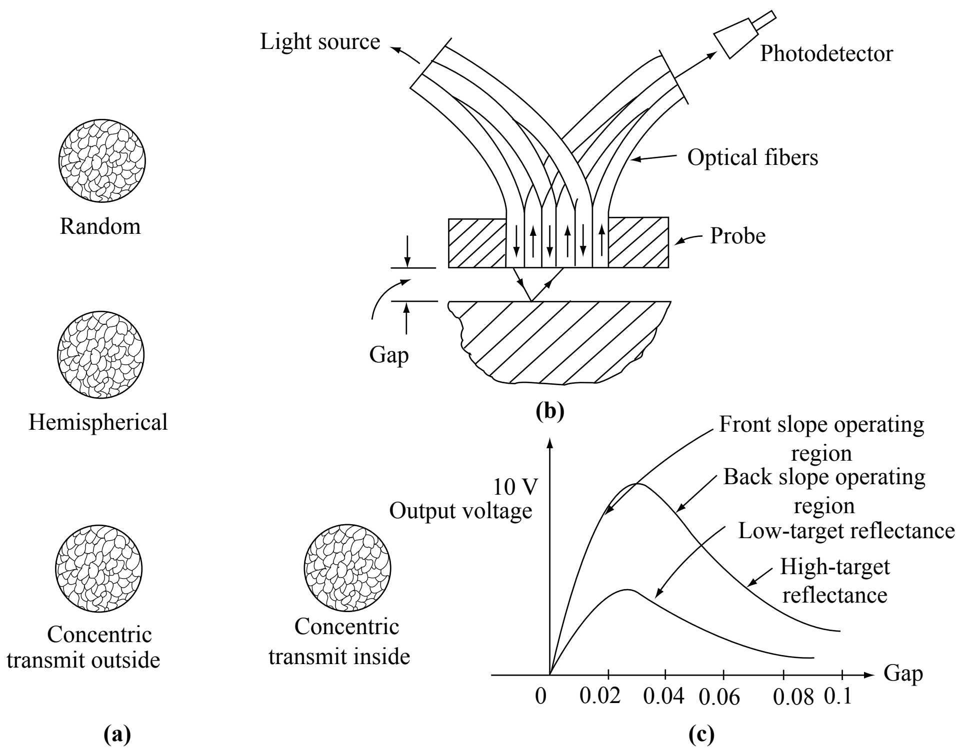


Figure 8.96 Fibre-optic transducer: (a) cross-section (b) principle and (c) characteristics

both target illumination and the fraction of reflected light gathered by the sensor decreases roughly according to an inverse square law. This back slope region is also useful for measurement, but is less linear and sensitive. Theory shows that maximum sensitivity in the front slope region is achieved by arranging the fibres in a precise geometric pattern in which each receiving fibre is surrounded by four transmitting fibres. As it is not practicable from a manufacturing point of view, a random distribution has been found to give nearly identical results. Typical characteristics for a standard (random distribution) probe of 0.109 cm diameter for the front slope region are as follows: sensitivity 1.6 mV/μ in, linear (±1%) range 0.001–0.004 cm, static resolution 1 μ, frequency response flat ±3 dB from DC to 50 kHz, full-scale output 0 – 10 V DC, output signal ripple is 50 mV p.p.(peak to peak). Since the target surface is usually an existing machine part, static calibration with that specific surface is generally required; however, automatic compensation for a slowly changing surface reflectivity is sometimes possible.

8.28.1 Absolute Motion Devices

8.28.1.1 Seismic Devices (Spring-Mass Type)

In these devices the base of the device or transducer is attached to the object whose motion is to be measured as shown in Fig. 8.97. Inside the transducer is a mass '*m*' supported on a spring of stiffness '*k*' and a viscous damper, with damping coefficient '*c*'. The motion of the mass relative to the frame or base gives an indication of the motion of the object and is the output of the instrument.

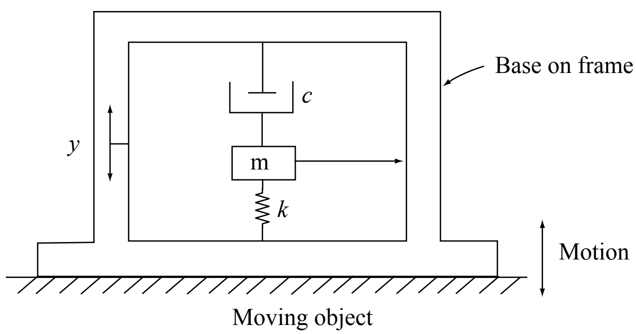


Figure 8.97 Seismic instrument

Basically, a seismic transducer consists of a mass supported on a spring and with a damper, with a relative motion transducer (Fig. 8.98) to measure motion of mass relative to the frame. In Fig. 8.98(a) and (b) relative motion transducers shown are of the potentiometric type and the strain gauge type. The relative motion induces strain in a cantilever spring, which is measured by using strain gauges R_1 and R_2 since R_1 and R_2 would have strains of opposite nature due to bending of the cantilever. The two are arranged in adjacent arms of the bridge to increase the output.

An alternative arrangement using strain gauges is shown in Fig. 8.99. In this arrangement, the longitudinal strain induced in the strip is measured. Since the strips are quite stiff in the longitudinal direction compared to the same size, the arrangement of Fig. 8.99 would have a high undamped natural frequency and thus would act as an accelerometer.

Figure 8.100 shows a seismic transducer using a piezoelectrical crystal. This is commonly used for shock and vibration measurements. The crystal is fairly stiff and held in compression by a spring.

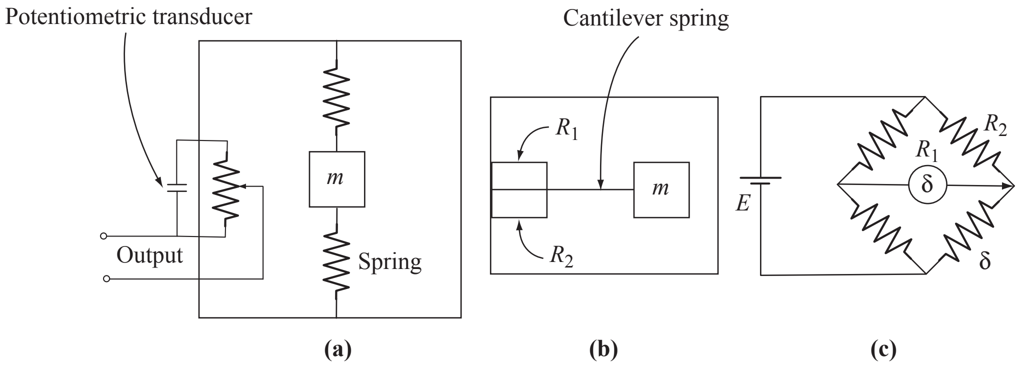


Figure 8.98 Seismic transducer using relative motion transducers: (a) equivalent circuit (b) R_1, R_2 resistance strain gauges and (c) bridge circuit

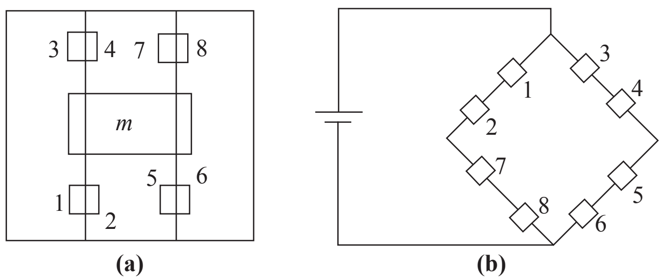


Figure 8.99 (a) Layout and (b) bridge circuit layout

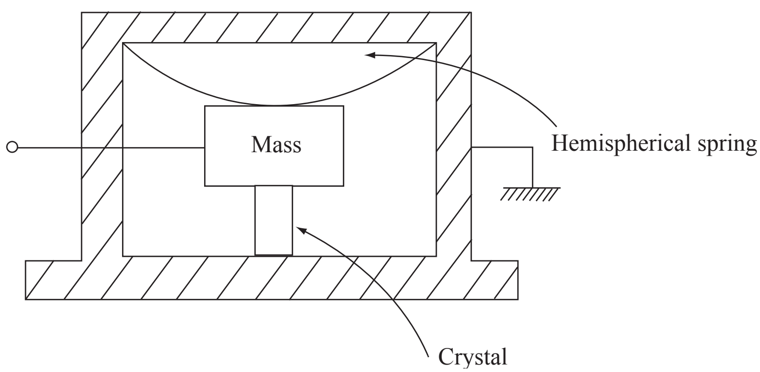


Figure 8.100 Piezoelectric accelerator

Because of the motion of the frame due to the moving object to which it is attached, an output voltage proportional to the acceleration of the moving object is obtained. Such devices are very sensitive, lightweight and can be used over a wide frequency range.

8.28.3 Measurement of Velocity

Here we consider devices for measuring the velocity of translation, along a line, of one point relative to another and the plane rotational velocity about a single axis of one line relative to another. In the first unit, two devices, variable reluctance pickup and photo tube pulse pickup, were described for the measurement of speed (angular velocity) of a shaft.

8.28.4 Translational Velocity Transducer Moving Coil Pickup

Moving coil pickup (Fig. 8.101) is based on the law of induced voltage $e_o = (B\ l\ v_l)10–8$. Where e_o = terminal voltage, B = flux density, l = length of coil (cm) and v_l = relative velocity of coil and magnet, in cm/s. Since B and l are constant, the output voltage follows the input velocity linearly and reverses polarity when the velocity changes sign. Since the flux density available from permanent magnets is limited to the order of 10,000 G, an increase in sensitivity can be achieved only by an increase in the length of wire in the coil. To keep the coil small, this requires fine wire and thus high resistance. High-resistance coils require a high-resistance voltage measuring device at e_o to prevent loading. A typical pickup of 500 Ω resistance has a sensitivity of 0.15 V/(cm/s) and a full-scale displacement of 0.15 cm with a non-linearity of ±1%. A more sensitive coil used in a seismometer (instrument to measure earth shocks) has 500,000 Ω and a sensitivity of 115 V/(cm/s).

The transducer shown in Fig. 8.102 uses a permanent-magnet core moving inside a form wound with two coils connected as shown. Units are available in full-range strokes from about 0.05 to 20 cm. Sensitivity varies from about 0.5 to 0.05 V/(cm/S), and non-linearity is about 1%.

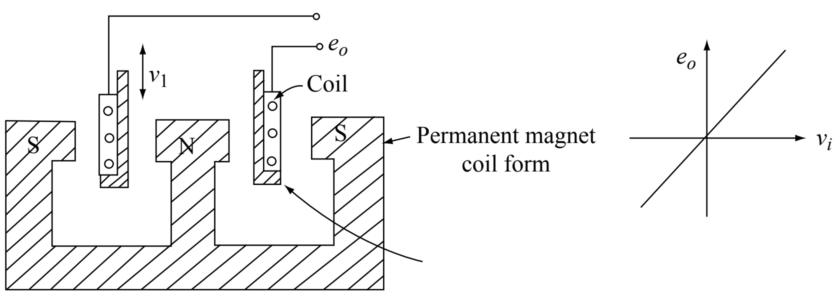


Figure 8.101 Moving coil pickup

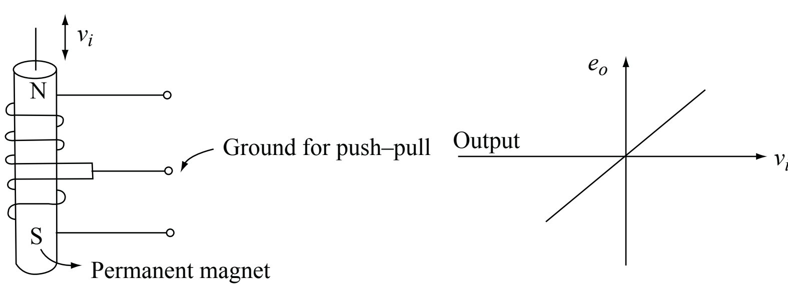


Figure 8.102 Velocity pickup

8.29 DC TACHOMETER GENERATORS FOR ROTARY VELOCITY MEASUREMENT

An ordinary DC generator produces an output voltage roughly proportional to speed. By emphasizing certain aspects of design, such a device can be made an accurate instrument for measuring speed rather than a machine for producing power. The output voltage is given as

$$e_o = \frac{n_p n_c \phi N}{60 n_{pp}} \times 10^{-8}$$

where e_o is the average output voltage, V_{np} the number of poles n_c the number of conductors in armature, ϕ the flux per pole, lines N the speed in revolutions/min and n_{pp} the number of parallel paths between positive and negative brushes. The voltage e_o reverses polarity when angular velocity reverses. A typical high accuracy unit has a sensitivity of 7 V/1000 rpm, a rated speed of 5000 rpm, non-linearity of 0.07% over a range 0 to 3, 600 rpm, ripple voltage 2% of average voltage for speeds above 100 rpm, friction torque of 0.2 in oz, rotor inertia 7*g*cm^2, output impedance of 2800 and a total weight of 3 oz.

A special DC tachometer of unique design that is used where a limited (±15°) angular travel is acceptable exhibits a very high sensitivity. A 2.5 cm diameter model gives 500 V per 1000 rpm, whereas a 7.5 cm diameter gives 30 kV per 1000 rpm. The non-linearity is ±9% for ±15° travel and the operating torque is 500*g*cm. In this generator, the permanent magnet rotates while the coil is stationary, and no commutator is needed.

8.29.1 AC Tachogenerator for Rotary Velocity Measurement

An AC two-phase squirrel-cage induction motor can be used as a tachometer by exciting one phase with its usual AC voltage and taking the voltage appearing at the second phase as output. When the rotor is stationary, the output voltage is essentially zero. Rotation in one direction causes an AC voltage at the output of the same frequency as the excitation and of an amplitude proportional to the instantaneous speed. This output voltage is in phase with the excitation. Reversal of rotation causes the same action, except that the phase of the output shifts to 180°. While squirrel-cage rotors are sometimes used, the most accurate units employ a drag-cup rotor. This does not change the basic operating characteristics.

A typical high-accuracy unit is excited by 115 V/400 Hz voltage, has a sensitivity of 2.8 V per 1000 rpm, a non-linearity of 0.05% from 0 to 3600 rpm, negligible rotor friction, a rotor inertia of 7*g*cm^2, and a total weight of 6.7 oz. Most commercial AC tachometers are designed to be used on either 60 or 400 Hz.

8.30 FORCE MEASUREMENTS

The transducers or devices used for force measurement are also called load cells. Both static and dynamic force measurements will be considered. The magnitude of the static or dynamic forces being considered may vary from a fraction of a Newton to several meganewtons and then special devices may be required, especially for extreme cases. For dynamic measurements, electromechanical transducers are often used. The following types of devices will be discussed here:

1. Balance.
2. Hydraulic load cell.
3. Elastic force devices.

8.30.1 Balance

A simple lever system shown in Fig. 8.103 called a balance has long been used as a force-measuring device. To measure the unknown force F at a distance 'L' from the pivot, a mass 'm' at a distance 'e' from the pivot is used. The system is in equilibrium when

$$FL = mgl$$

With the knowledge of other parameters, i.e., L, l, mass 'm', and gravitational constant 'g', force F can be calculated. Several versions of such balances are available for various force ranges and degree of accuracy. In this type of device, hydraulic pressure (Fig. 8.104) is used to indicate the force F applied to the diaphragm or some other type of force-transmitting element. When force F is applied, pressure is developed in the fluid, which is normally oil. This can be used upto very large forces of the order of millions of Newtons.

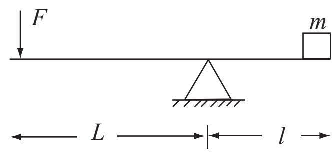


Figure 8.103 Balance principle

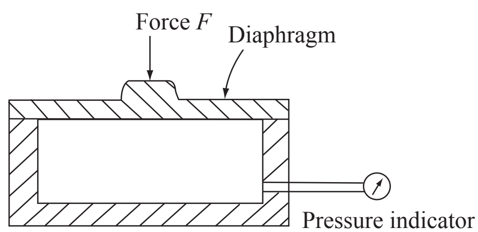


Figure 8.104 Hydraulic load cell

8.30.2 Elastic Force Devices

These are important devices for measurement of both static and dynamic forces. In such devices, the force applied to the elastic member results in a displacement or strain in the elastic member, which is sensed by mechanical or electromechanical means. The elastic members may be in the form of rings, diaphragms, strips, cylinders, etc. Relations between strain and stiffness for some of the members are given here.

1. *Axially loaded member*: For the member shown in Fig. 8.105, strain in axial direction (ε) = Strain and stiffness K in the same direction are:

$$\varepsilon = F/AE \quad \text{and} \quad K = EA/L$$

where P is the force, E Young's modulus, A the area of cross-section, L the member length, and k the stiffness.

2. *Cantilever-type elastic element*: Strain at the root of the cantilever

$$\varepsilon = 6FL/Ebt^2$$

$$\text{and stiffness, } K = \frac{\text{Force } P}{\text{Deflection at free end}} = \frac{Ebt^3}{4L^3}$$

where F is the force, E Young's modulus, A the area, L the length, K the stiffness, ε the strain, b the width, and t the thickness (Fig. 8.106).

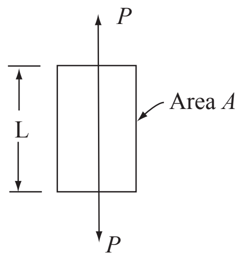


Figure 8.105 Axially loaded elastic member

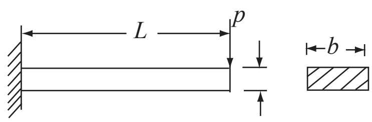


Figure 8.106 Cantilever-type elastic member

8.31 ELECTROMECHANICAL METHODS

For measuring displacement or strain in the above cases, these methods use electromechanical transducers that are discussed in earlier chapters.

A piezoelectric transducer employing a piezoelectric crystal is a useful device for dynamic force measurement. Such transducers are very sensitive and can be used for measurement of forces over a wide range of values.

Figure 8.107 shows an LVDT type of transducer for force measurement. The deflection of the elastic diaphragm due to force '*F*' is measured by a LVDT, the principle of which has already been discussed. This type of device can be used for static as well as dynamic force measurement.

Figure 8.108 illustrates the use of resistance strain gauges for force measurement. The figure shows a cantilever-type load cell, with four gauges bonded at the root such that strains in R and R_3 are opposite in nature to those in R_2 and R_4. In Fig. 8.109 the resistance gauges are bonded on the inside and the outside of an elastic ring. In Fig. 8.110 a hollow cylinder is loaded in an axial direction, inducing both longitudinal strains while R_2 and R_4 measure circumferential strains. The gauge arrangement has been shown with adjacent arms in the Wheatstone bridge having strains of opposite nature.

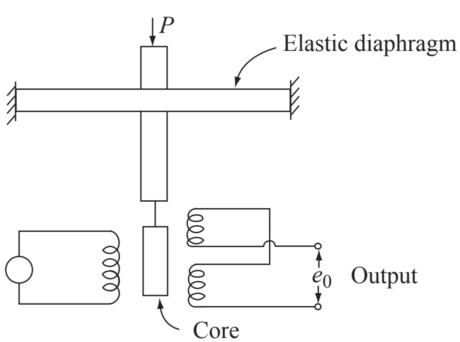


Figure 8.107 LVDT-type force transducer

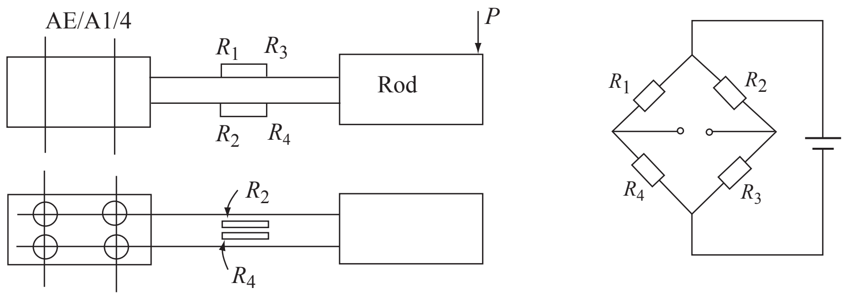


Figure 8.108 Cantilever-type load cell

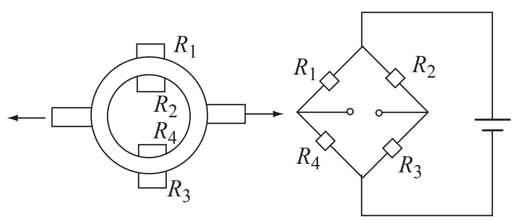


Figure 8.109 Force measurement ring-type cell

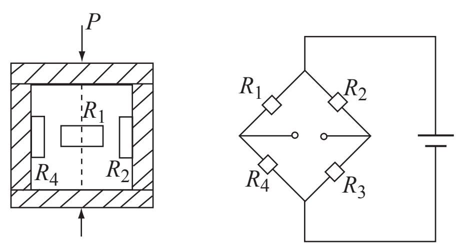


Figure 8.110 Force measurement cylindrical-type cell

8.32 MEASUREMENT OF PRESSURE

Pressure means force per unit area, exerted by a fluid on the surface of the container. Absolute pressure means the fluid pressure above the reference value of a perfect vacuum or the absolute zero pressure. Gauge pressure represents the value of pressure above the reference value of atmospheric pressure. It is the difference between the absolute and local atmospheric pressures. The atmospheric pressure at the sea level is 760 mm of Hg. Figure 8.111 shows various terms used to express pressure. Vacuum represents the amount by which the atmospheric pressure exceeds the absolute pressure. From the figure the gauge pressure corresponding to absolute pressure $P_1 = P_1 - P_a$, and vacuum corresponding to absolute pressure $P_2 = P_a - P_2$.

The techniques for pressure measurement are quite varied, depending on whether pressure is moderate, very high, or very low and also whether it is static or dynamic.

Pressures higher than 1000 atm are usually regarded as very high while those of the order of 1 mm Hg or below are regarded as very low.

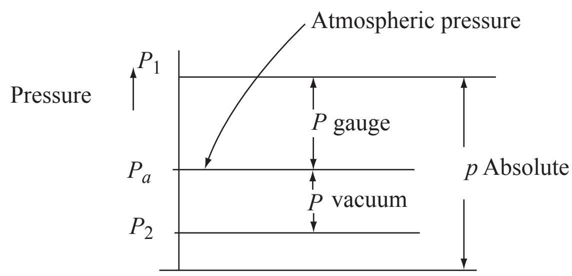


Figure 8.111 Various pressure terms used

8.32.1 Moderate Pressure Measurement

Two types of devices are included in this category of measurement, i.e., manometer and others using elastic elements. Manometers are meant for measuring *static pressure,* while devices using elastic elements may be used for both *static* and *dynamic measurements*.

8.32.2 Manometers

A manometer is a simplest device used for measuring static pressure. A simple 'U' type using water, mercury, or any other suitable fluid is shown in Fig. 8.112. The difference in levels 'h' between the two limbs is an indication of the pressure difference $P_1 - P_2$ between the two limbs:

$$h = (p_1 - P_2)/\rho$$

where ρ is the mass density of liquid in manometer and g the gravitational constant. If one of the limbs is exposed to atmospheric pressure (say limb 2) the above difference gives the gauge pressure applied to limb 1. The desirable characteristics of a manometer fluid are:

1. It should be non-corrosive and not have any chemical reaction with the fluid whose pressure is being measured.
2. It should have low viscosity and thus quick adjustment with any pressure change.
3. It should have negligible surface tension and capillary effects.

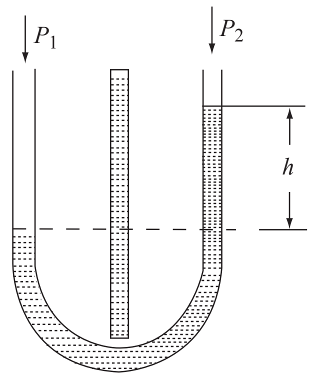


Figure 8.112 U-tube-type manometer

Several types of modified manometers are available, which have the advantages of ease in use and high sensitivity. One such device is the cistern or well-type manometer shown in Fig. 8.113.

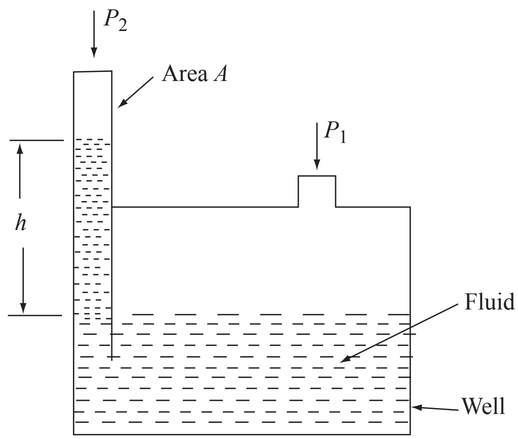


Figure 8.113 Well-tube manometer

In this type, the area is large compared to that of the tube. Thus, only a single leg reading may be noted and the change in level in the well may be ignored. Here,

$$P_l A - P_2 A = Ah\rho g, \quad \therefore \frac{P_1 - P_2}{\rho g} = h$$

where p is the density of fluid, h the height of column, and g the gravity.

An inclined-type manometer is another device, which is sensitive and convenient to use. In such a manometer, the length 'l' along the inclined tube is read as a measure of the pressure difference $P_1 - P_2$ as shown in Fig. 8.114. If $A_1 >> A_2$ or A_2/A_1 is negligible, then $P_1 - P_2 = \rho gl \sin\theta = \rho gh_2$. In such cases the reading on one side only, i.e.,'l' is required.

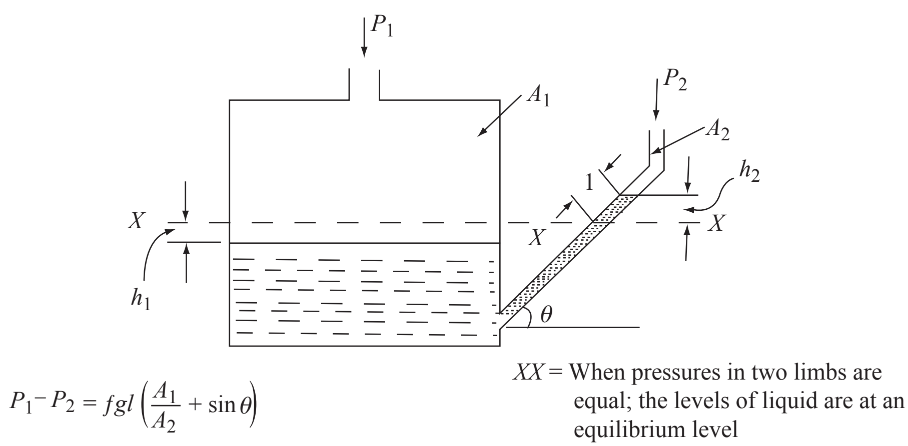


Figure 8.114 Inclined tube manometer

For increased accuracy in reading the output of the manometer, i.e., liquid displacements can be measured with micrometer heads, and the contact between the micrometer movable points and the liquid may be sensed electrically or visually.

8.33 ELASTIC TRANSDUCERS

Elastic elements when subjected to pressure get deformed. The deformation when measured gives an indication of the pressure. These elements may be in the form of diaphragms, capsules, bellows, Bourdon, or helical tubes (Fig. 8.115). The deformation may be measured by mechanical or electrical means. These devices are convenient to use and can cover a wide range of pressure depending upon the design of the elastic elements.

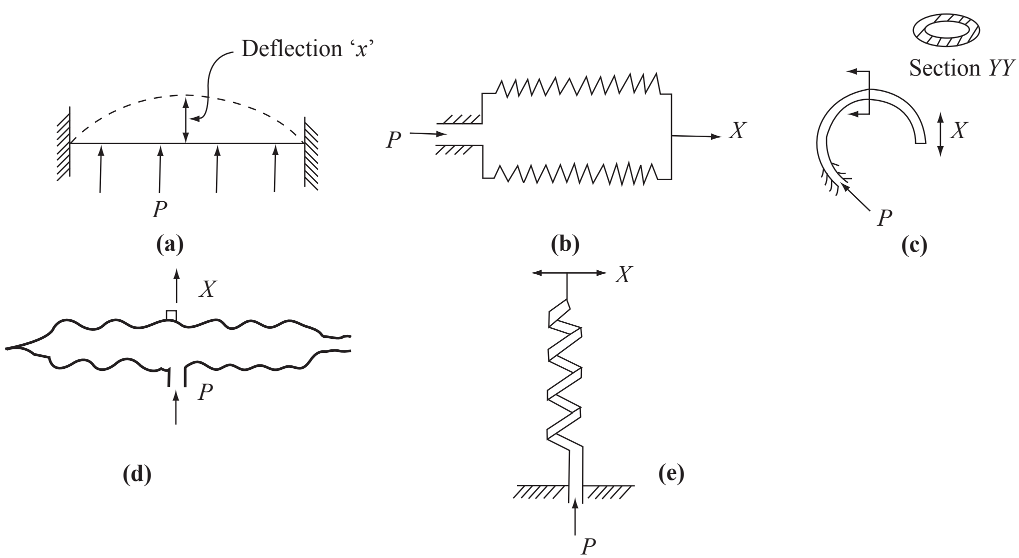


Figure 8.115 Elastic elements: (a) diaphragm (b) bellows (c) Bourdon tube (d) capsule and (e) helical tube

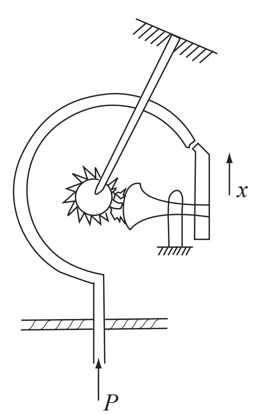


Figure 8.116 Bourdon tube pressure gauge

Types of measuring devices: A Bourdon gauge is commonly used for measuring pressure. This incorporates an elastic Bourdon tube. The cross section of the tube, due to pressure, tends to round out changing its shape. The tube uncoils, since the inner and outer arc lengths remain almost equal to their original lengths. The motion of the end of the tube is amplified and indicated by a printer moving on a calibrated scale (Fig. 8.116).

Usually, an electromechanical transducer is used along with the elastic element, especially when dynamic pressures are to be measured (Fig. 8.117). The output voltage can be indicated by an oscilloscope or recorder. Figure 8.118 shows an LVDT type of pressure transducer with bellows, which is communicated to the bellows, whose motion gives an output voltage proportional to the pressure to be measured. Figure 8.119 shows a variable capacitance-type pressure transducer in which due to pressure P the elastic diaphragm deflects, changing the capacitance between it and the fixed electrode. The output of the bridge is proportional to pressure P. Similarly, a piezoelectric crystal can be used for measuring only dynamic pressures by exposing one of the surfaces of the crystal to the pressure.

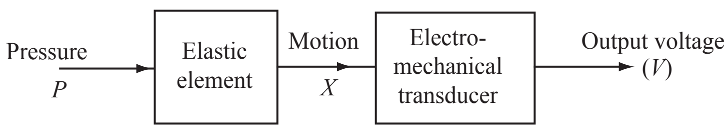


Figure 8.117 Pressure measurement using an elastic element and an electromechanical transducer

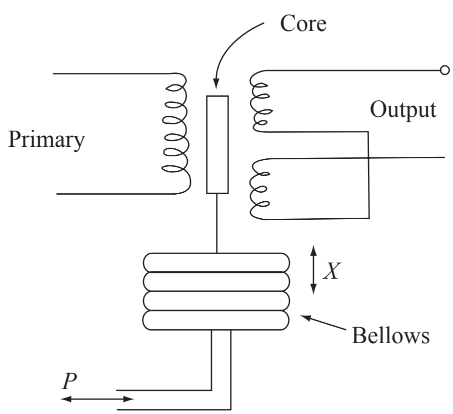


Figure 8.118 LVDT-type pressure transducer

Three types of pressure transducers, using resistance gauges are shown in Figs. 8.120, 8.121, and 8.122. In Fig. 8.120 resistance gauges R_1 and R_3 are bonded, so as to measure radial strain near the outer radius of the diaphragm while R_2 and R_4 are bonded near the centre and measure tangential strains. Expressions for radial stress σ_r and tangential σ_t at a radius 'r' are

$$\sigma_r = \frac{3pR^2v}{8t^2}\left[\left(\frac{1}{\gamma}+1\right)-\left(\frac{3}{\gamma}+1\right)\frac{r^2}{R}\right]$$

$$\sigma_t = \frac{3pR^2v}{8t^2}\left[\left(\frac{1}{\gamma}+1\right)-\left(\frac{1}{\gamma}+3\right)\frac{r^2}{R}\right]$$

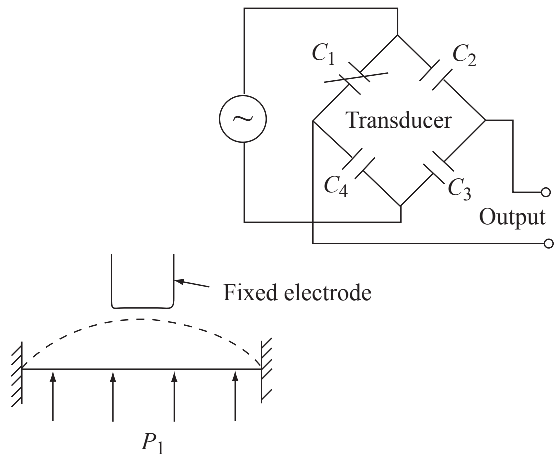


Figure 8.119 Capacitance-type pressure transducer

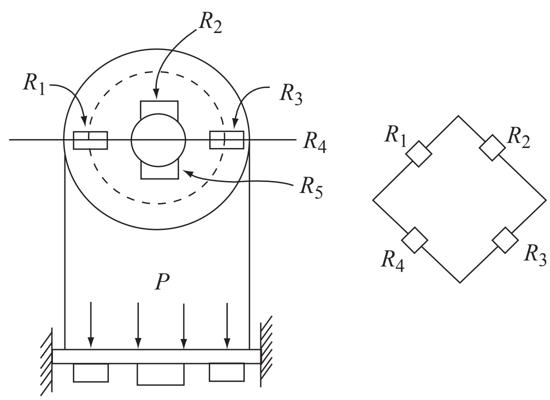


Figure 8.120 Diaphragm-type strain gauge pressure transducer

where 't' is the thickness of diaphragm, 'γ' Poisson's ratio of diaphragm material, and P the pressure. A plot of radial and tangential stresses (Fig. 8.120) shows that stresses in R_1 and R_3 would be opposite in nature to those in R_2 and R_4.

Figure 8.121 shows a strain gauge pressure transducer in which the resistance gauge is bonded on a member that is strained longitudinally. This type of transducer can be used for high pressure. Figure 8.122 shows a strain gauge pressure transducer, meant for measuring the pressure of a fluid in a pipe, without disturbing the fluid. Strain gauge R measures the strain in the pipe due to the pressure of the fluid.

Except for the piezoelectric type, other transducers here can be used for both static and dynamic measurements. The advantage of a piezoelectric-type transducer lies in its increased sensitivity and thus higher output for a given pressure change (Fig. 8.123).

Elastic element characteristics: Elastic elements used for pressure transducers have to be carefully designed for:

1. Deflection due to pressure.
2. Dynamic considerations.

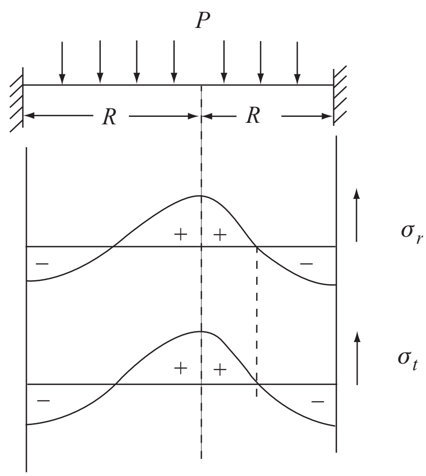


Figure 8.121 Plot of radial and tangential stresses

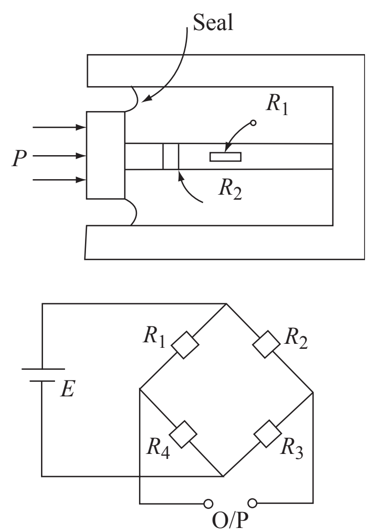


Figure 8.122 Piston-type strain gauge pressure transducer

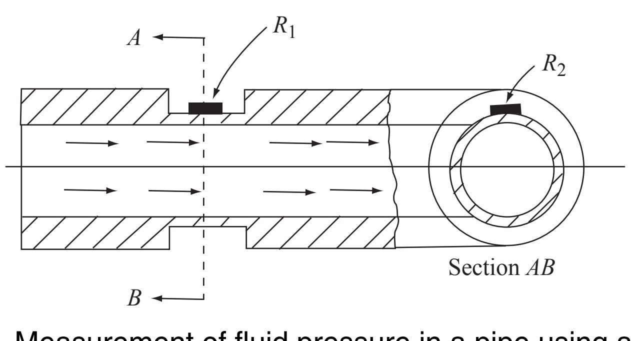


Figure 8.123 Measurement of fluid pressure in a pipe using a strain gauge

It is known that an elastic diaphragm would remain linear for small deflections only and hence maximum deflection '*y*' should be < *t*/3, '*t*' being the diaphragm thickness. For dynamic considerations it is important to check that the fundamental frequency of variations of the elastic element is higher than that of the exciting frequency due to fluctuating pressure.

It may be further seen that the dynamic characteristics of a pressure measuring system are dependent on not only the characteristics of the elastic-transducing element but also upon the characteristics of the pressure-transmitting fluid and the connecting tubing. The effective mass of the moving system depends upon the mass of the fluid and the connecting tubing. The above factors may be considered in predicting the dynamic response.

8.34 HIGH-PRESSURE MEASUREMENT

For pressure above 1000 atm, special techniques have to be used. One such technique is based upon the electrical resistance change of a manganin (alloy of Cu, Ni, Mn) or gold chrome wire, with hydrostatic pressure, due to bulk compression effect. Figure 8.124 shows an outline diagram of this type of device. Usually, the coil is enclosed in flexible bellows (not shown in figures) filled with kerosene for transmitting the pressure to be measured to the coil. The change in the resistance of the wire between *A* and *B* is measured by usual methods such as a Wheatstone bridge etc.

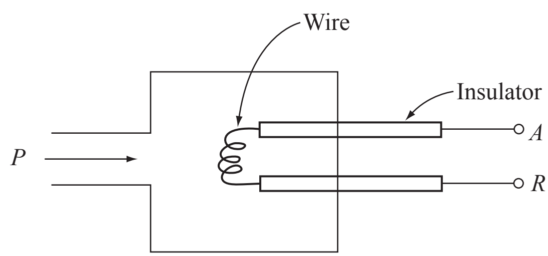


Figure 8.124 High-pressure transducer

For, manganin, the sensitivity is $2.5 \times 10^{-11}\Omega/\Omega$-Pa (pascals) while for gold chrome, the same is $9.85 \times 10^{-12}\Omega/\Omega$-Pa. Even though gold chrome is less sensitive, it is preferred to manganin, since the former is less temperature sensitive than the latter.

8.35 LOW-PRESSURE MEASUREMENT (VACUUM MEASUREMENT)

Units of vacuum measurement are Torr and micron. *One Torr is a pressure equivalent of 1 mm Hg while one micron is* 10^{-3} *Torr*. Manometer and elastic element gauges can be used to about 0.1 Torr. Below these ranges, other types of vacuum gauges are required.

8.35.1 Thermal Conductivity Gauge or Pirani Gauge

It consists of a platinum filament enclosed in a chamber as shown in Fig. 8.125. The wire forms an arm of a Wheatstone bridge. The temperature of the wire, for a given magnitude of current, depends on the rate of heat dissipation, which in turn depends on the conductivity of the surrounding medium

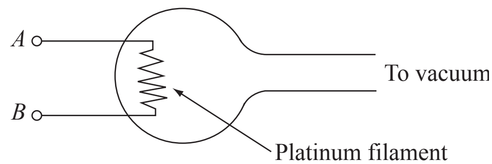


Figure 8.125 Pirani gauge

and hence its pressure. Thus, with change in pressure of the medium, the temperature and hence the resistance of the wire changes, which can be measured by using the Wheatstone bridge.

Ionisation gauge This is used for measurement of very low pressure, of the order of 1 μ and below. The gauge consists of a triode vacuum tube (Fig. 8.126). The heated cathode emits electrons that are accelerated by the positively charged grid. As the electrons move towards the grid, they ionise the gas molecules through collisions. The plate is maintained at a negative potential so that positive ions are collected by the grid. They ionise the gas molecules through collisions. The plate is maintained at a negative potential, so that positive ions collect there, producing plate current i_l. The electrons and negative ions are collected by the grid, producing grid current i_2. It is found that the pressure of the gas is given by

$$p = \frac{1}{k}\frac{i_1}{i_2}$$

where k is termed the sensitivity of the gauge.

The following table gives the approximate range for various types of pressure-measuring devices discussed:

Measuring Device	Pressure Range
Manometer	$10 - 16^6$ Pa (N/m^2)
Bourdon gauge	$108 - 5 \times 10^8$ Pa
Elastic diaphragm with LVDT or capacity or resistance strain gauge transducer	$100 - 10^8$ Pa
Piezoelectric transducer	$10^4 - 10^8$ Pa
Hydrostatic compression gauge	100 – 105 atm
Pirani gauge	$10^{-4} - 1$ Torr
Ionisation gauge	$10^{-12} - 10^{-3}$ Torr

8.36 TEMPERATURE MEASUREMENTS

Temperature is probably the most widely measured and frequently controlled variable in the numerous industrial processes. In addition it forms an important governing parameter in thermodynamics, heat transfer, and a number of chemical reactions/operations. It is also a fundamental quantity in much the same way as mass, length, and time. Temperature is defined as the degree of hotness or coldness of a body or an environment on a definite scale (in layman's language). We can also define temperature as a condition of a body by virtue of which heat is transferred to or from other bodies. However, there is a

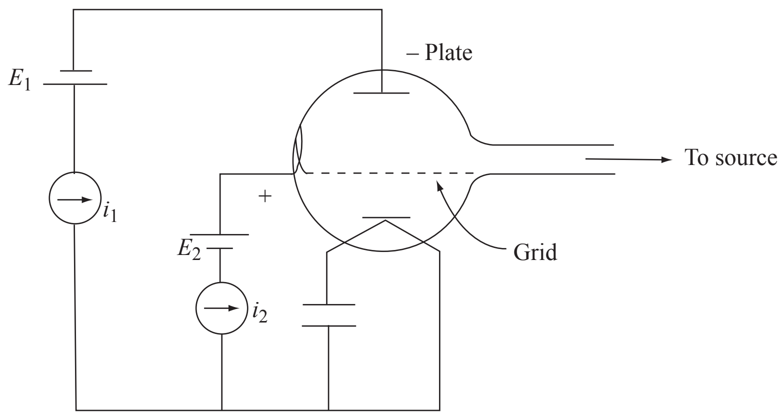


Figure 8.126 Ionisation gauge

marked difference between the quantity's temperature and heat. Temperature may be defined as '*Degree' of heat* whereas heat is taken to mean as '*quantity' of heat*.

Temperature cannot be measured directly but must be measured by observing the effect that temperature variation causes on the measuring device. Temperature measurement methods can be broadly classified as follows:

1. Non-electrical methods
2. Electrical methods
3. Radiation methods

Non-electrical methods: The non-electrical methods of temperature measurement can be based on any one of the following principles:

1. Change in the physical state.
2. Change in chemical properties.
3. Change in physical properties.

Electrical methods and radiation methods are explained in previous sections.

8.37 DATA ACQUISITION SYSTEMS

Data acquisition systems (DAS) as the name implies acquire data and record the data. The signals are obtained from transducers or measuring instruments and are processed (signal conditioning) and displayed/recorded. These are used in instrumentation systems for the measurement and subsequent processing, display, or recording of the parameters.

The signals in DAS are obtained from:

1. *Direct measurement*: Signals are obtained from measuring instruments such as DC/AC voltmeters, ammeters, frequency counters, *R*, *L*, *C* bridge circuits, etc. These methods are used in electronic component testing, environmental studies, etc.
2. *Transducers*: Signals are obtained from various sensors for pressure, force, velocity, etc., strain gauges, thermocouples, etc. Physical quantities such as acceleration, flow, etc. are converted into electrical signals by the transducer and are given to the DAS.

Data acquisition systems can also be classified as follows:

1. Analog systems.
2. Digital systems.

The block diagram of an analog DAS is shown in Fig. 8.127.

The signal conditioning device may be an amplifier, a filter, a rectifier, a mixer, etc. The display or recording device may be a Cathode Ray Oscilloscope, X–Y recorder, magnetic tape recorder, analog display, or electronic display system. The block diagram of a digital DAS is shown in Fig. 8.128.

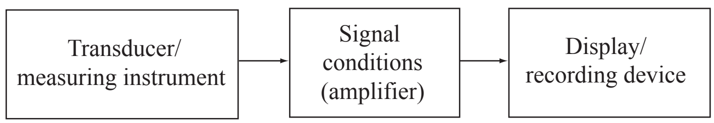


Figure 8.127 Analog data acquisition system

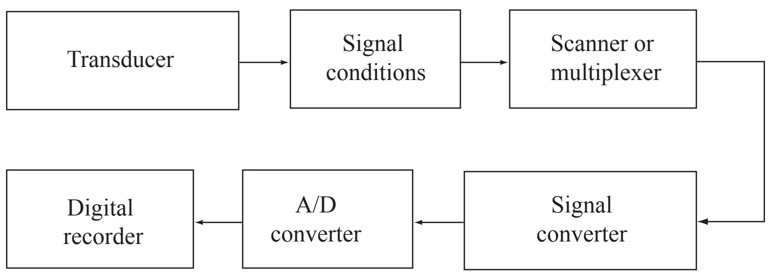


Figure 8.128 Block schematic of digital data acquisition system

A scanner or a multiplier accepts multiple analog inputs and sequentially connects them to one measuring instrument.

A signal converter converts the signal to a form at an acceptable level to the A/D converter (DAS).

Data acquisition systems are used where large numbers of parameters are to be measured, monitored, and controlled.

Some of the areas where these are used are as follows:

1. Industries.
2. Biomedical applications.
3. Scientific research.
4. Aerospace.
5. Telemetry.

The output of a transducer is usually very small. In addition, in the industrial environment, electrical noise signals will be predominant. Therefore, the signal must be amplified and noise must be filtered. Instrumentation amplifiers are used in DAS, which provide the following:

1. Large differential gain Ad, large Common Mode Rejection Ratio (CMRR), to reduce noise.
2. High-input impedance and low-output impedance. The value of CMMR for low-frequency signals such as powerline frequencies is as high as 100 db.

The connecting wires are shielded from an external signal pickup by 'guarding'. The circuit as is shown in Fig. 8.129(b).

In Fig. 8.129(a), the signal from the transducers or source is connected to the meter through wires. These wires are close to a powerline wire, It is capacitively coupled and, therefore, couples powerline frequency voltages into a signal amplifier. The powerline interference is conducted to the ground with identical currents from each side of the line, if the amplifier is perfect. However, it is not the case practically. If the leakage resistance or capacitance is different for one line relative to the other, the currents to the instrument ground are from only one side of the differential line. By adding a shield, which is connected to one side of the signal and to the instrument case, the capacitively coupled signals from the power frequency line are coupled signals from the power frequency line and are safely conducted to the instrument case and to the ground (Fig. 8.129).

If the strength of a noise signal is very large and more shielding will not serve the purpose, isolation amplifiers are to be used.

If a low-level signal is transmitted over a significant distance, errors occur. Such errors can be due to the resistance of interconnecting wires. The circuit is shown in Fig. 8.130.

The expression for the voltage across the signal load V_L is

$$V_L = \frac{V_{out} + I_{supply} R R_L}{2R + R_L}$$

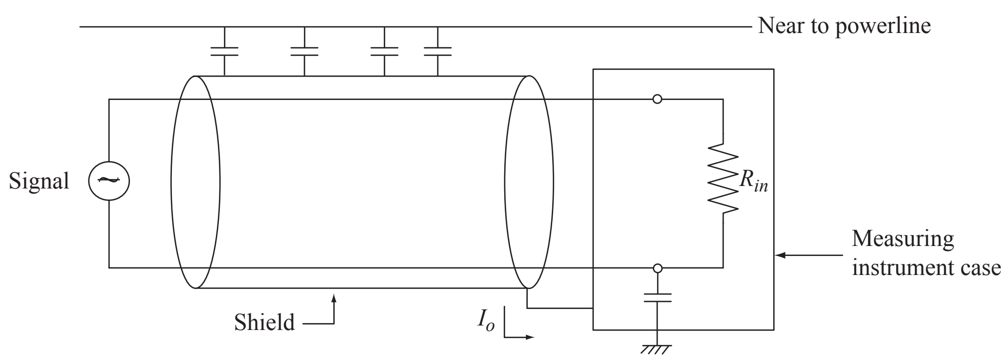


Figure 8.129 Generating circuit for shielding wires from external signal pickup

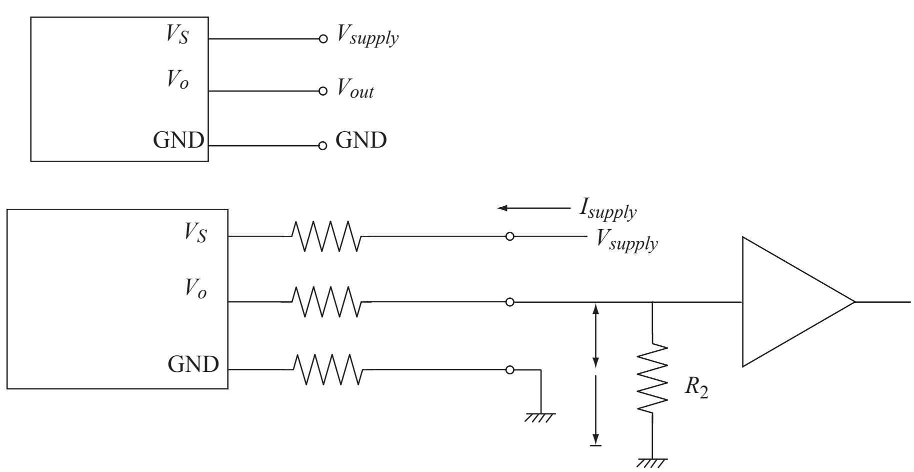


Figure 8.130 Effect of lead resistance

Therefore, current, rather than voltage, is used for transmission of signals. A typical current loop transmission system uses currents from 4 to 20 mA for the full-scale transmission of a parameter. A 4 mA level current includes the power supply current. A voltage-to-current converter circuit for 4-to-20 mA current loop is used. Two voltage-controlled current sources are used in the converter. One current source senses the power supply current for the amplifier plus the current from the voltage-controlled current source, I_2 and the sum is equal to 4 mA. The second voltage-controlled current source I provides a variable current as a function of the transducer voltage as provided by the instrumentation amplifier. The second voltage-controlled current source provides from 0 to 16 mA for a total current from 4 to 20 mA.

The 4 to 20 mA current loop method of transmission reduces errors caused by the resistance of the interconnecting wires. However, in situations where a current loop technique or using an isolation amplifier do not reduce errors, the analog signal is connected to a digital form for transmission. A *V/F* conversion technique is one such method.

8.38 SUMMARY

Synchros are used for displacement measurements. Angular displacement 'θ' is converted into electrical output 'e_0' Microsyn is another variable reluctance element. It is widely used in sensitive gyroscopic instruments.

In flow-measuring instruments, the movement of the fluid stream is directly or indirectly used to actuate a secondary device. Rotameter is another type of flow meter classified under area flow meters.

Variable capacitance transducers are used to determine parameters such as like force, displacement, pressure, etc. Carbon microphone is a capacitance transducer. The sound vibrations produced on a thin metal membrane diaphragm cause vibration in capacitance and the output voltage changes. Piezoelectric transducers are active transducers. Due to piezoelectric effect, electrical output voltage is produced due to mechanical input such as force, pressure, strain, etc. The natural materials exhibiting this effect are Quartz and Rochelle salt. The synthetic materials are ammonium dihydrogen phosphate, Barium Titanate, Lead Zirconate Titanate, etc., d_{33} is a parameter indicating the electrical charge produced to the force applied. Its units are Coulombs/Newton.

Points to Remember

- Synchros are used for angular displacement measurement.
- Microsyn is another variable reductance element.
- For flow measurement, head flow meters and area flow meters are used.
- Rotameters are one type of area flow meters.
- Variable capacitance transducers are used for displacement, force, etc. When air is the dielectric due to change in spacing between the places, capacitance changes, which is a measure of the physical parameter (force) thus causing change in spacing.
- Carbon microphone is also a capacitance-type transducer. The sound vibrations on a thin membrane cause a change in capacitance.
- Quartz and Rochelle salt are naturally occurring piezoelectric materials.
- PZT, PDP, $BaTiO_3$ are synthetic piezoelectric materials.
- Piezoelectric transducers are active transducers.
- 'd'coefficient of a piezoelectrical material denotes the electric charge output provided for the force applied. Its units are Coulombs/ Newton.

Objective-type Questions

1. Synchros are used for the measurement of ________.
2. Selsyn is also known as ________.
3. Selsyn and synchros are ________ types of transducers.
4. The two types of flow meters are (A) ________ (B) ________.
5. Rotameters are used for the measurement of ________.
6. Carbon microphone is ________ type of transducer.
7. Piezoelectric transducers are ________ type of transducers.
8. Naturally occurring piezoelectric materials are ________.
9. Synthetic piezoelectric materials are ________.
10. The units of '*d*' coefficient of a piezoelectric materials are ________.
11. The relation between '*g*' coefficient and '*d*' coefficient is ________.
12. The relation between '*g*' coefficient and '*h*' coefficient is ________.
13. A single wire type of a transducer for the measurement of flow responds essentially to the component of velocity ________ to it, if the angle between wire and velocity vector is between 90° and 25°.
14. In the case of flow-measuring-type hot wire anemometers, the heat loss is greater than predicted by the component of velocity vector $V \sin \theta$, if θ is ________.
15 Flutter effects in hot wire anemometers occur at high speeds because ________.
16 In an electrolytic hygrometer, the materials used
17 Photo-conductive cells are basically ________ type transducers.
18 In photoconductive cells, lead sulphide material can be used upto a wavelength of ________.
19 In photoconductive cells, the material that can be used upto a wavelength of 0.9 μ is ________.
20 The frequency range of a carbon microphone transducer is ________.

Review Questions

1. Explain the principle and working of resistance thermometers.
2. What are the advantages and disadvantages of resistance thermometers?
3. What are the different types of hot wire anemometers? Explain the principle of each type.
4. Explain the constructional details, principle, and working of an LVDT.
5. Give the specifications and typical values of the parameters for an LVDT.
6. Describe the applications of an LVDT.
7. Explain about non-electrical and electrical methods of temperature measurement.
8. Give the constructional details of thermocouples.
9. State and explain the laws of thermocouples.
10. Give the constructional details, applications, and materials used for thermistors.
11. Explain the principle of flow measurement.
12. Discuss the theory and characteristics of head flow meters.
13. Describe the principle and constructional details of magnetic flow meters.
14. Describe the principle of variable capacitance transducers.
15. What are the materials employed for piezoelectric transducers? Mention these characteristics.
16. Describe the relationship between different piezoelectric coefficients.
18. Describe the principle and working of a piezoelectric accelerometer.
18. Explain about the principle and applications of a magnetostrictive transducer.
19. What are the different types of errors that occur in magnetostrictive transducers? Explain.

Unsolved Problems

1. The voltage sensitivity of a piezoelectric crystal is 0.06 V m/N. The thickness of the crystal is 3 mm. Determine the output voltage when it is subjected to a pressure of 2×10^6 N/m^2.
2. A $BaTiO_3$ piezoelectric pickup has dimensions of 5 mm × 5 mm × 1.8 mm. The voltage sensitivity of the transducer is 0.015 Vm/N. Determine the voltage developed when a force of 1 kg is applied to the transducer.
3. A variable reluctance-type tachometer has 100 teeth on the rotor. The speed of shaft on which the rotor is mounted is 1500 rpm. Determine the frequency.
4. An accelerometer has a mass of 0.05 kg and a spring constant of 5000 N/m. The maximum mass displacement is ± 0.03 m. Determine (a) measurable acceleration and (b) notional frequency.
5. The *d*-coefficient of a piezoelectric material is 0.8 × 10–12 Coulombs/N. Absolute dielectric constant is 4 × 8.85 × 10–12 F/m. Determine the value of *g*-coefficient and *h*-coefficient if Young's modulus of elasticity of the material is 1.2×10^6 kgf/cm^2.

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Appendix A

Answers to Objective-Type Questions

Chapter 1

1. Measurand.
2. Amplifier/filter/modulator/AC modulator/A/D converter/D/A converter.
3. Digital.
4. Resolution.
5. (a) Zero drift or calibration drift, (b) span drift, (c) zone drift.
6. Fidelity.
7. Lag.
8. $\pm 0.67450\ \sigma$.
9. $y = \dfrac{h}{\sqrt{\pi}} e^{-h^2 w^2}$.
10. $T = BAIN$.
11. Ammeters/current meters.
12. $10^5/f_c$.
13. A.
14. Resolution.
15. Hysteresis.
16. Thermal zero shift.
17. Middle of the scale.
18. Two.
19. Spring.
20. High.

Chapter 2

1. AF and RF.
2. Standard signal generator.
3. AF and RF.
4. Test oscillator.
5. Sine, square, triangular, ramp, pulse.
6. Duty cycle, frequency, and amplitude.
7. Small variation in amplitude and frequency that can be produced.
8. Frequency.
9. Sine waves only.
10. Synthesised, triangular.
11. 0.01 Hz.
12. Diode resistor network.
13. Low.
14. Signal generator.
15. mW.
16. Sine wave.
17. Positive feedback.
18. Audio.
19. Three.
20. The timing range is limited.

Chapter 3

1. Distortion and stability of output.
2. Harmonic distortion analysers, wave analysers, and spectrum analysers.
3. 600 dB/octave.
4. 75 db.
5. 10 kHz to 18 MHz.
6. Wave analysers.
7. Locally generated RF signal.
8. The sizes of L and C are large.
9. The mixer introduces spurious cross - modulation products.

10. $B = \sqrt{D_2^2 + B_3^2 + D_4^2} \ldots$, where

$D_2 = \frac{B_2}{B_1}; D_3 = \frac{B_3}{B_1}; D_4 = \frac{B_4}{B_1}$.

11. 0–30 Hz.
12. Separate two signals, closely repeated.

Chapter 4

1. *Y*-amplifier.
2. Saw-tooth waveform.
3. Of the order of kV.
4. True representation of input signal.
5. Thermionic emission.
6. Graticules.
7. DC offset-voltage adjustment.
8. Free-running stable or scanning mode.
9. A straight line passing through the origin.
10. (a) Mesh storage, (b) phosphor storage.
11. Pre-accelerating and accelerating anodes.
12. Barium and strontium oxide-coated cathode.
13. Inside the surface only.
14. To collect secondary emission electrons.
15. Focusing anode voltage.
16. Proper synchronisation between the signal and sweep generator.
17. Changing the time base.
18. Millers' sweep circuit.
19. Lissajous figures.
20. Circle.

Chapter 5

1. (a) Alternate mode, (b) chopped mode.
2. 100 kHz.
3. 1 Ω shunted by 12 pf.
4. (a) Direct probe, (b) high-impedance probe, (c) high-voltage probe, (d) detector probe.
5. Large.
6. 80 nsecs to 200 nsecs.
7. (a) Lumped parameter delay zone,
   (b) distributed parameter delay line.
8. t_d = 180 nsec/m.
9. Input signal.
10. Gating signal.
11. $f_o = \sqrt{f_c}$.
12. Period mode.
13. Astigmatism.
14. To lock the display of signal.
15. 5 MHz.
16. PBA voltage.
17. Non-triggered oscilloscope.
18. Double beam oscilloscope.
19. One.
20. Storage CRO.
21. DC.
22. Current.

Chapter 6

1 Wheatstone bridge and Kelvin's bridge.
2 AC, DC.
3 Kelvins.
4 1–0.0001Ω.
5 AC, both magnitude and phase angle are to be balanced.
6 Inductance.
7 Inductance.
8 Wien bridge.
9 Wagner's ground connection.
10 Opposite angle bridge.
11 Inductance cannot be measured over a wide range.
12 Inductance.

13 Maxwell's bridge.
14 $V_{I_c} \tan \delta$.
15 Anderson's bridge.
16 Hay bridge.
17 Few volts.
18 Vibration galvanometer.
19 Schering bridge.
20 Eliminating the effect of inter-component capacitances.

Chapter 7

1. Two types—active transducers and passive transducers.
2. Three types—resistance, capacitance, and inductance.
3. Photoconductive cell.
4. Change in dimensions due to magneto-strictive.
5. Semiconductor type.
6. G.F = 1 + 2γ.
7. Nylon.
8. Negative.
9. Constant current type, constant temperature type.
10. Lithium chloride.
11. Few microns to few cms.
12. Quartz and Rochelle salt.
13. Strain gauge, straining member.
14. Nylon.
15. Teflon.
16. High.
17. High.
18. The gird support, the gauge.
19. Negligible.
20. Liquid level.

Chapter 8

1. Angular displacement.
2. Microsyn.
3. Inductive.
4. (a) Head flow meters, (b) area flow meters.
5. Flow.
6. Capacitance.
7. Active.
8. Quartz and Roche salt.
9. ADP, $BaTio_3Z$, P.
10. Coloumbs/Newton.
11. $\frac{d}{f} = \varepsilon$ (dielectric constant).
12. $g \times y = h$, where y = Young's modulus.
13. Perpendicular.
14. Less than 25°.
15. Wires may vibrate due to aerodynamic loads.
16. Lithium chloride salt in a 3% concentration.
17. Variable resistance.
18. 3.5 μ.
19. Lead sulphide.
20. 30–3000 Hz.

References

1. E.O. Doebelin, 2004. *Measurement Systems: Application and Design*. McGraw-Hill: Boston.
2. A.D. Helfrick, W.D. Cooper, 1990. *Modern Electronic Instrumentation and Measurement Techniques*. Prentice-Hall: Englewood Cliffs, NJ.
3. D.A. Bell, 2007. *Electronic Instrumentation and Measurements*. Oxford University Press: Canada.
4. R.A. White, 2002. *Electronic Test Instruments: Analog and Digital Measurements*. Prentice-Hall: Englewood Cliffs, NJ.
5. A.S. Morris, 2001. *Measurement and Instrumentation Principles*. Butterworth-Heinemann: Oxford.
6. C.S. Rangan, G.R. Sarma, V.S.V Mani, 1983. *Instrumentation, Devices and Systems*. Tata McGraw-Hill: New Delhi.
7. Available from www.Tektronix.com.
8. Available from www.Keithley.com.

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