DESIGN AND CONSTRUCTION OF HIGH SENSITIVITY A.C. SUSCEPTIBILITY BRIDGE

BY

FARKHUND SHAKEEL MAHMOOD

A DISSERTATION SUBMITTED IN THE PARTIAL FULFILMENT OF THE REQUIREMENTS OF THE DEGREE OF MASTER OF PHILOSOPHY IN PHYSICS

DEPARTMENT OF PHYSICS
QUAID-I-AZAM UNIVERSITY
ISLAMABAD

(1987)

C E R T I F I C A T E

Certified that the experimental work in this dissertation was carried out and completed under my supervision.

(Dr. Syed Khursheed Hasanain)
Assistant Professor
Department of Physics
Quaid-i-Azam University
Islamabad.

Submitted through

(Prof. Dr. Masud Hussain)
Chairman
Department of Physics
Quaid-i-Azam University
Islamabad

DEDICATION

DEDICATED

TO

MY PARENTS AND WIFE WHOSE
COOPERATION AND ENCOURAGE-
MENT LED ME TO COMPLETE MY
RESEARCH WORK.

CONTENTS

CONTENTS

ACKNOWLEDGEMENT

ACKNOWLEDGEMENT

All praise to Almighty Allah, the most Merciful and most compassionate, the creator of the universe, who enabled me to complete this research successfully. I offer my humblest and sincerest words of thanks to His Holy Prophet Muhammad (Peace be Upon Him) who is for ever a torch of guidance and knowledge of humanity.

It is a great pleasure for me to acknowledge my supervisor, Dr.Syed Khursheed Hasanain for his keen interest in the research work and his personal attention, scrupulous guidance, sympathetic attitude, valuable discussions and encouragement. Under the supervision of whom the present research was completed.

I am grateful to Professor Dr. Masud Hussain, Chairman Department of Physics, Quaid-i-Azam University, Islamabād for providing all the facilities during my research work.

I have no words to express gratitude for Dr.Asghari Maqsood and all the staff members of laboratory for their very well behaviour and cooperation, during my stay at laboratory without which the manuscript would have never attained its present shape.

I feel pleasure to acknowledge my friends Mr.Muhammad Saeed, Mr.Muhammad Rashid Mirza, Mr.Muhammad Asghar Gill and Mr. Malik Dur-i-Iman due to their cooperation and fruitful discussion in the accomplishment of this research work.

My thanks are also due to Professor Dr.Ghulam Mohammad Malik, Former Principal Govt. College, Multan, and Professor Mohammad Bashir, Principal(Retd.) and Professor Mohammad Saeed Ashraf, Principal, Govt. College, Multan their moral encouragement, cooperation and sincerest advice have paved my path to success.

Last but not least, the author also owes a debt of gratitude to his affectionate parents, wife, sister, brothers and daughter who remembered him in their prayers and encouraged him throughout the span of study.

(Farkhund Shakeel Mahmood)

ABSTRACT

ABSTRACT

We have used an existing design of an inexpensive mutual inductance bridge circuit to construct a high Sensitivity A.C. Susceptibility bridge. which uses integrated operational amplifiers. This bridge circuit in conjunction with a lock in amplifier can be used for measurement of magnetic susceptibility of ferro, ferri, and paramagnetic samples. The circuit employs operational amplifiers both for balancing the bridge and for detecting the differential signal from the sample coil. The simple design of the entire system and the use of low cost components are the special features of this system.

At present the sensitivity of the bridge is approximately, 0.1mV, and further improvements are possible. The effect of temperature on the electronic noise in the system was also determined. We also find good agreement between the known value of permeability of magnetic steel and the value by us, using this bridge circuit measured.

INTRODUCTION

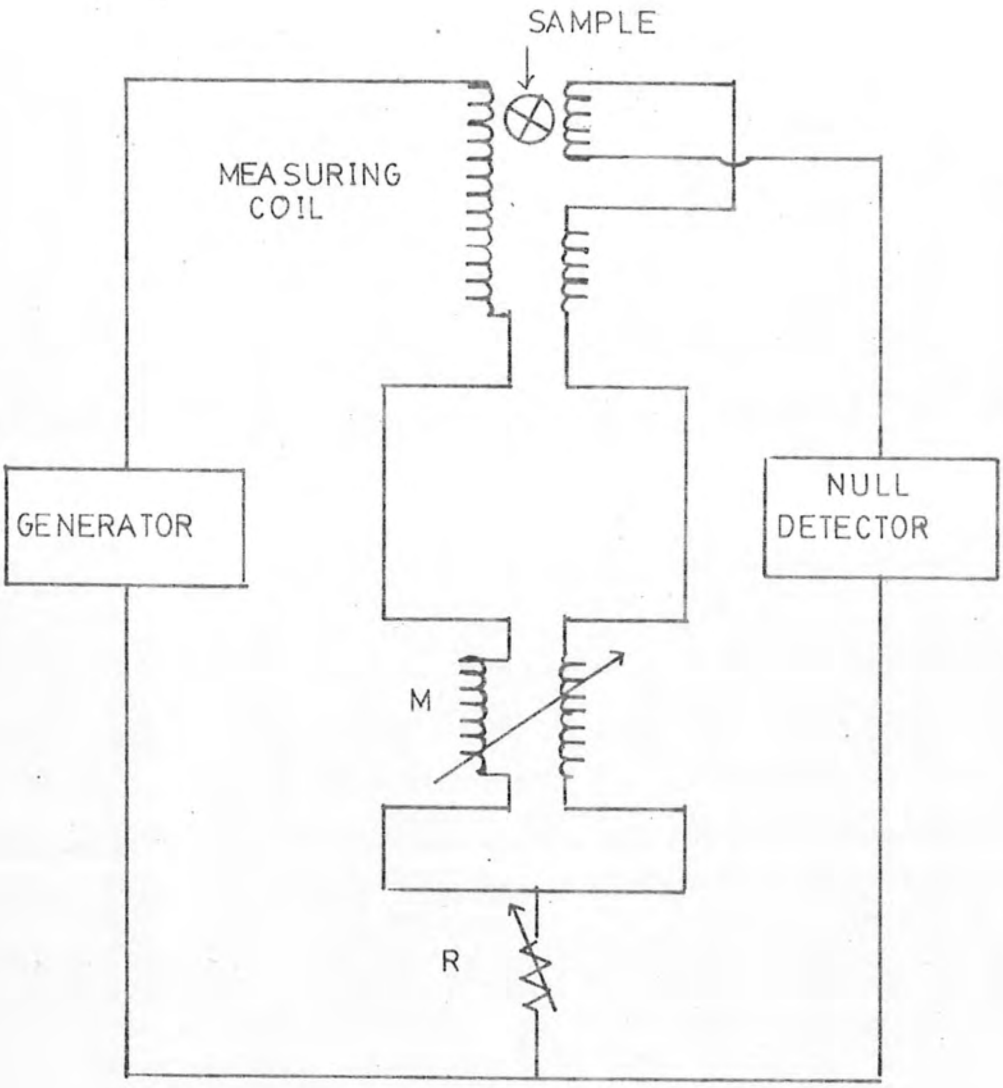
INTRODUCTION

A large number of AC mutual inductance bridge for magnetic susceptibility have been reported in the world- viz-a viz Maxwell 1965, Daybell 1967, Parks and Swvenson 1967, Anderson 1968, Anderson et al 1970, Halperin et al 1981. All the systems use expensive ratio transformers to minimise the noice and increase the sensitivity of the mutual inductance bridge used in their system. Other designs have been presented namely. Pundon and Kretschmar 1978, Brodbleck et al 1978, Ran et al. 1978, Gentile et al 1981, for mutual inductance bridges having operational amplifiers Although these designs are successful in providing economical and versatile alternatives to conventional ratio transformer design, their sensitivity is limited by finite gain and the inherent noise of the operational amplifiers.

The basic form of the conventional AC mutual inductance bridge was first indroduced by Hartshorn. The basic Hartshorn bridge circuit is shown in figure 3.12. The measuring coil typically is placed in the cryostate and consists of two identical secondaries and a coaxial primary. The secondaries are separated along the axis of the primary. ~~The secondaries are separated along the axis of the primary~~ coil and are connected in opposition. In principle, the net voltage induced across the two secondaries is zero until a paramognetic sample is inserted into one of them. The resulting induced e.m.f. may then be opposed by the voltage across the secondary of the mutual inductance M and the resistance R. At balance, M is proportional to the real part of the susceptibility of the sample and R is proportional to the power dissipation. For thermometry it is desirable to minimize the latter which is proportional to ωH^2. Where H is the amplitude of the magnetic field at the sample and ω is its frequency.

A disadvantage of Hartshorn bridge is that it requires a precise variable mutual inductance which is both expensive and limited range. It is possible, however, to overcome this drawback by simulating a variable mutual inductance electronically with an operational amplifier and fixed inductance. The Brodleck bridge incorporated three basic modifications of the conventional Hartshorn bridge. First, a ground is inserted between the two secondaries of the sample coil placing them in parallel rather than in series. Secondly, the standard input-resistance adjustment of the differential operational amplifier serves as a simple resistive balancing mechanism in the secondary circuit. This eliminates the need for a complex resistive balancing network which directly couples the secondary and primary circuits. Thirdly, a variable-gain op-amp. replaces the vacuum triod used in the earlier circuits. This modification substantially broadens the range over which the bridge can be balanced.

BASIC HARTSHORN BRIDGE CIRCUIT



CHAPTER ONE

NATURAL MAGNET:

In most of the countries there exists a peculiar kind of iron ore which has a wonderful power of attracting iron. This natural magnet was known in the remotest times, being regarded as an object of curiority by the people of many nations. Homer, Thales, Pythagoras and Cicero all mention its attractive power. The ore is called the 'loadstone' from the Saxon Word Laeden and its power was named magnetism by the Greeks, from magnesia, a distric in Macedonia, where loadstone was clearly found.

Even in olden times it was known that if a piece of load stone is suspended so that it savings freely it always turn to the north. Magnetic effects appear to be concentrated in certain region of magnetics only. The centres of these regions are called the poles of the magnet. Magnetic poles are of two kinds, namely north and south poles. Like poles repel each other while unlike poles attract each other. The lines of magnetic field come out of north pole and end on south pole. The direction of magnetic field can be defined as the direction of the force with which the field acts a north magnetic pole.

Electro-magnet:

In 1820 Oersted first discovered that a current in a wire can also produce magnetic effects, namely, iron flings to cling to the wire while the current is flowing, but drop off when the current is stopped, further, it can change orientation of a campass needle. The magnetic effect of a current in a wire can be intensified by placing an iron core in it. This sort of magnet is known as Electromagnet.

Intensity of magnetization: The intensity of magnitisation is defined as magnetic moment per unit volume.
Magnetic flux: Number of lines of force linked with certain material is called flux. Flux density or magnetic induction B Magnetic lines of force falling on a unit area is known as flux density.

Permeability(u): It is the conducting power of the material for magnetic lines of force a compared with air. Mathematically permeability is the ratio of the magnetic induction in the material to the magnetizing force.

$$u = \frac{B}{H}$$

Magnetising force H: The magnetising force(H) inside a material is defined as the force on a unit north pole placed at the centre of a long and a narrow tunnel drawn in the direction of magnetization.

MAGNETIC SUSCEPTIBILITY AND PERMEABILITY

Nearly all substances can be classified into two general group as regards their magnetic properties. One group, comprising the diamagnetic and paramagnetic materials, is very weakly magnetic. It is linear and isotropic in behaviour, so that M,H and B are all linear (i.e. proportional). The other group, consisting of the ferromagnetic substances, is very strongly magnetic and its member and all decidedly non-linear.

In a large class of materials there exists an approximately linear relationship between M and H. If the material is isotropic as well as linear

$$\overline{M} = X\,\overline{H} \qquad (1)$$

Where the dimensionless scalar quantity X is called the magnetic susceptibility in analogy with the electric susceptibility we met in electrostatics.

If X is positive, the material is called paramagnetic and the magnetic induction is strength d by the presence of magnetic material, since $\overline{M}$ and H are parallel pointing in the same direction.

If K is negative, the material is dimagnetic and the magnetic induction is weakned by the presence of the material, since $\overline{M}$ and $\overline{H}$ are antiparallel. The value of X is quite small for paramagnetic as well as for diamagnetic materials. The value of X is very high for ferromagnetic materials and is a function of temperature. Sometimes, it varies quite drastically with temperature, transforming the ferromagnetic behaviour into a paramagnetic behaviour of the specimen, (Curie's law $M = C\frac{H}{T}$ or $X = \frac{C}{T}$)

By combining the following equation

$$\bar{M} = X\,\bar{H}$$

$$\bar{B} = \mu_o\,\bar{H} + \mu_o M \qquad (2))$$

$$\bar{B} = \mu_o\,H + \mu_o\,X\bar{H}$$

$$\bar{B} = \mu_o(1+X)H$$

or $$B = \mu H \qquad (3)$$

where $$\mu = \mu_o\,(1+X) \qquad (4)$$

is the magnetic permeability of the material. In free space eq.(3) reduces to the vacuum equation i.e.

$\bar{B} = u_o\bar{H}$ since X is zero in vacuum. The linear relation between B and H predicted by eq.(3) implies that B and H are parallel. For an isotropic materials B and H are not parallel. In permanent magnets B,H and M may all point in different directions.

By analogy with the dielectric constant (relative permitively) $K = \frac{\Sigma}{\Sigma o}$ in electrostatics we define a dimensionless quantity

$$K_m = \frac{\mu}{\mu_o} = 1+X \tag{5}$$

This quantity is called the relative permeability. From above, we can write

$$\begin{aligned} X &= \frac{\mu}{\mu_o} - 1 = K_m - 1 \qquad \text{MKS} \\ X &= \frac{\mu - 1}{\mu\pi} \qquad \text{Gaussian} \end{aligned} \tag{6}$$

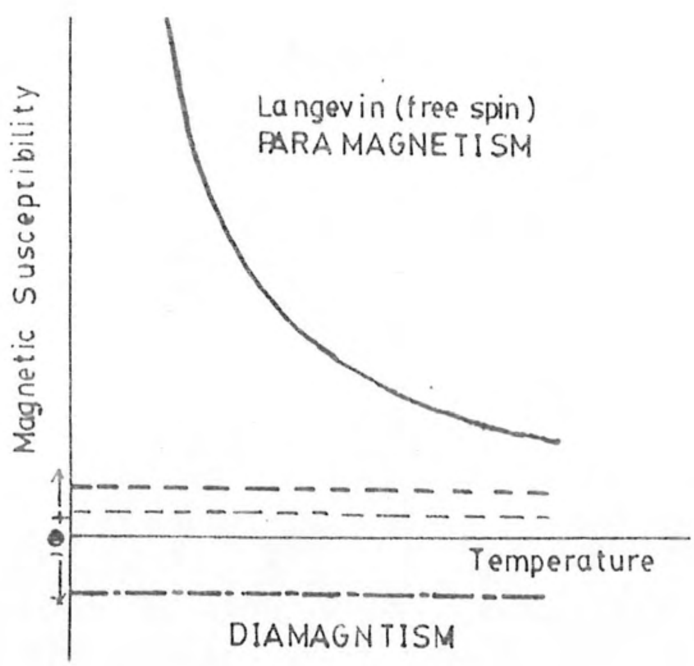
In guassian units $K_m = \frac{\mu}{\mu_o}\mu$ as $\mu_o = 1$ for free space in this system of units. We may also classify magnetic materials in terms of the magnetic permeability conclaine in eq.(5). Thus

$\mu < \mu_o$ for diamagnetic $\therefore$ X is negative

μ_o for paramagnetic $\therefore$ X is positive

$\mu >> \mu_o$ for ferromagnetic $\therefore$ X is positive and large enough.

The eq.(1) and (4) with constant X and μ donot apply to ferromagnetic materials, since they are non-linear in behaviour.


Langevin (free spin)
PARA MAGNETISM
Magnetic Susceptibility
Temperature
DIAMAGNTISM

THEORIES OF MAGNETISM DIAMAGNETISM

All matter should have diamagnetic properties. Atmos or molecules whose atomic currents have zero resultant magnetic dipole moments will not show any paramagnetic or ferromagnetic effects that mark their diamagnetism The diamagnetic susceptibility - calculations by semi-classical theory of atomic structure gives the susceptibility as proportional to the average moment of inertia of the electrons in the atom or molecules. The modern quantum theory gives the same formula, but the calculation of moments of inertia is difficult. One important result of the theory is that diamagnetism is not associated with thermal agitation and is most substance. There is a temperature dependence in certain exceptional cases that may be due to a temperature effect on the electron orbits.

PARAMAGNETISM

Paramagnetism occurs not for substances whose atoms and molecules each have a resultant magnetic moment, but also for conductors, since the spins of the conduction electrons can also be rotated by an external field. Resultant magnetic moments of molecules may arise because the orbital motions or spins of the outer electrons are not completely arranged in equal and opposite pairs. The complete inner shells of atoms always have a vanishing resultant magnetic moment. It is only incomplete shells that can contribute to paramagnetism.

The simplest theory of paramagnetism assumes that each atom with a resultant moment will tend to turn in a precessional fashion toward the direction of an applied field and that thermal agitation would tend to oppose the lining up tendency. The result is that to a good approximation, the susceptibility should behave like temperature dependent part of the electric susceptibility. That is

$$X = \frac{nm^2 u_o}{3kT} \qquad \text{MKS}$$

$$= \frac{nm^2}{3kT} \qquad \text{Gaussian} \tag{1}$$

$$X = C/T$$

where n is the number of molecules or atoms per unit volume and m is the permanent magnetic moment of each molecule or atom. The paramagnetism of the conduction electrons of metals is temperature independent, because of the location of most electrons in the filled levels of the conduction bands that are not affected by temperature.

FERROMAGNETISM

To have a vivid picture of ferromagnetism, it is of necessity to introduce the concept of molecular field, The magnetic field which is effective in its interactions with atomic currents in an atom or molecule is called the moleculecular field or local field $B_m = u_o H_m$. The molecular field H_m is the magnetic field at a molecular position in the material. It is produced by all external sources and by all molecular dipoles in the material with the exception of the one molecule or atom at the point under consideration. To calculate the molecular field, we take a material object of arbitrary shape, uniformly magnetised with magnetisation M. Let us make a spherical cavity inside this material object. The molecular field inside this cavity is given by

$$H_m = H + H_s + H' \qquad (2)$$

where H is the macroscopic magnetic intensity in the specimen, H_s is the contribution from the surface pole density σ_m' on the cavity surface and H' is the contribution of the various dipoles inside the cavity. Now H = 0 for a large class of materials and H_s is seen to be

$$H_m = H + \frac{1}{3} M \qquad (3)$$

and $$B_m = \mu_o H_m$$

This equation gives the molecular field in terms of the macroscopic magnetic intensity and the magnetisation in the sample. For most diamagnetic and paramagnetic materials the term $\frac{1}{3} M = \frac{1}{3} X_m H$ is negligibly small, but for ferromagnetic materials the correction is quite important. With the introduction of molecular field, the Eq.(   ) is equivalent to (since $M = X_m H$)

$$M = \frac{nm^2 u_o}{3kT} H_m \qquad \text{MKS} \qquad (4)$$

and reduces to

$$M = \frac{nm^2 u_o}{3kT} H \qquad \text{MKS} \qquad (5)$$

for paramagnetic materials. This is the case of weak local interaction.

In ferromagnetic materials the atomic (or molecular) magnetic moments are very nearly aligned even in the absence of an applied field. The cause of this alignment is the molecular field H_m defined by Eq.( ) and does not vanish even if $H = 0$ unless M vanishes simultaneously. Consequently, for ferromagnetic materials we replace $\frac{M}{3}$ by a larger expression M in Eq.( ) and the expression ( ) reduces to

$$M = \frac{nm^2 u_o (H + \gamma M)}{3kT} \qquad \text{MKS} \qquad (6)$$

Solving for M we get

$$M = \frac{nm^2 u_o H}{3k(T-0)} \qquad \text{MKS} \qquad (7)$$

where

$$0 = \frac{nm^2 u_o}{3k(T-0)} \qquad \text{MKS} \qquad (8)$$

$$= \frac{4\pi m^2 \gamma}{3k} \qquad \text{Gaussian}$$

It follows that for $T \gg \theta$, the susceptibility behaves in the usual paramagnetic way, with γ playing no significant role. On the other hand, if $T>>\theta$, there will be a large value of M even for a small H and for $T < \theta$, the formula gives physically meaningless results, indicating that the interaction has become so strong that M can sustain itself without any applied field H and the assumptions underlying Eq.( ) remain no longer valid.

It is in fact true that many substances are paramagnetic above a certain temperature and follow Eq.(5) fairly well. Below the temperature O, called the 'cruite temperature' these substance become ferromagnetic. The theory we have just given is the Weiss theory of ferromagnetism, first announced in 1907. Curie temperatures of iron alloys are of the order of 700 or $800^{O}K$. For $0 = 800^{O}K$

$m = m_B = 10^{-20}$ erg/gauss, $nm_B = \frac{10^4}{4\pi}$ for saturation and $k = 1.38 \times 10^{-6}$ erg/degree, the value of γ in gaussian units is found to be

$$= \frac{3k}{4\pi nm^2}\theta = \frac{3k}{4\pi nm_B^2}\theta = 4.1$$

$$= 4.1 \times 800 = 3300 \text{ (approx)}.$$

showing the enormous interaction strengths we must assume to explain ferromagnetism. The term ferromagnetism was coined originally to described the forrit-type ferromagnetic spin order and by extension the term covers almost any compound in which some ions have a mount antiparallel to other ions. Many ferromagnetics are poor conditions of electricity, a quality expolited in device applications.

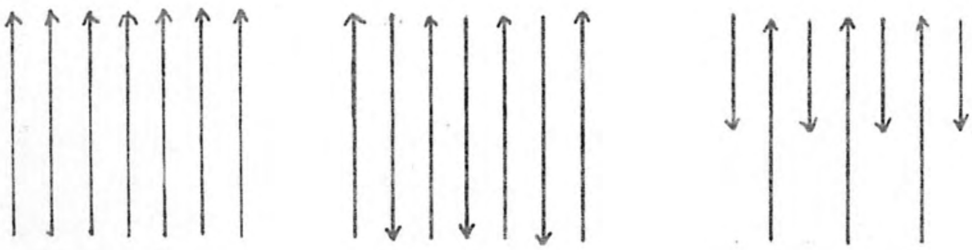
A ferromagnetic has a spontaneous magnetic moment a magnetic moment even in zero applied magnetic field. A spontaneous moment suggests that electron spins and magnetic moments are arranged in a regular manner. The order need not be simple all of the spin arrangements sketch in fig. 2 The susceptibility of a ferromagnete above the curie temperature has the form.

$$X = \frac{C}{(T-T_c)} \qquad \text{(Curic - Weiss laws)}$$

in the mean field approximation.

The elementary excitations in a ferromagnetic are magnons. Their dispersion relation for $Ka<<1$ has the form $h\,\omega \simeq jK^2a^2$ in zero external magnetic field. The thermal excitation of magnons leads at low temperatures to a heat capacity and to a fractional magnetization change both proportional to $T^{3/2}$. The origin of the internal field in ferromagnetic materials was due to the exchange interaction that lined up neighboring spin moments in the solid. In some salts of the transition metals that contain ferromagnetic elements and that crystallize in a certain structure called the spinel structure, the spacing

ORDERED ARRANGEMENTS OF ELECTRON SPINS



SAMPLE FERROMAGNET   SAMPLE ANTIFERROMEGNET   FERRIMAGNET

F. 2

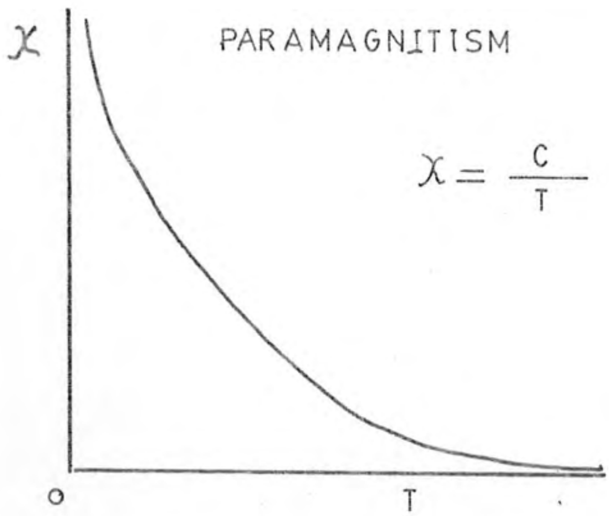
CANTED ANTIFERROMEGNET   HELICAL SPIN ARRAY of the atoms is such that the exchange interaction is negative. Below a critical temperature, called the neel temperature T_N, this can lead in such structures O anti-parallel alignment of electron spins in neighboring atoms. If the two spin systems are balanced, the solid belongs to the antiferromagnetic class i.e. salts of transition elements and, if unbalanced, to the ferrimagnetic class; Figure shows the spin arrangements characteristic of ferromagnetism, antiferromagnetism, and ferrimagnetism, the temperature in each being below the critical temperature for spin disordering.

A typical antiferromagnetic solid is MnO. Figure shows the magnetic susceptibility of MnO plotted as a function of temperature. The maximum is characteristic of antiferromagnetic behavior. In addition, above the temperature of the maximum, r varies as C/(T+0).

TABLE

Type	Magnitude of susceptibility	Temperature dependence	Examples
Diamagnetic	Small, negative intermediate, negative	Independent varies with field and temperature below $20^{o}K$(not discussed in text)	Organic materials, Light element, Alkali earth Bismuth
	Large, negative	Esists only below critical temperatures (see Section)	Superconducting metals
Paramagnetic	Small, positve	Independent	Alkali metals Transition metals
	Large, positive	$X = \frac{C}{T-0}$	Rare earth
Antiferromagnetic	Small, positive	When $T > T_N$; $X = \frac{C}{T+0}$ When $T < T_N$; $X \alpha T$	Salts of transition elements
Ferromagnetic	Very large, positive	$T > T_c$; $X = \frac{C}{T-0}$ $T < T_c$;	Some transition and rare earth metals
Ferrimagnetic	Very large, positive	$T < T_N$; $X = \frac{C}{T\pm 0}$ $T < T_N$	Ferrites

PARAMAGNITISM

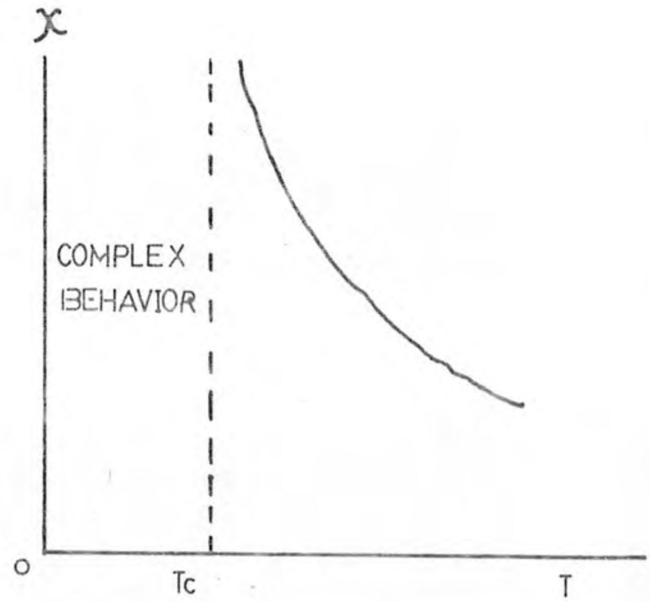


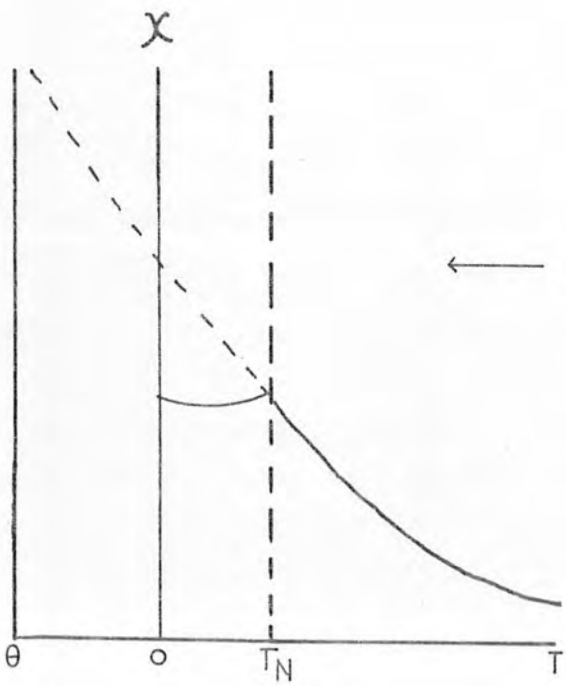
$$\chi = \frac{C}{T}$$ CURIE LAW

FERROMAGNETISM ⟶

$$\chi = \frac{C}{T - T_c} \qquad (T > T_c)$$

CURIE-WEISS LAW





⟵ ANTIFRROMAGNETISM

$$\chi = \frac{C}{T + \theta} \qquad (T > 0)$$

Table *The Electronic Structure of Atoms*

ATOMIC NUMBER	ELEMENT SYMBOL	NUMBER OF ELECTRONS K	L		M			N				IONIZATION ENERGY, eV	ATOMIC RADIUS, Å	MAGNETIC SUSCEPTIBILITY
		1*s*	2*s*	2*p*	3*s*	3*p*	3*d*	4*s*	4*p*	4*d*	4*f*			
1	H	1										13.53	0.53	-1.97×10^{-6}
2	He	2										24.47	0.30	−0.47
3	Li	2	1									5.37	1.50	+0.50
4	Be	2	2									9.28	1.19	−1.00
5	B	2	2	1								8.25	0.85	−0.69
6	C	2	2	2								11.20	0.66	−0.49
7	N	2	2	3								14.47	0.53	−0.8
8	O	2	2	4								13.55	0.45	+106.2
9	F	2	2	5								18.6	0.38	
10	Ne	2	2	6								21.47	0.32	−0.33
11	Na	2	2	6	1							5.12	1.55	+0.51
12	Mg	2	2	6	2							7.61	1.32	+0.55
13	Al	2	2	6	2	1						5.96	1.21	+0.65
14	Si	2	2	6	2	2						8.08	1.06	−0.13
15	P	2	2	6	2	3						11.11	0.92	−0.90
16	S	2	2	6	2	4						10.31	0.82	−0.49
17	Cl	2	2	6	2	5						12.96	0.75	−0.57
18	A	2	2	6	2	6						15.69	0.67	−0.48
19	K	2	2	6	2	6	—	1				4.32	2.20	+0.52
20	Ca	2	2	6	2	6	—	2				6.09	2.03	+1.10
21	Sc	2	2	6	2	6	1	2				6.7	1.80	
22	Ti	2	2	6	2	6	2	2				6.81	1.66	+1.25
23	V	2	2	6	2	6	3	2				6.76	1.52	+1.4
24	Cr	2	2	6	2	6	5	1				6.74	1.41	+3.08
25	Mn	2	2	6	2	6	5	2				7.40	1.31	+11.8
26	Fe	2	2	6	2	6	6	2				7.83	1.22	Ferromag.
27	Co	2	2	6	2	6	7	2				8.5	1.14	Ferromag.
28	Ni	2	2	6	2	6	8	2				7.61	1.07	Ferromag.
29	Cu	2	2	6	2	6	10	1				7.68	1.03	−0.086
30	Zn	2	2	6	2	6	10	2				9.36	0.97	−0.157
31	Ga	2	2	6	2	6	10	2	1			5.97	1.13	−0.24
32	Ge	2	2	6	2	6	10	2	2			8.09	1.06	−0.12
33	As	2	2	6	2	6	10	2	3			10.5	1.01	−0.31
34	Se	2	2	6	2	6	10	2	4			9.70	0.95	−0.32
35	Br	2	2	6	2	6	10	2	5			11.30	0.90	−0.39
36	Kr	2	2	6	2	6	10	2	6	—	—	13.94	0.86	−0.35

(N.B. The magnetic susceptibility values in this table are in cgs units. To convert to mks units, multiply the cgs value by 4π.)

Table (*continued*)

ATOMIC NUMBER	ELEMENT SYMBOL	NUMBER OF ELECTRONS N				O					P	IONIZATION ENERGY, eV	ATOMIC RADIUS, Å	MAGNETIC SUSCEPTIBILITY
		4s	4p	4d	4f	5s	5p	5d	5f	5g	6s			
37	Rb	2	6	—	—	1	—	—	—	—	—	4.16		$+0.21 \times 10^{-6}$
38	Sr	2	6	—	—	2	—	—	—	—	—	5.67		-0.20
39	Y	2	6	1	—	2						6.5		$+5.3$
40	Zr	2	6	2	—	2						6.92		-0.45
41	Nb	2	6	4	—	1						6.8		$+1.5$
42	Mo	2	6	5	—	1						7.06		$+0.04$
43	Tc	2	6	6	—	1						7.1		
44	Ru	2	6	7	—	1						7.7		$+0.50$
45	Rh	2	6	8	—	1						7.7		$+1.11$
46	Pd	2	6	10	—	—						$8.\dot{3}$		$+5.4$
47	Ag	2	6	10	—	1						7.54		-0.20
48	Cd	2	6	10	—	2						8.96		-0.18
49	In	2	6	10	—	2	1					5.76		-0.11
50	Sn	2	6	10	—	2	2					7.30		-0.25
51	Sb	2	6	10	—	2	3					8.35		-0.87
52	Te	2	6	10	—	2	4					8.96		-0.31
53	I	2	6	10	—	2	5					10.44		-0.36
54	Xe	2	6	10	—	2	6					12.08		-0.34
55	Cs	2	6	10	—	2	6				1	3.87		-0.22
56	Ba	2	6	10	—	2	6				2	5.19		$+0.9$
57	La	2	6	10	—	2	6	1			2	5.59		$+1.04$
58	Ce	2	6	10	2	2	6	—			2	6.54		$+15.0$
59	Pr	2	6	10	3	2	6	—			2	5.8		$+25.0$
60	Nd	2	6	10	4	2	6	—			2	6.3		$+36.0$
61	Pm	2	6	10	5	2	6	—			2	6.3		
62	Sm	2	6	10	6	2	6	—			2	6.6		
63	Eu	2	6	10	7	2	6	—			2	5.64		$+22.0$
64	Gd	2	6	10	7	2	6	1			2	6.7		Ferromag.
65	Tb	2	6	10	8	2	6	1			2	6.7		
66	Dy	2	6	10	9	2	6	1			2	6.8		Ferromag.
67	Ho	2	6	10	10	2	6	1			2			
68	Er	2	6	10	11	2	6	1			2			
69	Tm	2	6	10	12	2	6	1			2			
70	Yb	2	6	10	13	2	6	1			2	6.2		
71	Lu	2	6	10	14	2	6	1			2	5.0		
72	Hf	2	6	10	14	2	6	2	—	—	2	5.5		

(N.B. The magnetic susceptibility values in this table are in cgs units. To convert to mks units, multiply the cgs value by 4π.)

CHAPTER TWO

MUTUAL INDUCTANCE

If magnetic flux from one circuit threads the current path of another circuit, changing current in the first circuit influences the current in the second circuit. The quantitative aspects of this mutual interaction are best discussed in terms of a purely geometric quantity called the mutual inductance. The mutual inductance M gives the magnitude of the e.m.f. induced in one circuit in the other. The e.m.f. Σ_2 induced in circuit '2' by a changing current 2, in circuit 'i'. When the circuits do not move is given by

$$\Sigma_2 = - M_{21} \frac{di}{dt} \qquad (1)$$

The mutual inductance between two circuits is the same, regardless of which circuit is considered the inducing agent and which circuit acted upon. Reversing the argument in eq.(1) we can write

$$\Sigma_1 = - M_{12} \frac{di_2}{dt} \qquad (2)$$

It can be shown that

$$M_{12} = M_{21} = M(\text{say}) \qquad (3)$$

The MKS unit of mutual inductance is henry similar to self inductance and is named after the American scientist Joseph Henry, who developed the idea of inductance almost simultaneously with Faraday. The mutual inductance of circuits is one henry if the current changing at the rate of 1 amp/sec. in one circuit induces an e.m.f. of one volts in the other. In the light of eq.(3) we can write

$$\Sigma = - M \frac{di}{dt}$$

which yields

$$M = \frac{\Sigma}{di/dt} \quad \text{numerically}$$

$$M(\text{henries}) = \frac{\Sigma(\text{volts})}{di/dt(\text{amp/sec}}$$

$$= \frac{\Sigma}{di/dt} \quad \text{volts-sec/amp}$$

In gaussian units one henry is equal to 10^9 emu.

The mutual inductance M is the geometric parameter which describes the inductive of one circuit on the other-similar to self-inductance we can calculate M from the relation

$$M = \frac{d\Phi}{di} \tag{4}$$

Calculation of M between two coaxial solenoids consider two identical solenoidal coils each of length l and number of turn N_1 and N_2 wound one upon the other on the same core but perfectly insulated from each other. A current i_1 in the first coil produces a magnetic induction.

$$B = \frac{\mu_o N_1 i_1}{l}$$

and the flux, linking its each turn is

$$\Phi_{11} = BA = \frac{\mu_o N i_1}{l} \qquad (5)$$

Total flux linking the N_1 turns in

$$\Phi_{11} = N_1 \quad \Phi_{11} = \frac{\mu_o \; N_1^2 i_1 A}{l} \qquad (6)$$

Total flux linking the N_2 turns of the second coil is

$$\Phi_{21} = N_2 \quad \Phi_{11} = \frac{\mu_o N_1 N_2 i_1 A}{l} \qquad (7)$$

From these fluxes it follows that

$$L_1 = \frac{d\Phi_{11}}{di_1} = \frac{\mu_o N_1^2 A}{l} \qquad (8)$$

as before, and

$$M_{21} = \frac{d\Phi_{21}}{di_1}$$

$$M_{21} = \frac{\mu_o N_1 N_2 A}{l} \qquad (9)$$

Reversing the procedure and considering a current i_2 in the second coil gives

$$L_2 = \frac{d\Phi_{22}}{di_1} = \frac{\mu_o N_2^2 \; A}{l} \qquad (10)$$

and

$$M_{12} = \frac{d\Phi_{12}}{di_2} = \frac{\mu_o N_1 N_2 A}{l} \qquad (11)$$

thus justifying our conviction that $M_{12} = M_{21} = M$ (say) from eqs. 8,9,10 and 11 we can write

$$\left(\frac{\mu_o N_1 N_2 A}{l}\right) = \frac{\mu_o N_1^2 A}{l} \times \frac{\mu_o N_2^2 A}{l}$$

or

$$M = \sqrt{L_1 L_2} \qquad (12)$$

This relation represents a limit that is imposed on the mutual inductance between two circuits, viz, it is always less than or equal to the square root of the product of the self-inductance of the two circuits. In view of this limit, a coupling coefficient K is often introduced and defined by

$$M = K\sqrt{L_1 L_2} \qquad |K|>1 \qquad (13)$$

If K = 1 then all the flux of one circuit links the other circuit. In addition to the induced e.m.f.s in each circuit induced e.m.f. due to its own current will also be present

$$\Sigma_1' = L_1 \frac{di_1}{dt}$$

and

$$\Sigma_2'' = L_1 \frac{di_2}{dt}$$

CHAPTER THREE

BRIDGE:

A bridge is an instrument, or an intermediate means in a measurement, system, that or all of a bridge circuit, and by means of which one or more of the electrical constants of a bridge circuit may be measured. The operation of a bridge consists of the insertion of a suitable electromotive force and a suitable detecting device in branches that can be made conjugate and that do not include the branch whose constants are to be measured, followed by the adjustment of one or more of the branches until the response of the detecting device becomes zero or an amount measurable by the detector for the purpose of interpolation.

CAMPBELL BRIDGE:

A form of mutual-inductance bridge, used for the comparison of mutual inductances. It was designed by Campbell in 1910. One of the mutual inductances is a standard for the calibration of the other. Two balances are required: (1) with the detector at 1-1 (Fig.3-1).

$$L_1/L_2 = R_1/R_2 = R_4/R_3$$

(2) with the detector at 2-2 (maintaining all bridge elements constant except M_1 and varying M_1 for balance),

$$R_1/R_2 = R_4/R_3 = L_1/L_2 = M_1/M_2$$

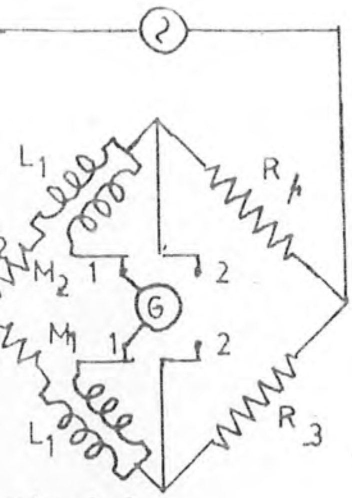


Fig.3.1

CAMPBELL BRIDGE

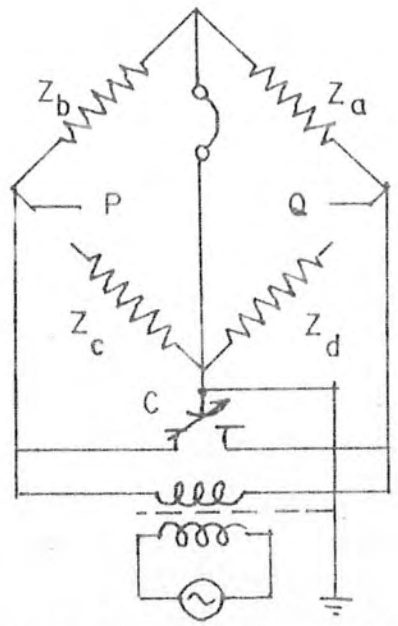
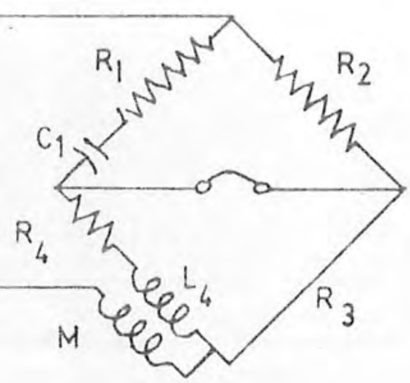


Fig.3.2 CARACITANCE BRIDGE



.3.3 CAREY FOSTER BRIDGE

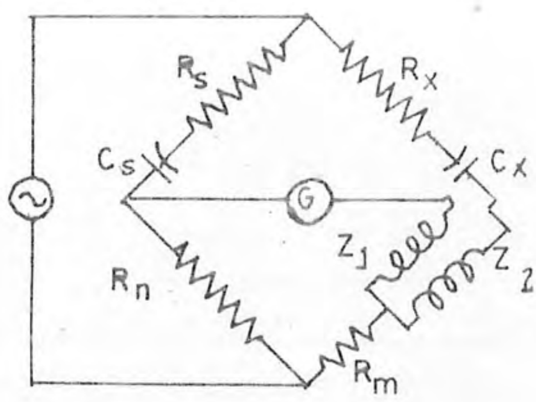


Fig.3.4 DAWES-HOOVER BRIDGE

CAPACITANCE BALANCE BRIDGE:

A bridge useful for measuring an unknown impedance Z_x when one point of the bridge must be grounded directly. The bridge is first balanced with the arms P and Q closed (Fig.3.2) by means of the variable capacitor C. Then P and Q are opened, and the bridge rebalanced. P and Q are then again closed, and the balance checked.

CAREY-FOSTER BRIDGE(Heydweiller)

A form of the mutual impedance bridge, originally designed by Carey-Foster in 1887, for the comparison of mutual inductance by the ballistic method. It was modified for a-c use in 1894 by Heydweiller. At balance

$$M = C_1 R_2 R_4$$

$$R_4 = M \left(1+\frac{R_1}{R_2}\right) = C_1 R_4 (R_1+R_2)$$

The circuit is shown in figure 3.3

DAWES-HOOVER BRIDGE

A bridge of the Heaviside type, adapted for work on high high-voltage cables at power distribution frequencies. The circuit diagram of the bridge is shown in Fig.3.4. R_x and C_x are the capacitance and series loss resistance of the cable. R_s and C_s are the capacitance and series loss of a standard capacitor. By substituting various values of variable mutual inductors Z_1 and Z_2, a very wide range of power factors may be measured, ranging from 0.1% to 80%, which is required when the insulation of the high voltage cables is studied at high temperatures or voltages beyond the ionization level. The balance conditions are

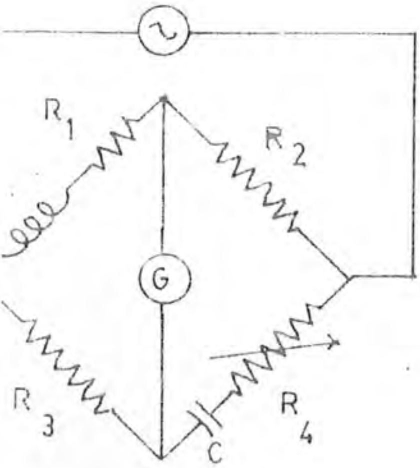
$$R_x \cong M/R_n C_s; \qquad C_x = R_n C_s/R_m$$

and the power factor, $PF = \omega M/R_m$.

HAY BRIDGE

The Hay bridge, normally used for the measurement of inductance in terms of capacitance, resistance, and frequency, is a four-arm a-c bridge, in which the arms adjacent to the unknown impedance are non-reactive resistor in series with a resistor, shown in Fig.3.5. Usually the bridge is balanced by adjustment of the resistor, which is in series with the capacitor, and of one of the non-reactive arms. The balance depends upon the frequency. It differs from the Maxwell bridge in that in the arm opposite the inductor, the capacitor is in series with the resistor. The balance equations are

$$L_1 = R_2 R_3 \frac{C}{1+\omega^2 C^2 R_4^2} \qquad R_1 = R_2 R_3 \frac{\omega^2 C^2 R_4}{(1+\omega^2 C^2 R_4^2)}$$



.3.5 Hay bridge

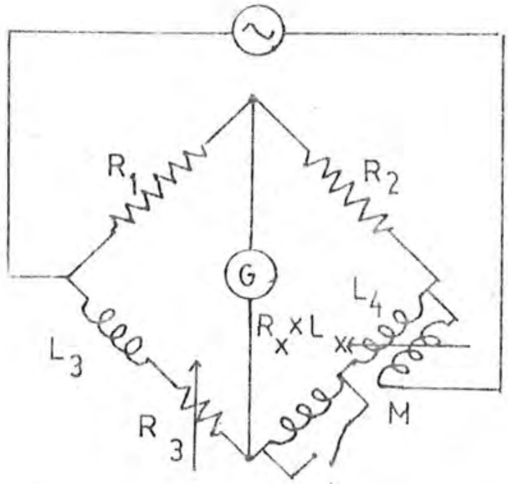
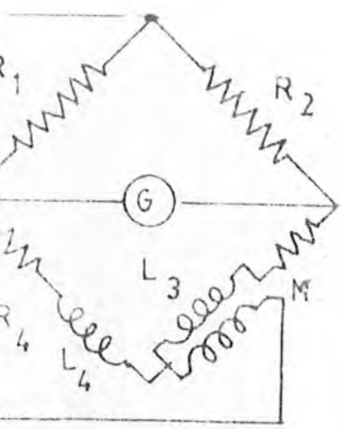


Fig.3.6A Frequency independent Heaviside-Campbell mutual-inductance bridge/



6 Heaviside-Campbell mutual-inductance bridge.

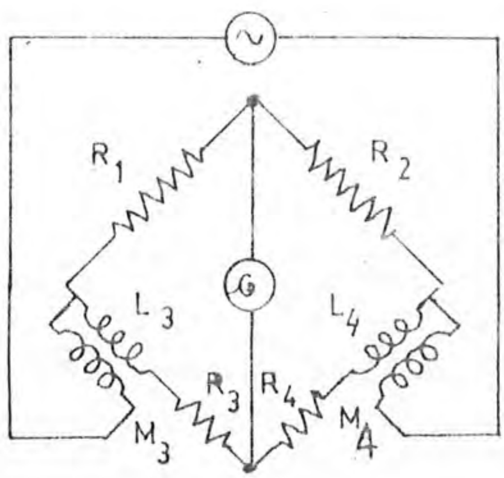


Fig.3.7 Heaviside mutual-inductance bridge.

HEAVISIDE-CAMPBELL BRIDGE

A form of the mutual-impedance bridge. It is used for the comparison of self-and mutual inductance. At balance,

$$M = \frac{R_2L_4 - R_1L_3}{R_1 + R_2}$$

If L_3 and L_4 are known, M is determined. Similarly, if M and L_4 are known, L_3 may be determined.

Another form of this bridge is shown in Fig.3.6, in which one of the inductive arms contains a separate inductor, which is included in the bridge arm during the first of a pair of measurements and is short-circuited during the second. The balance is independent of the frequency, and the balance equations are given by

$$R_x = (R_3 - R_3)\ \frac{R_2}{R_1}\ ; \qquad L_x = M-M')\ (1+ \frac{R_2}{R_1}\ ).$$

HEAVISIDE MUTUAL INDUCTANCE BRIDGE

A Heaviside mutual-inductance bridge, normally used for the comparison of self and mutual inductances, is an a-c bridge in which two adjacent arms contain self-inductance, and one or both of these have mutual inductance to the supply circuit, the other two arms being normally non-reactive resistors, as shown in figure 3.7.

The balance is independent of the frequency. The balance equations are

$$R_1R_4 = R_2R_3, \qquad L_3-L_4\left(\frac{R_1}{R_2}\right) = -(M_3-M_4)\left(1+\frac{R_1}{R_2}\right)$$

MAXWELL BRIDGE

The Maxwell bridge, normally used for the measurement of inductance (or capacitance) in terms of resistance and capacitance (or inductance), is a four-arm a-c bridge characterized by having in one arm an inductor in series with a resistor, and in the opposite arm a capacitor in parallel with a resistor, the other two arms being normally non-reactive resistors as shown in figure 3.8. The balance is independent of the frequency, and at balance the ratio of the inductance to the capacitance is equal to the product of the resistances of either pair of opposite arms. It differs from the Hay bridge in that in the arm opposite the inductor, the capacitor is shunted by the resistor. The balance equations are:

$$R_1R_4 = R_2R_3 = L/C.$$

MAXWELL INDUCTANCE BRIDGE

The Maxwell inductance bridge, normally used for the comparison of self-inductances, is a four-arm a-c bridge characterized by having inductors in two

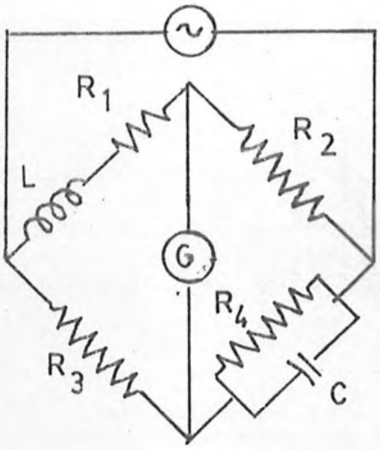


Fig. 3.8 Maxwell Bridge

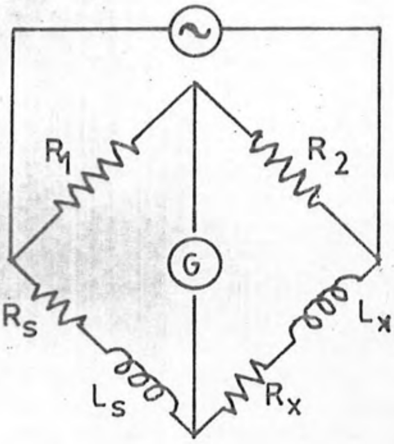


Fig.3.9 Maxwell Inductance Bridge

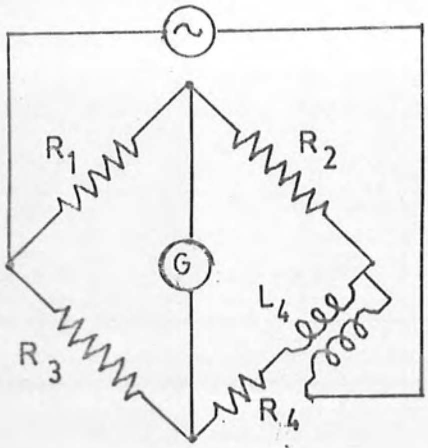


Fig.3.10 Maxwell Mutual Inductance bridge.

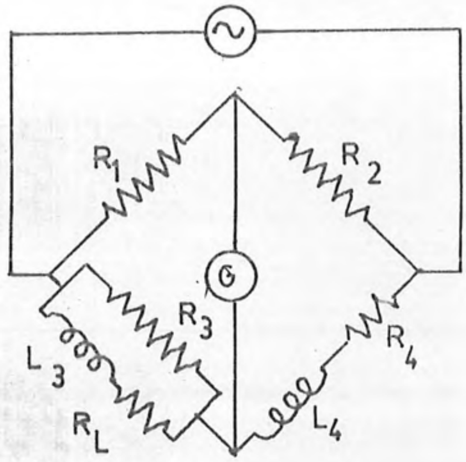


Fig.3.11 Wien Inductance Bridge

adjacent arms and, usually, non-reactive resistors in the other two arms, as shown in figure 3.9. The balance is independent of the frequency, and the equations for balance are

$$L_x = L_s \frac{R_2}{R_1}, \qquad R_x = R_3 \frac{R_2}{R_1}$$

MAXWELL MUTUAL-INDUCTANCE BRIDGE

The Maxwell mutual-inductance bridge, normally used for the measurement of mutual inductance in terms of self-inductance, is an a-c bridge characterized by the presence of mutual inductance between the supply circuit and that arm of the network which includes one coil of the mutual inductor, the other three arms being normally non-reactive resistors, as shown in figure 3.10. The balance is independent of the frequency, and the equations for balance are given by

$$R_1 R_4 = R_2 R_3, \qquad L_4 = M \left(1 + \frac{R_2}{R_1}\right.$$

WIEN INDUCTANCE BRIDGE

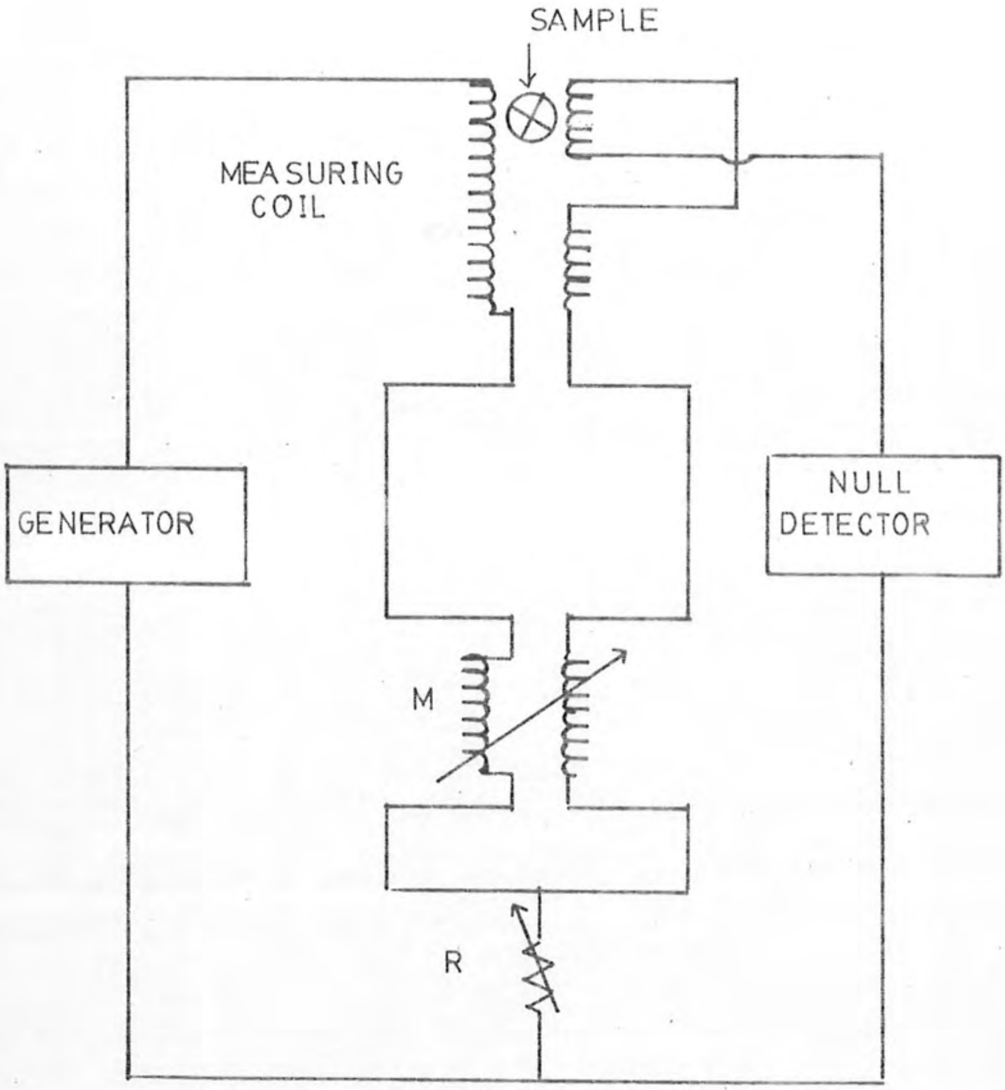
The Wien inductance bridge, normally used for the measurement of inductance in terms of resistance and frequency, is a four-arm a-c bridge characterized by having in two adjacent arms inductors respectively in series and in parallel with resistors, while the other two arms are normally non-reactive resistors figure 3.11. The balance depends upon frequency, but from the balance conditions the inductances of either or both inductors can be computed from the resistances of the four arms and the frequency. The balance equations are:

$$\frac{L_3}{L_4} = \frac{R_1(R_L+R_3)}{R_2R_3-R_1R_4}\ , \ \omega^2 L_3L_4 = R_4(R_L+R_3) - R_LR_3\frac{R_2}{R_1}$$

HARTSHORN BRIDGE

The basic Hartshorn bridge circuit is shown in figure 3.12. The measuring coil typically is placed in the cryostate and consists of two identical secondaries and a coaxial primary. The secondaries are separated along the axis of the primary. The secondaries are separated along the axis of the primary coil and are connected in opposition. In principle, the net voltage induced across the two secondaries is zero until a paramagnetic sample is inserted into one of these. The resulting induced e.m.f. may then be opposed by the voltage across the secondary of the mutual inductance M and the resistance R. At balance, M is proportional to the real part of the susceptibility of the sample and R is proportional to the power dissipation. For thermometry it is desirable to minimize the latter which is proportional to ωH^2. Where H is the amplitude of the magnetic field at the sample and ω is its frequency.

BASIC HARTSHORN BRIDGE CIRCUIT



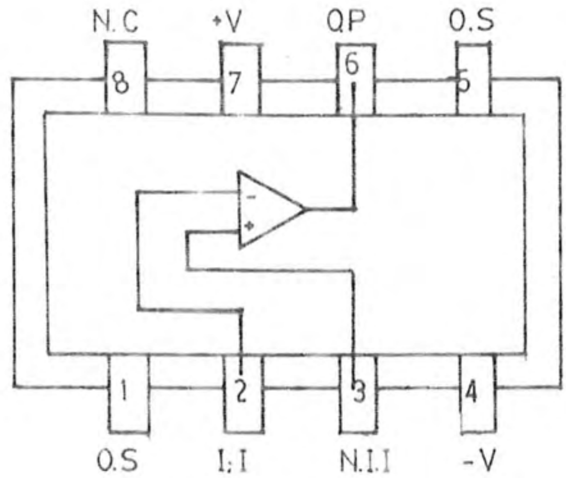
A disadvantage of Hartshorn bridge is that it requires a precise variable mutual industance which is both expensive and limited range. It is possible, however, to overcome this draw-back by simulating a variable mutual inductance electronically with an operational amplifier and fixed inductance.

CHAPTER FOUR

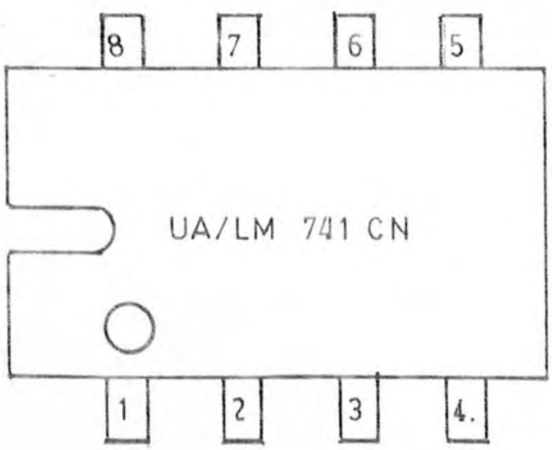
INTEGRATED AMPLIFIER: (Operational Amplifier)

Amplifier is a very important electronic circuit and in widely used in various electronic instruments. So instead of making amplifier circuits by discrete components, the whole of the amplifier circuit is integrated on a small chip of silicone and enclosed into a capsule. Pins are connected to suitable points of the amplifier as the input, out-puts terminals, ground terminal, $+V_{CC}$, $-V_{CC}$ points etc. These pins project outside the capsule as shown in the figure 4.1.

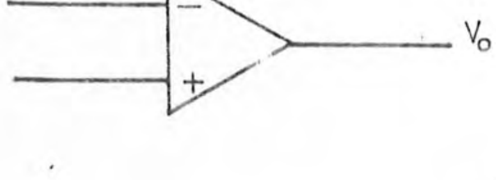
All that we have to do is to connect the signal to be amplified at the input pin, the ground pin is connected to ground, $+V_{CC}$, $-V_{CC}$ are connected to their pins, only one or two external ēlements are connected and the out-put is available at the out-put pin. Thus everything has been done by the manufacturer who also gives all the characteristics of this amplifier. All that we have to do is to just plug in. These amplifiers are becoming more popular every-day and at present it has so many applications that transistors are being replaced by them. (Hence we will learn about these amplifier and perform a few basic experiments with them). The basic part of this amplifier is a differential amplifier whose circuit is shown in figure 4.2.



4.1



4.8



4.3

It consists of two exactly identical transistors Q_1 and Q_2, emitters of which are coupled together and connected to $-V_{CC}$ through a resistance R_E. The base-emitter junction is forwarded because the emitter(n) is connected to $-V_{CC}$. Collector is reverse biased by $+ V_{CC}$. $-V_{CC}$ injects a current I_E into the emitters. Here is the electronic current.

$$V_{CC} = V_E + V_{BE}$$

$$\therefore \quad I_E = \frac{(-V_{CC}) - V_{BE}}{R_E}$$

As $\quad V_{CC} >> V_{BE}$

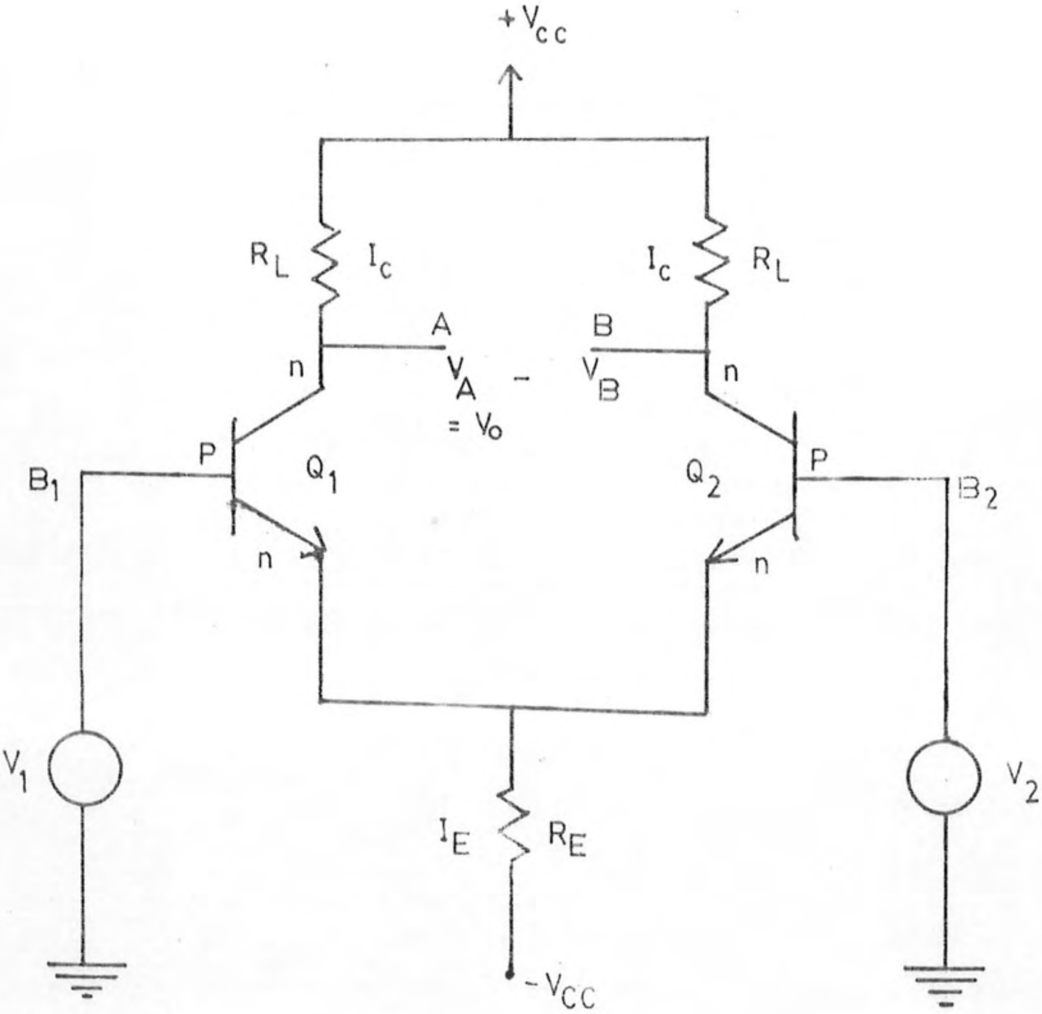
$$\therefore \quad I_E = \frac{-V_{CC}}{R_E}$$

Thus V_{CC} and R_E form a constant current generator injecting a constant current I_E into the emitters of the two transistors. The base potential V_1, V_2 of the two transistors are equal (to zero) then the current I_E will be equally divided among the two transistors. If I_{c_1} is the current through Q_1 and I_{c_2} is the through Q_2, then in this case $I_{c_1} = I_{c_2} = I_c$

$$\therefore \quad I_E = I_{c_1} + I_{c_2} = 2I_c \; .$$

collector potential of Q_1 is given by

$$V_A = V_{CC} - I_{c_1} R_L = V_{CC} - I_C R_L$$



4.2

collector potential of Q_2 is given by

$$V_B = V_{CC} - I_{C_2} R_L = V_{CC} - I_C R_L$$

In differential amplifiers the out-put is taken across the collectors of the two transistors.

$$\therefore \quad V_O = V_A - V_B = (V_{CC} - I_C R_L) - (V_{CC} - I_C R_L) =$$

Thus when base-potentials are equal i.e.

$V_1 = V_2$, then the output is zero.

Now let V_2 be zero and V_1 to be increased from 0 to V. As base of Q_1 is at a higher potential than that of Q_2 , so Q_1 will draw more current. Let $I_{C_1} = I_C + \Delta I_C$. As the total injected current I_E remains constant, so naturally I_{C_2} will decrease by

$$I_C . \qquad \therefore \quad I_{C_2} = (I_C \quad I_C)$$

Now $V_o = \{V_{CC} - (I_C + \Delta I_C) R_L\} - \{V_{CC} - (I_C - \Delta I_C) R_L\}$

$$= - 2 \Delta I_C R_L$$

Thus a signal applied at the base of Q_1, is amplified with a change of sign. Hence B_1 (base of Q_1) is known inverting terminal or - terminal because a signal applied at this terminal will be amplified with a change of sign or with a phase shift of 180^o.

Similarly it can be shown, that if $V_1 = 0$ and $V_2 = + \Delta V$, the $V_o = + 2 \Delta I_C R_L$. Thus a signal applied at B_2 i.e. base of Q_2 is amplified without any change of sign. So this terminal is known as non-inverting terminal or + terminal.

It should be noted current flowing out of the input terminals is the base current which is very very small and the resistance seen by signal sources is nearly infinity.

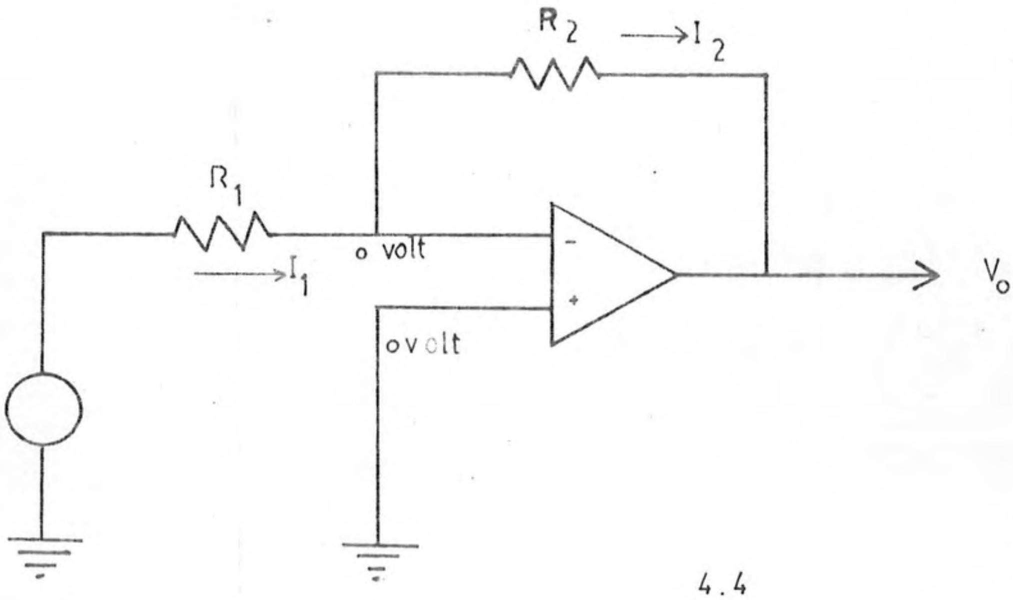
The whole amplifier is symbolically represented by fig.4.3

The amplifier has a very high gain 50,000. Infinite input resistance and practically zero out-put resistance. Integrated Amplifier as an inverting amplifier as shown in figure 4.4

INVERTING AMPLIFIER

It may be recalled that + and - terminals are actually the bases the transistor and the resistance between two terminals is practically infinity, so no current passes from + terminal to - terminal hence the potentials of these two terminals will be the same. As the terminal is at zero volts, so - terminal will also at zero volts.

$$\therefore \text{ current through } R_1 = I_1 = \frac{V_1 - 0}{R_1} = \frac{V_1}{R_2}$$



4.4

$$\therefore \text{ current through } R_2 = I_2 = \frac{0-V_o}{R_2} = -\frac{V_o}{R_2}$$

$$\text{As } I_1 = I_2 \quad \therefore \frac{V_1}{R_1} = -\frac{V_o}{R_2} \quad \therefore A = \frac{V_o}{V_1} = -\frac{R_2}{R_1}$$

Thus the gain of the amplifier entirely depends upon the resistances R_1 and R_2 and it is independent of the internal structure of the integrated amplifier.

If we want an amplifier of gain 10, then

R_1 = 1K, and R_2 = 10K are suitable resistance. Note that V_1 and V_2 are both a.c. voltages.

As the initial d.c. out-put voltage is z zero, so the output signal can swing from 0 to V_{CC}.

These amplifiers amplify d.c. as well as a-c voltages. So it is very essential that there is no d.c. voltage across the terminals of the amplifier. The gain of the amplifier is very larger. So if a small d.c. voltage exists across the two terminals, it will be amplified and the d.c. voltage at the out-put terminals will not be zero. This is known as d.c. off-set voltage. It can be selected the d.c. off-set voltage will reduce the swing of a.c. signal at the out-put. Referring to the discussion of differential amplifier, the inverting(-) and non-invarring(+) terminals of the amplifier are actually base terminals from which a small portion of the current injected at the emitter, flows out. In the circuit of the amplifier this current flowing out of - terminals sees a resistance R_1 ll R_2, but the current flowing out of + terminal comes across no resistance due to d.c. which a potential is developed across the two input terminals. This gives rise to a large d.c. offset. This offset voltage can be minimised by connecting a resistance $R = R_1$ ll R_2 between the positive terminal of the amplifier and the ground. So the circuit of non-inverting amplifier is as shown figure 4.6.

As the gain of the amplifier is 10, allowing an swing of 10 volts at the out-put, the maximum swing of V_S will be 1 volt.

Integrated amplifier as a non-inverting amplifier:

Previous argument, the potential of the negative terminal will be V_S.

$$\therefore \quad I_1 = \frac{0 - V_S}{R_1} = -\frac{V_S}{R_1}$$

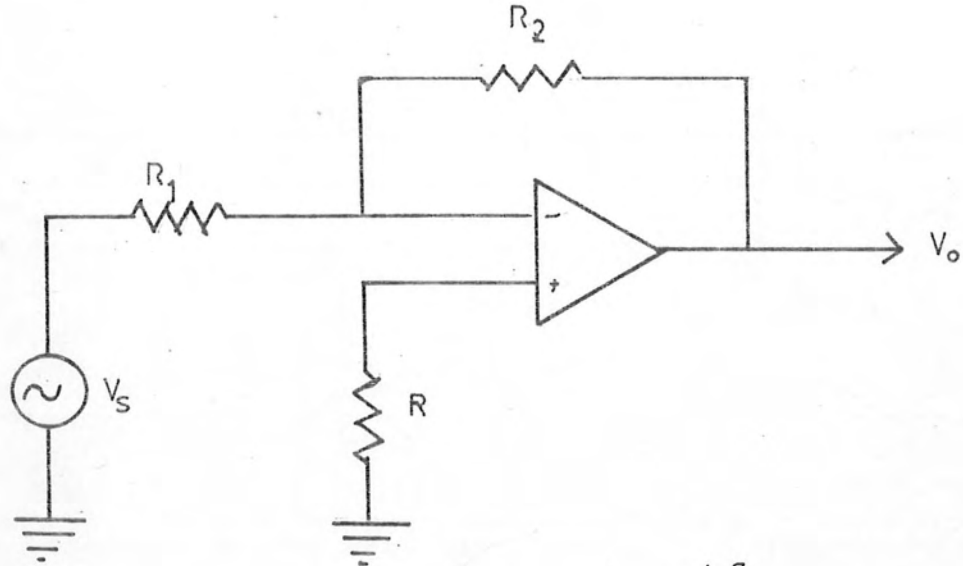
and
$$I_2 = \frac{V_S - V_o}{R_2}$$

As
$$I_1 = I_2 \quad \therefore \quad \frac{-V_S}{R_1} = \frac{V_S - V_o}{R_2}$$

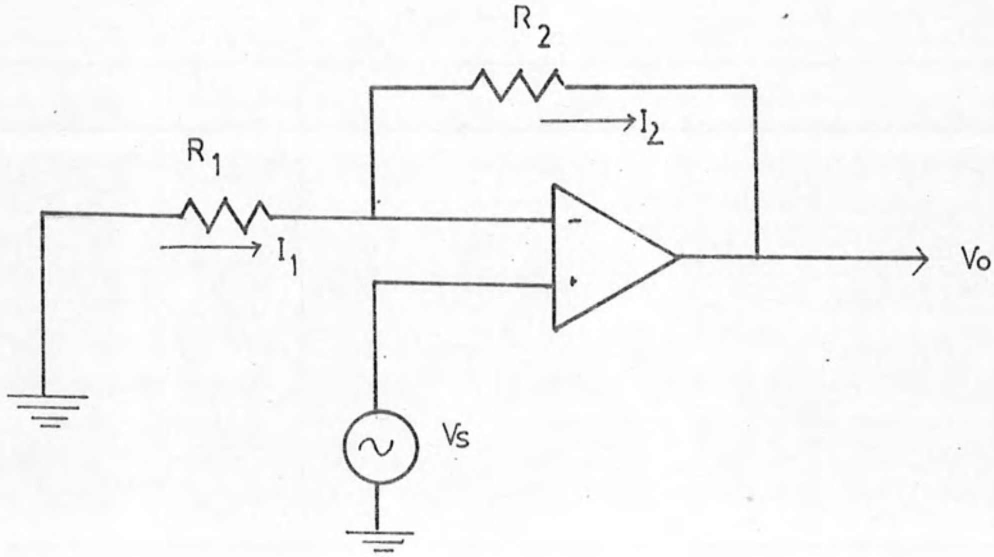
$$\text{or} \quad -V_S R_2 = R_1 V_S - V_o R_1$$

$$\text{gain} = A = \frac{V_o}{V_S} = 1 + \frac{R_2}{R_1}$$

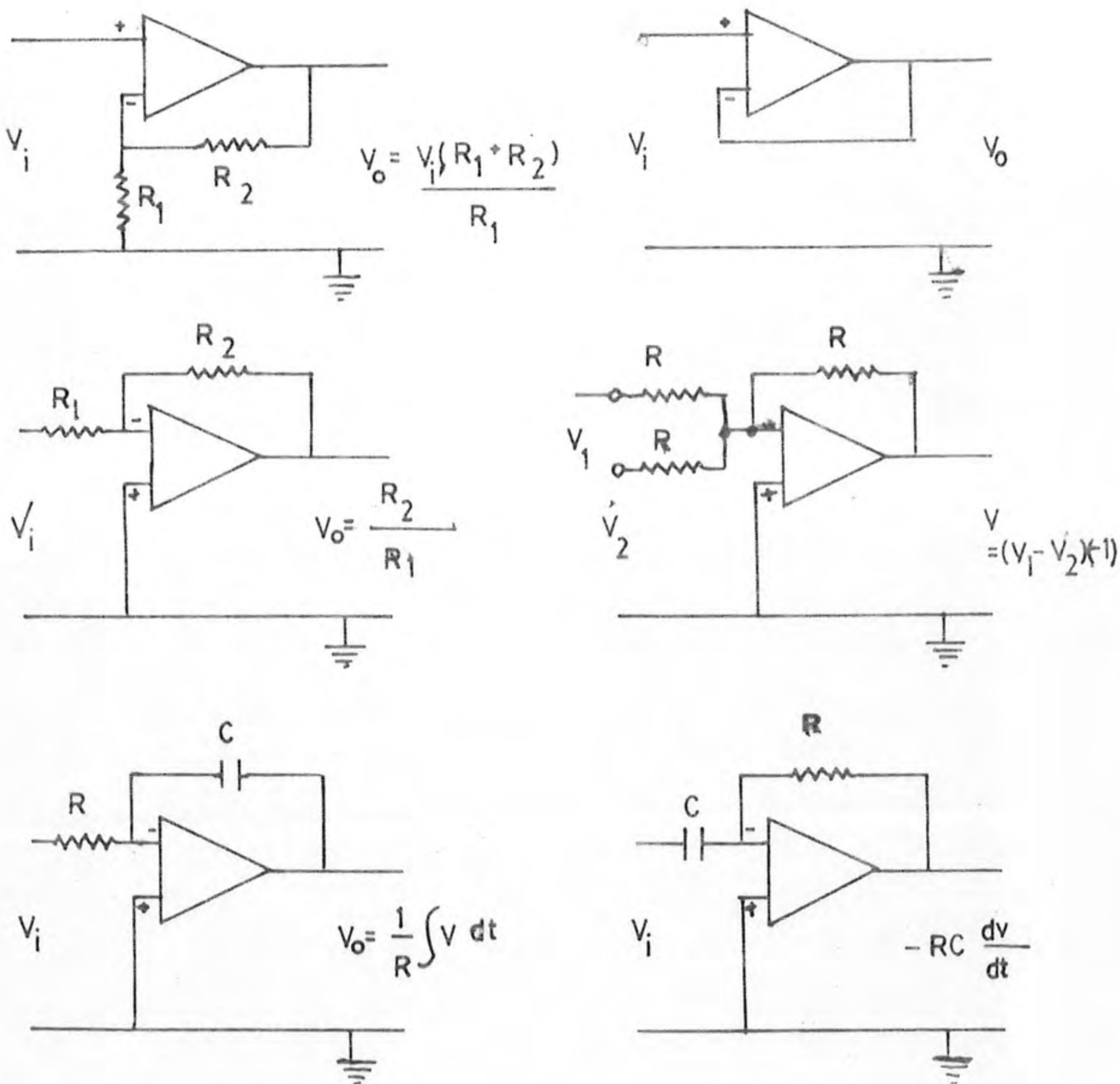
The resistance to minimise the d.c. is still connected in series its positive terminal as shown in figure 4.7



4.7



If R_S is the resistance of the generator V_S, so the additional of resistance be, connected with positive terminal is $R - R_S = R$ our case $R_S = 600$ ohms $\therefore R = 400$ ohms. The integrated amplifier is u 741 whose pin base connections are shown in figure 4.8.


Vi
R1
R2
Vo= Vi(R1+R2)/R1
Vi
Vo
R2
R1
Vi
Vo= R2/R1
R
R
V1
V2
V =(V1-V2)(-1)
C
R
Vi
Vo= 1/R ∫V dt
R
C
Vi
- RC dv/dt

CHAPTER FIVE

LOCK IN AMPLIFIER

Lock in amplifier is a synchronous detector. A dector which responds only when the input signal is synchronous which the frequency of a control signal locally generated. Such detectors are used as small indicators in bridge circuit.

Lock in Amplifier is a percision instrument crafted to permit detail analysis of signals burried in noise or other interference. The lock-in-amplifier in operation comprises three distinct section.

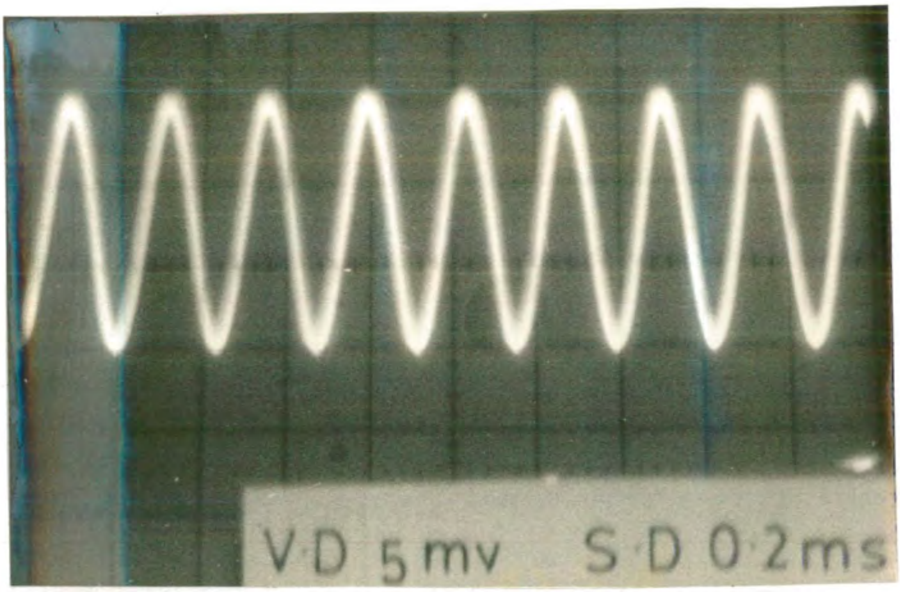
(1) The signal channel

(2) R.C.

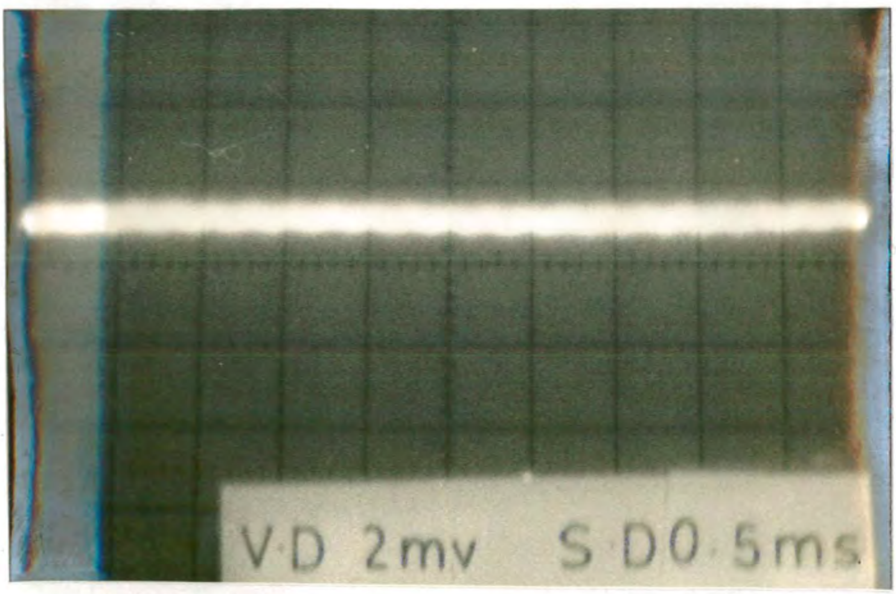
(3) PSD alongwith output its output circuitry.

The signal being measured, combined with any interference, is amplified by the signal channel which has a variable band width capable of rejecting a large amount of the interference before submitting the signal to the detector. The reference channel has the two fold function of driving the experiment so that the frequency of its signal output is known, and switching the Detector at exactly the same frequency but with an adjust able phase. Therefore, the Detector receives both a signal with its interference and a frequency reference of known phase. From these two inputs, the Detector produces sum and difference frequencies.

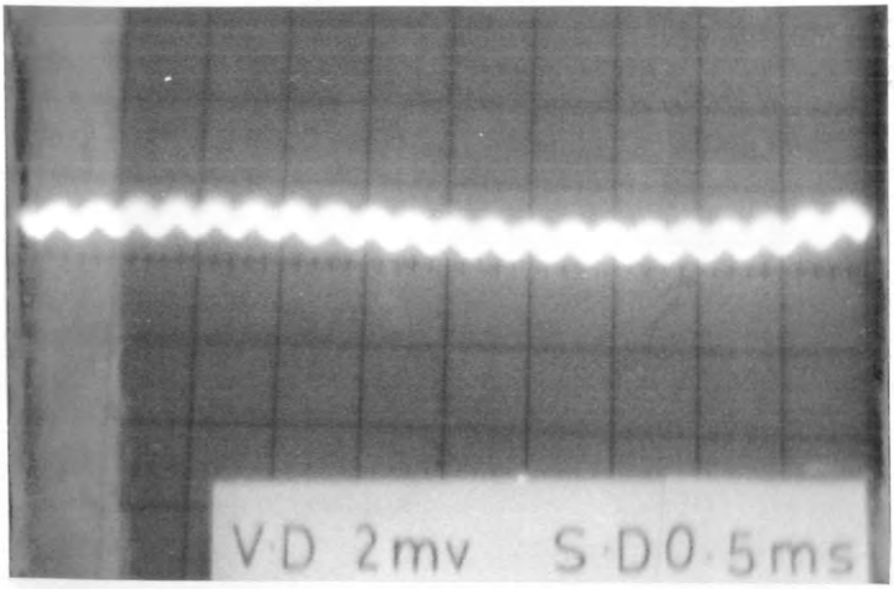
The difference between the signal and the reference is zero frequency or simply d.c., with an amplitude proportional component of the signal at the set phase. All interference and noise not at the reference appear at the Detector's output at frequencies greater than zero. By sending the total Detector output through a low pass filter, only the d.c. is present to detect the meter a calibrated amount.



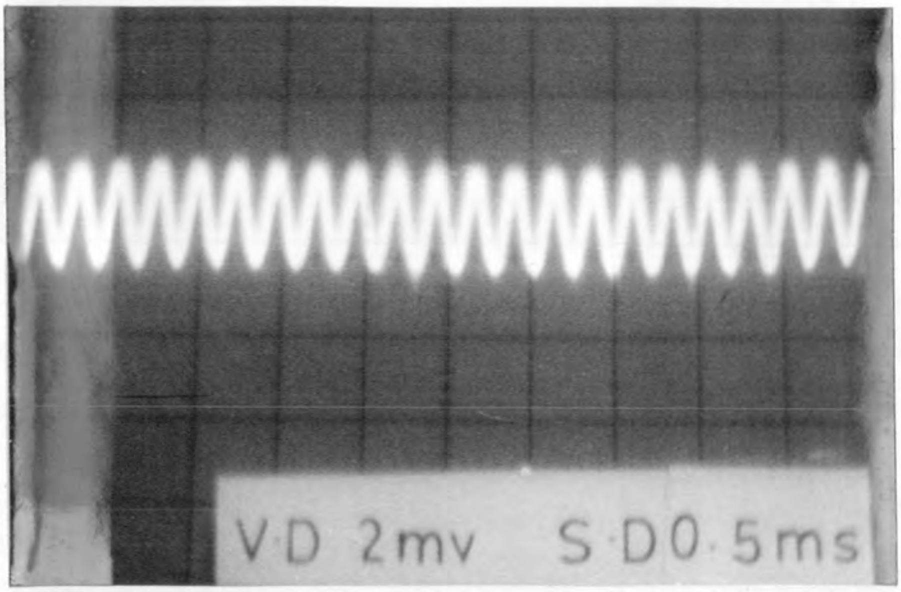
INPUT SIGNAL TO M_{1p} FROM SIGNAL GENERATOR



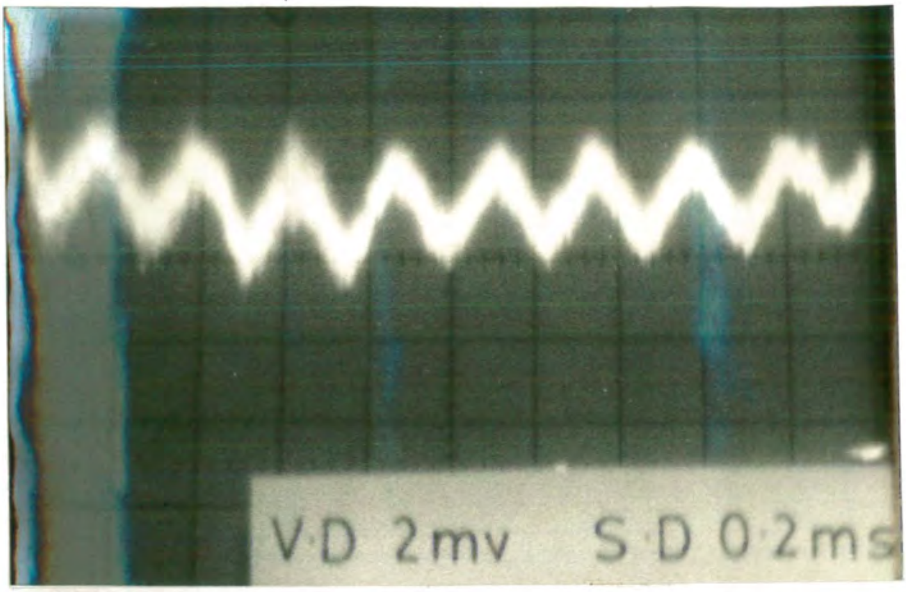
INPUT AT OPERATIONAL AMPLIFIER A_1 Pin 2 WHEN R_1 IS AT pt.1



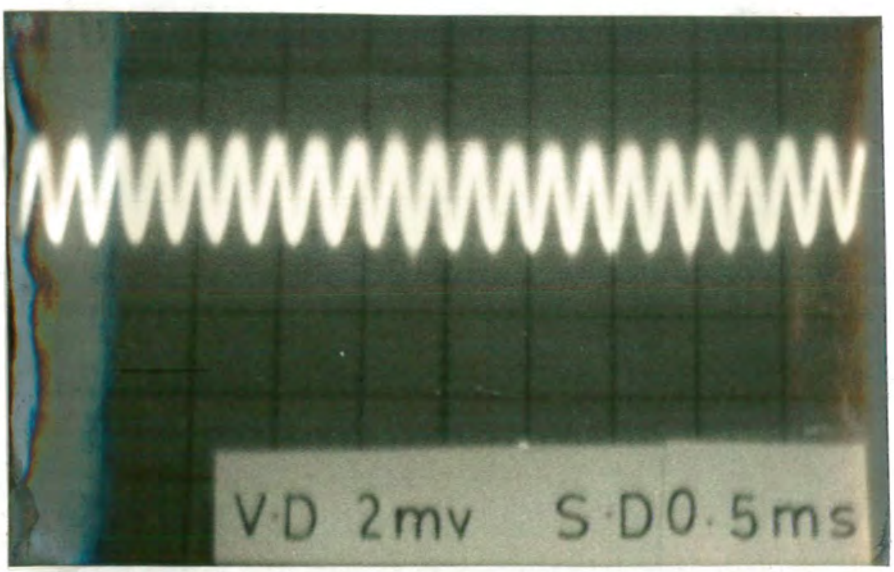
INPUT AT OPERATIONAL AMPLIFIER A_1 pin 2 WHEN R_1 IS AT POINT 3.



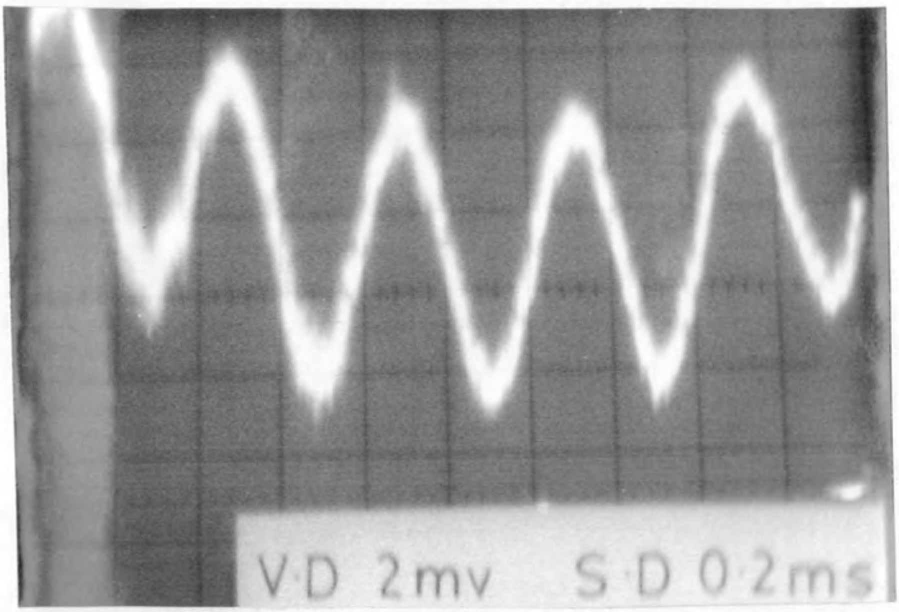
WAVE FORM AT INPUT OPERATIONAL AMPLIFIER A_2 AT pin 2.



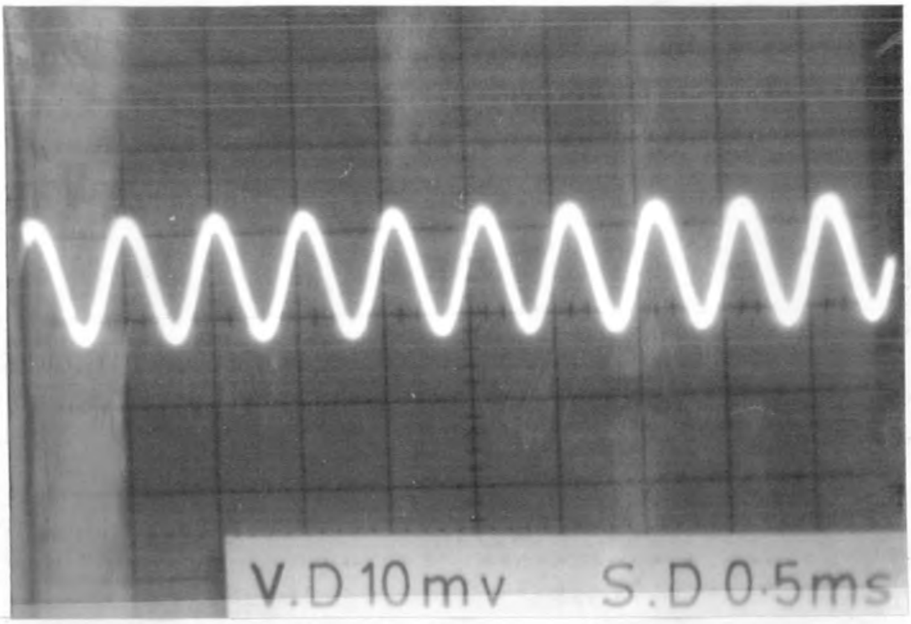
OUTPUT OF OPERATIONAL AM FIER AT pin 6 WITH NOISE



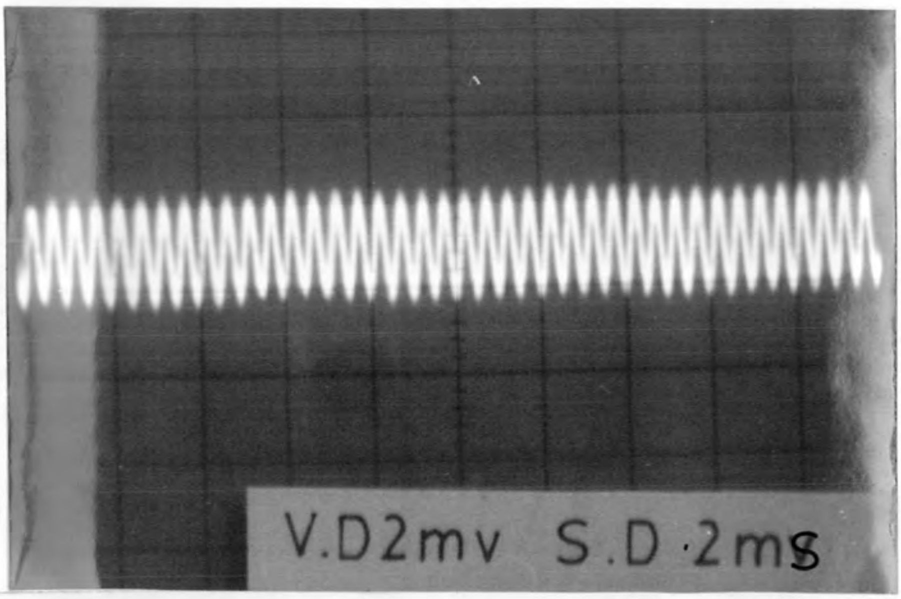
WAVE FORM INPUT OF OPERATIONAL AMPLIFIER A_2 AT pin 3.



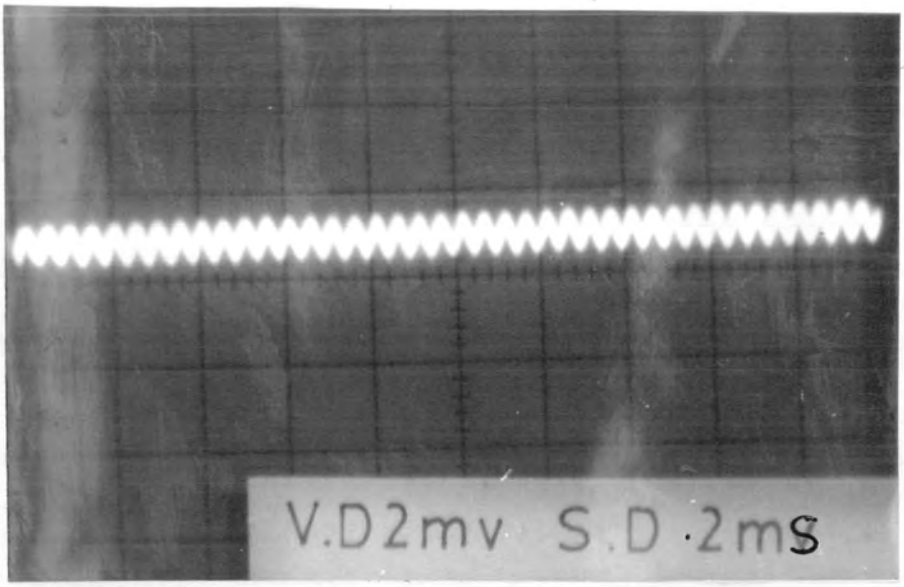
OUTPUT OF OPERATIONAL AMPLIFIER A_2 at pin 6 WITH HIGH NOISE.



INPUT SIGNAL TO M_{1p} FROM LOCK IN AMPLIFIER MODEL NO. 126.

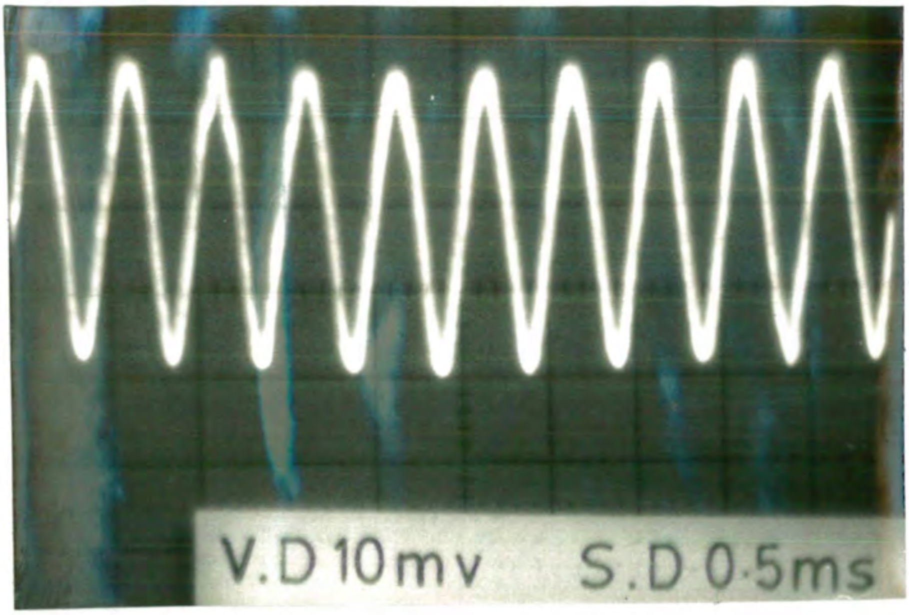


WAVEFORM AT INPUT OPERATIONAL AMPLIFIER A_2 AT pin 2.

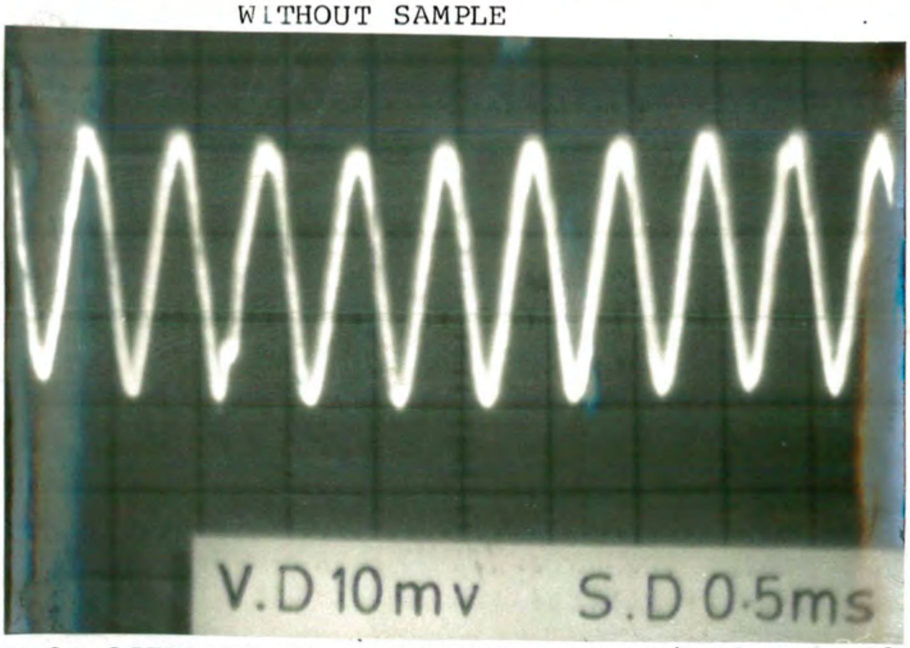


INPUT OF OPERATIONAL AMPLIFIER A_1 AT pin 3 BY USING LOCK IN AMPLIFIER

OUTPUT OF OPERATIONAL AMPLIFIER AT A_1 AT pin 6. BY USING LOCK IN AMPLIFIER



OUTPUT OF OPERATIONAL AMPLIFIER A_2 AT pin 6 WITH LOW NOISE.
WITHOUT SAMPLE

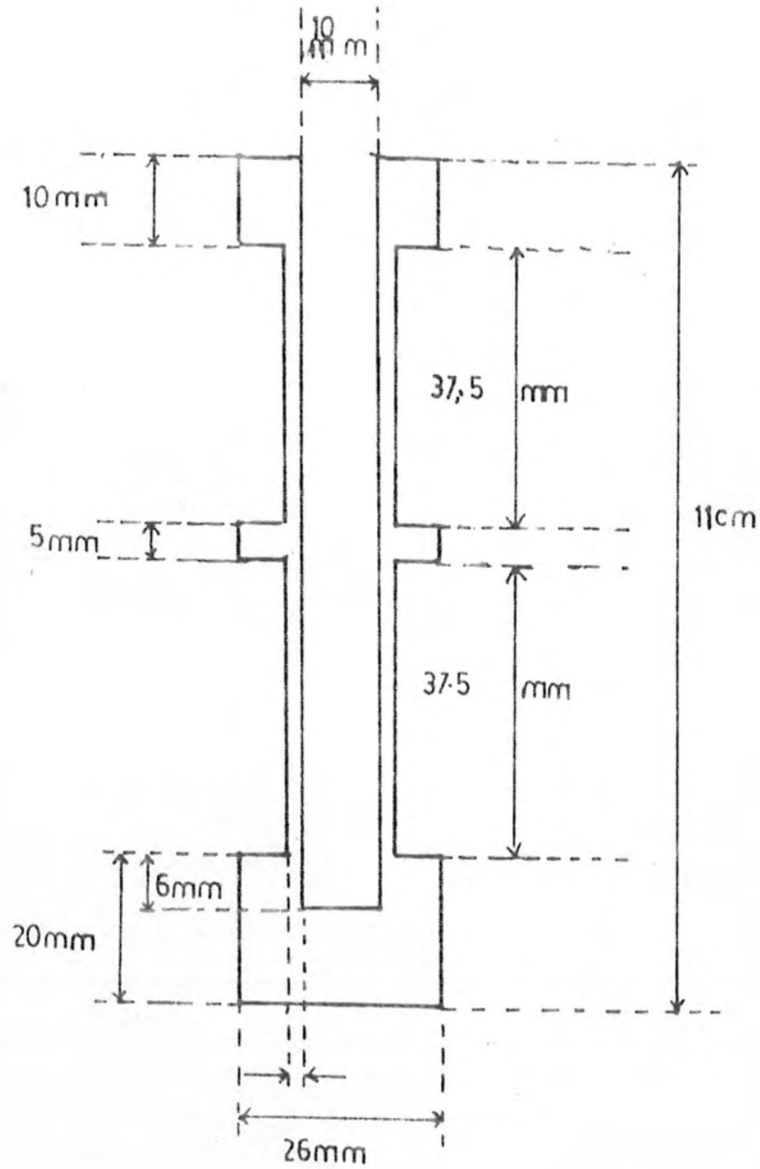


OUTPUT OF OPERATIONAL AMPLIFIER A_2 AT pin 6 WITH LOW NOISE
WITH SAMPLE

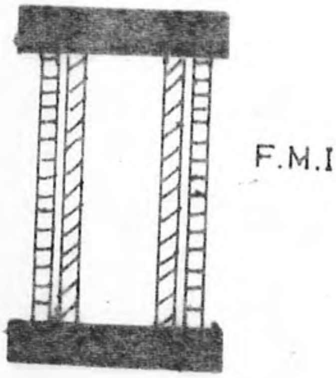
CHAPTER SIX

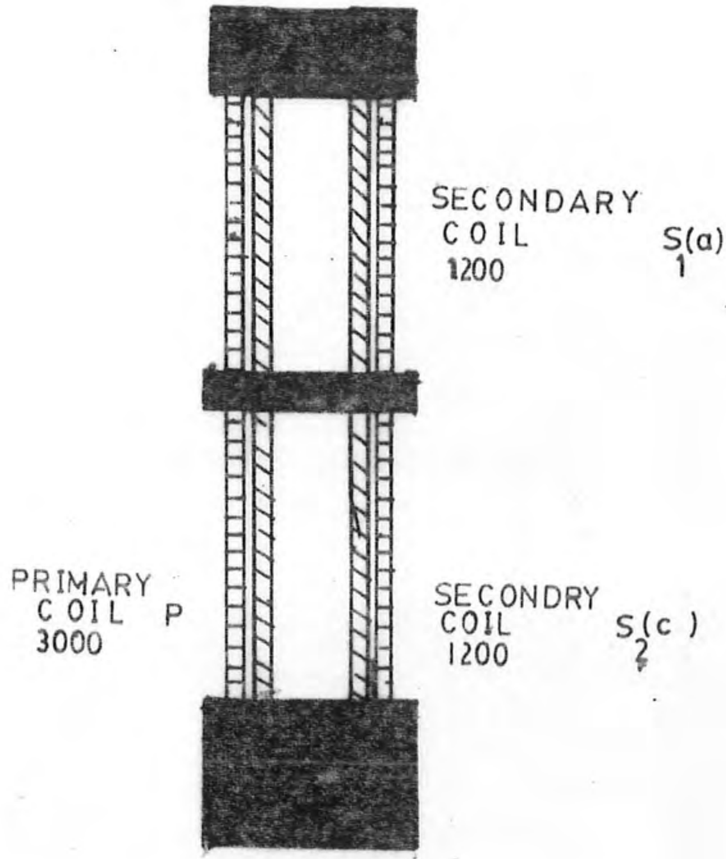
MEASURING COIL

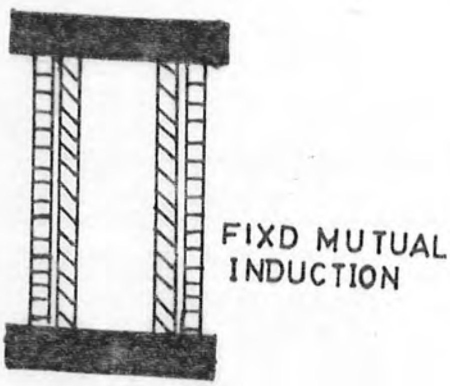
A former of abenoite material was crafted according to the dimension as shown in diagram. First of all secondary S_1 .was wound clockwise statically with the help of lathe machine by keeping its speed 50 rotations/min. In this way 1200 turn wound in 24 min. on the former. Similarly the secondary coil was wound in Anticlockwise in the same time because number of turns were same. In order to wind primary coil of 3000 turns on the both secondaries which were covered by two layers of paper tape, the same procedure was adopted. It took exactly 60 min. It also needed patience and full concentration to wind the coil layer by layer and side by side. This technique was employed due to unavailability of skilled person and proper machine. The windings are made with copper wire of diameter 0.1mm (# 36 copper wire) on the former - M_1. The coil M_2 and M_3 (turns 1:1 each) contain 100 and 750 turns respectively and made use of the same size and same diameter speed as M_1. The coils M_1, M_2 and M_3 were not worned astically.



Dimension of measuring coil former


F.M.I


SECONDARY
COIL
1200
S(a)
1
PRIMARY
COIL P
3000
SECONDRY
COIL
1200
S(c)
2


FIXD MUTUAL
INDUCTION

CHAPTER SEVEN

DESIGN AND CONSTRUCTION OF A HIGH SENSITIVITY A.C. SUSCEPTIBILITY BRIDGE

A schematic diagram of Brodbeck A.C. bridge circuit is shown in fig. The primary circuit is coupled via the transformers T(1:1) to an audio-frequency oscillator S. The sample is placed in one arm of the mutual inductor M_1 whose secondary is centertapped to ground. The mutual inductor M_2 is a coarse compensating inductance connected in series with M_1 to allow very large inductances to be measured.

The input signal for the op-amp A_1 is a portion of the primary circuit voltage determined by the potential divider R_1. The output of A_1 is linked to the secondary circuit via the fixed mutual inductor M_3. Hence, by varying R_1 the e.m.f. induced in the secondary circuit can be varied to produced an inductive balance in their circuit for different samples in M_1. The polarity of the emf induced by M_3 in the secondary circuit is reversibly by means of switch P.

A_1 is operated as a noninverting amplifier i.e. the input signal is applied between the non-inverting input and ground. In this configuration A_1 has a very high input impedance (about 10M ohms). Thus preventing current drain from the primary circuit. Gain control for A_1 is provided

by

by the variable negative feedback resistance R_2. By restricting R_2 to an upper limit of 25K ohms, noise regeneration in A_1 is kept to a minimum. A blocking capacitor at the output of A_1 minimizes dc drift.

The differential op-amp A_2 performs the function of subtracting the signal from the sample half of the secondary circuit from that of the compensating half containing M_2 and M_3. Because of the secondary windings of M_2 and M_3, the resistive elements of the two halves of the secondary circuit are normally slightly mis-matched. However, the effect of this mismatch can be made negligibly small at A_2 by balancing the input resistance and gain of A_2. The conditions for a balanced input resistance and balanced gain of A_2 are given by

$$R_a = R_i^2/(R_i + R_f)$$

$$R_b = R_i R_f/ (R_i + R_f)$$

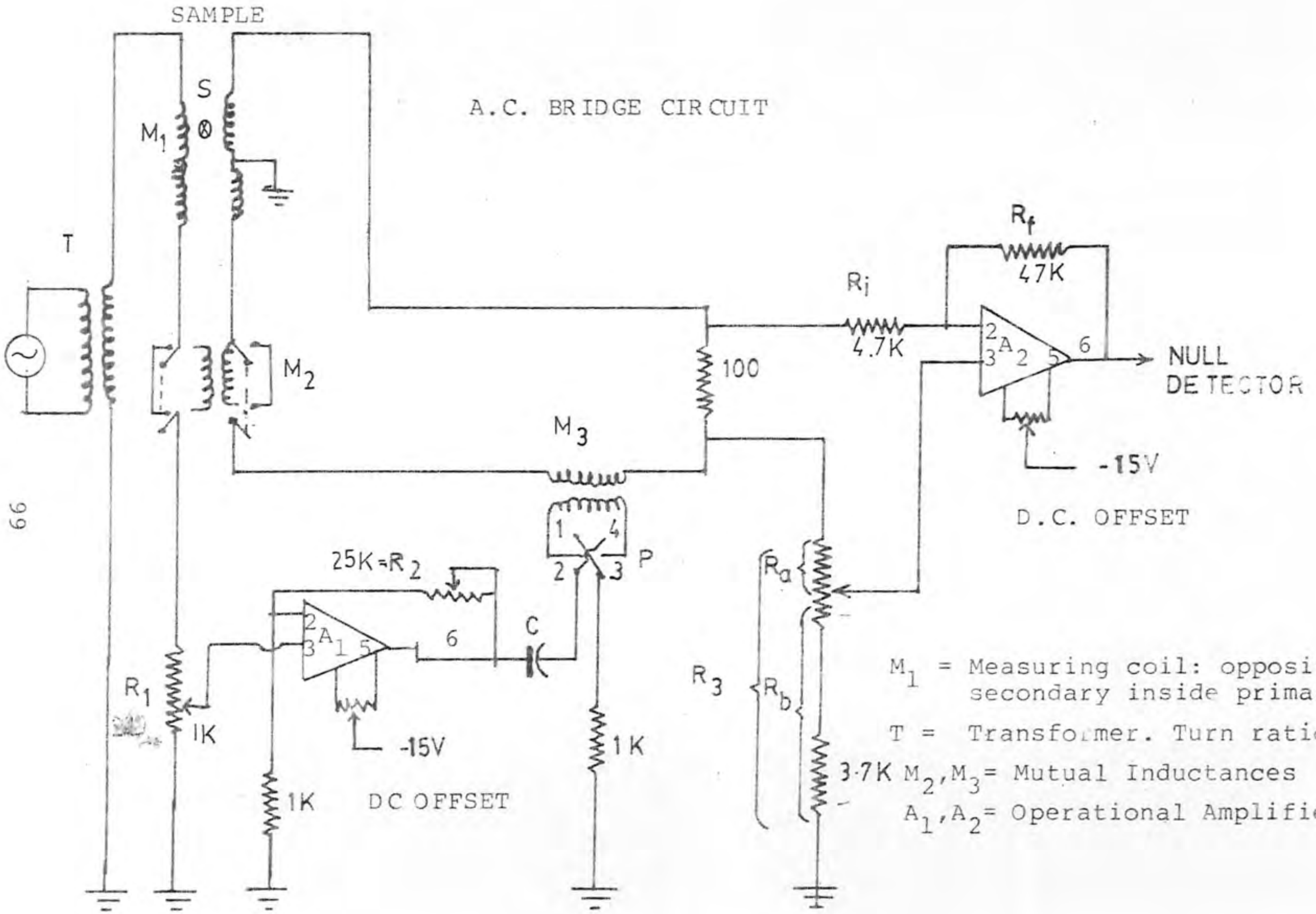
where R_a and R_b are the upper and lower portions of the potential divider R_3 respectively. R_i and R_f are the inverting input and feedback resistances of A_2 respectively. For limited gain of ten and an inverting input resistance R_i of 4.7KΩ, the balance conditions are satisfied for R_a = 0.43 K ohms and R_b = 4.7K ohms. Hence, by inserting a potential divider spanning 1K ohms at R_a, a large range of resistive losses in the secondary can be nullified at the input of A_2. In operation, the adjustment of R_3

producing a resistive balance is made initially with no sample in the circuit, and kept fixed thereafter for the suceptibility measurements.

The potential divider R_1 is a precision ten-turn potentiometer that should be equipped with a 1000 count digital dial. This allows three placd acceracy in the inductive balance adjustment. The variable portion of R_3 is a precision ten-turn potenteometer that can be locked at the correct setting for a resistive balance.

In order to calibrate the system, the resistance R_2 governing the gain of A_1 is set at a contant value with no sample in the circuit and the bridge is balanced by adjustment of R_1 and R_3 producing a minimum in the output of A_2.

Data is collected at R_1 = 900 ohms and R_3 = 360 ohms.

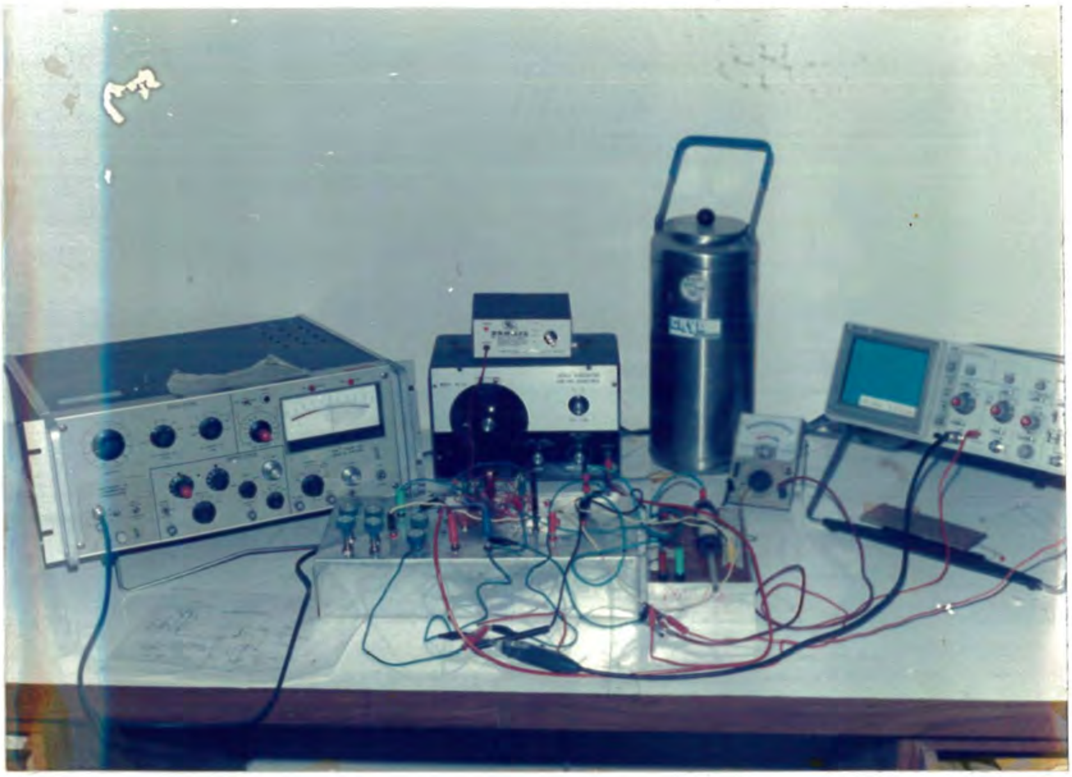


M_1 = Measuring coil: oppositly wound secondary inside primary.

T = Transformer. Turn ratio 1:1

M_2, M_3 = Mutual Inductances

A_1, A_2 = Operational Amplifier



A view of apparatus arrangement

1. Left to right lock in amplifier
2. Signal generator
3. Power supply
4. Liquid Nitrogen Fl sk
5. Avometer
6. Oscilloscope
7. A.C. Bridge circuit

DATA AND CALCULATIONS

BRIDGE COILS

Fixed mutual inductance coils (a statically wound)

Turns - primary (# 32 copper wire) $M_{2p} = 100$

Turns - secondary(# 32 copper wire) $M_{2s} = 100$

Turns - primary (# 32 copper wire) $M_{3p} = 750$

Turns - Secondary (# 32 copper wire) $M_{3s} = 750$

MEASURING COIL

Turns - primary (#36 copper wire) $M_{1p} = 3000$

Turns - secondary(# 36 copper wire) anti clockwire) $M_{1s1} = 1200$

Turns - secondary(# 36 copper wire) (clockwire) $M_{1s2} = 1200$

Magnetic field in primary M_{1p} = 2.8294 Ocrsted

Resistance of primary $M_{1p} = 0.2857 \times 10^3$ ohms

Capacitance of primary M_{1p} & M_{1s1} =

Resistance of secondary M_{1s1} = 95.8 ohms

Resistance of secondary M_{1s2} =100.43 ohms

CALCULATION OF AREA OF COIL M_{1p}

No. of turns of $M_{1p} = 3000$

Length of M_{1p} M_{1p} = 7.50 cm

N_{1p} = No. of turns/cm of $M_{1p} = \frac{3000}{7.5}$

N_{1p} = 400 /cm

No. of turns in 1 mm = 400

No. of turns in 0.1mm = 4

Diameter of wire = 0.1mm

gap between M_{1p} and M_{1s} = 0.3 mm

I. Gap between M_{1p} and M_{1s} on both side = 0.6mm

Thickness of Walt of former = 2mm

II. Diameter of the hole informer = 10 mm

III. Thickness of two walls of former = 4 mm

IV. Thickness of M_{1s} on both sides = 0.64mm

V. Thickness of M_{1p} on both sides = 0.8 mm

diameter of M_{1p} coil = I + II + III + IV + V

= 0.6+10+4+0.64+0.8

= 16.04 mm

= 1.04 cm

r_1 = radius of M_{1p} coil = 1.604/2 = 0.802 cm

A_{1p} area of M_{1p} coil = πr_1^2 = $22/7 \times (0.802)^2 = 2.021\ cm^2$

CALCULATION OF AREA OF COIL $M_{1s} = M_{2s}$

No. of turns of M_{1s} = 1500

length of M_{1s} = 3.75 cm, = 37.5 mm

No. of turns/mm of $M_{1s} = \frac{1200}{37.5}$

N_{1s} = No. of turns/cm of M_{1s} = 320

No. of turns in 0.1mm of M_{1s} = 3.2

I. Thickness of M_{1s} = 0.32 mm

II. Thickness of M_{1s} on both sides = 0.64 mm

III. Diameter of hole in former = 10 mm

IV. Thickness of walls of former on both sides = 4mm

Diameter of M_{1s} coil = I+II+III+IV = 0.32+0.64+10+4 =

$$= 14.96 \text{ mm} = 1.496\text{cm}$$

Radius of M_{1s} coil = 1.496/2

$$r_2 = 0.748 \text{ cm}$$

Area of M_{1s} coil = A_{1s} = $r_2^2 = \frac{22}{7} \times (0.748)^2$

$$A_{1s} = 1.758 \text{ cm}^2.$$

CALCULATION OF i_p

$$V = 1V$$

$$i_{ip} = V/R_{ip} = \frac{1}{0.2857\times10^3}$$

$$i_{ip} = 3.5 \times 10^{-3} \text{ amp}$$

Calculation of H_1

$$H_1 = N_{1p} \times i_{ip} \times A_{1p}$$

$$H_1 = 400 \times \frac{1}{0.2857\times10^3} \times 2.021$$

$$H_1 = 400 \times 3.5 \times 2.021 \times 10^{-3}$$

$$H_1 = 2829.4 \times 10^{-3}$$

$$H_1 = 2.8294 \quad \text{(Oersted)}$$

Calculation of e.m.f.

$$E = -\frac{1}{C}\frac{d}{dt}$$

and $\bar{E} = -u/C \dfrac{dH}{dt}$

$$\frac{E.dl}{A} = -\frac{1}{c}\frac{dB}{dt} \qquad A = A_{1s}$$

$$\oint E.dl = E(\text{volts})$$

$$E = -\frac{1}{c}\frac{dB}{dt} \times A_{1s}$$

$$E = \frac{1}{3\times10^{10}} \times \frac{d}{dt}(B_o S_i n wt) \times A_{1s}$$

$$E = \frac{1}{3\times10^{10}} \; u\, H_o\, w \,/\, \cos wt \quad \times A_{1s}$$

$$E = \frac{1}{3\times10^{10}} \times 100 \times 2.8294 \times 39.4 \times 10^{3} \times 1\times1.758$$

$$= 65.1 \times 10^{-5} \text{ volts}$$

$$\simeq 0.65 \text{ mV}$$

SAMPLE MAGNETICSTEEL

OF S.	SENSITIVITY [mV]	SIGNAL CHANNEL L.H [Hz]	H.F [KHz]	TIME CONT. [mSec.]	REF. CHANL. [KHz]	PHASE [Deg.]	S/W.S	FUNC. P.S.D	METER READING	V_{RMS} [mV]	DIFF. [mV]	ROOM TEMP. [°C]
	2	100	1	30	1	90°	S	X 1	0.42	0.84	0.12	29
	2	100	1	30	1	90	W.S	X 1	0.48	0.96		29
	2	100	1	30	2	90	S	X 1	0.50	1.00	0.28	29
	2	100	1	30	2	90	W.S	X 1	0.64	1.28		29
	2	100	1	30	3	90	S	X 1	0.21	0.42	0.46	29
	2	100	1	30	3	90	W.S	X 1	0.44	0.88		29
	2	100	1	30	3.4	90	S	X 1	0.11	0.22	0.46	29
	2	100	1	30	3.4	90	W.S	X 1	0.34	0.68		29
	2	100	1	30	3.5	90	S	X 1	0.1	0.20	0.40	29
	2	100	1	30	3.5	90	W.S	X 1	0.3	0.60		29
	2	100	1	30	3.6	90	S	X 1	0.08	0.16	0.38	29
	2	100	1	30	3.6	90	W.S	X 1	0.27	0.54		29
	2	100	1	30	3.7	90	S	X 1	0.06	0.12	0.36	29
	2	100	1	30	3.7	90	W.S	X 1	0.24	0.48		29
	2	100	1	30	3.8	90	S	X 1	0.05	0.10	0.36	29
	2	100	1	30	3.8	90	W.S	X 1	0.23	0.46		29

SAMPLE MAGNETICSTEEL

COILS COOLED TO LIQUID N TEMP.

[SE]NSI-VITY [mV]	SIGNAL CHAN-NEL L.H [Hz]	H.F[KHz]	TIME CONT. [mSec.]	REF. CHANL. [KHz]	PHASE [Deg.]	S/W.S	FUNC. P.S.D	METER READ-ING	V_{RMS} [mV]	DIFF. [mV]	ROOM TEMP. [^{0}C]
2	100	1	30	1	90°	S	X 1	0.15	0.30	0.40	29
2	100	1	30	1	90	W.S	X 1	0.35	0.70		29
2	100	1	30	1.7	90	S	X 1	0.74	1.48	0.48	29
2	100	1	30	1.7	90	W.S	X 1	0.98	1.96		29
2	100	1	30	3	90	S	X 1	0.39	0.78	0.06	29
2	100	1	30	3	90	W.S	X 1	0.42	0.84		29
2	100	1	30	3.4	90	S	X 1	0.26	0.52	0.12	29
2	100	1	30	3.4	90	W.S	X 1	0.32	0.64		29
2	100	1	30	3.5	90	S	X 1	0.23	0.46	0.12	29
2	100	1	30	3.5	90	W.S	X 1	0.29	0.58		29
2	100	1	30	3.6	90	S	X 1	0.20	0.40	0.12	29
2	100	1	30	3.6	90	W.S	X 1	0.26	0.52		29
2	100	1	30	3.7	90	S	X 1	0.17	0.34	0.10	29
2	100	1	30	3.7	90	W.S	X 1	0.22	0.44		29
2	100	1	30	3.8	90	S	X 1	0.16	0.32	0.08	29
2	100	1	30	3.8	90	W.S	X 1	0.19	0.38		29

NOISE REDUCED FROM 0.96 TO 0.70 = 0.26 mv (w.s)

" " " 0.84 TO 0.30 = 0.54 mv (s)

CONCLUSION

CONCLUSION

We have constructed an a.c. susceptibility bridge, capable of measuring absolute values, changes in susceptibility of magnetic materials. It has been found that at present the sensitivity of the bridge is about 0.1mV. We have also seen that the value of induced e.m.f. in the coils is ~ 0.48 mV. that is quite close to the theoretical value of 0.65mV expected for a permanent magnetic.

Differences are attributed to loss of e.m.f. the resistive windings of the secondary. We have also seen that a significant reduction in noise takes place. When the coils are cooled to liquid nitrogen temperature. This is very reasonable since at low temperature the electronic and thermal resistance are expected to decrease.

We were unable to measure the susceptibility of a cobalt ferrite liquid sample due to insufficient sensitivity. We believe that improvements suggested in Ref. 7 Can be incorporated in this bridge to increase the sensitivity to 1μV which is the kind of resolution required for measuring weakly magnetic systems.

We found that by variation of R_1 and R_2 the bridge could be zeroed , even after the insertion of the sample, hence the bridge has the required behaviour. We also note that the sensitivity of the bridge is maximum at an input frequency of about 1700 Hz. Hence we define this .. the operating frequency of the bridge.

In conclusion, the a.c. bridge has been successfully built and tested by us.

APPENDICE

APPENDIX A

348

... operational amplifiers of the **741** type in a single ... d.i.l. plastic package. Channel separation ...
... with LM348, MC3403, HA4741, etc.

stock no.	price each 1–24	25+
306–027	£1·24	£0·99

ZN459 – low noise amplifier

... high grade amplifier for applications where ... noise performance is required, e.g., low ... band amplifiers, microphone amplifiers, ... and infra red imaging, etc. The ... is matched to within ± 1 dB, thereby ... device ideal for multichannel ... 8-pin d.i.l. plastic package. ... operating temperature range: 0 °C to +70 °C. ... to ZN459CP.

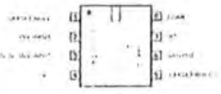
Technical specification
... voltage 6 V d.c. max. 5 V d.c. typ.
... current 2·5 mA typical
... gain 60 dB ± 1 dB max.
... coeff. of gain − 0·2%/°C typ.
... freq. (−3 dB) 15 MHz typ.
... resistance 7 kΩ at 10 kHz
... capacitance 80 pF
... level 2 V typ.
... current 0·8 mA typ.

... resistance 40 Ω typ.
... noise voltage 800 pV/√Hz } $R_S = 0\Omega$.
... noise 3 nV/√Hz
... typ. (1Vp-p, 10 kHz)

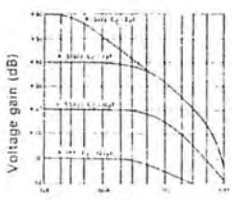
stock no.	price each 1–24	25+
307–468	£2·85	£2·57

For additional data on operational amplifiers refer to page 204.

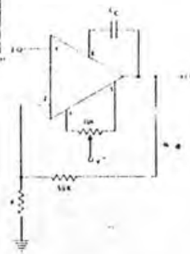
531



TOP VIEW



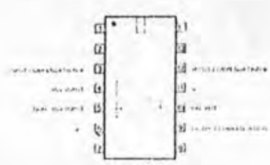
Closed loop non inverting voltage gain as a function of frequency.



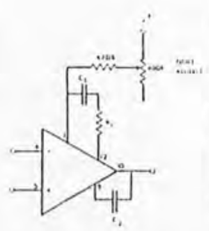
A high performance operational amplifier with a faster slewing rate than general-purpose types, giving superior large signal a.c. performance. Wide full-power bandwidth allows use without special precautions to achieve stability. External frequency compensation. 8-pin d.i.l. plastic package. Ready-made p.c.b. 434-065. Refer to the Hardware section.
Equivalent to NE531V.

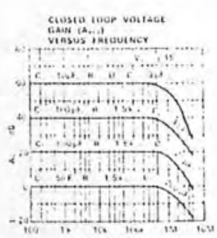
stock no.	price each 1–24	25+
305–872	£1·25	£1·13

709



TOP VIEW





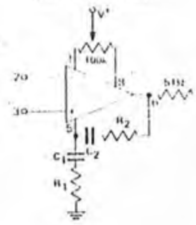
14-pin d.i.l. plastic package. Requires external frequency compensation.
Equivalent to SN72709N, MC1709G, etc.

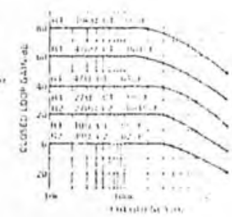
stock no.	price each 1–24	25+
305 254	£0·48	£0·43

725CN (instrumentation) amplifier

TOP VIEW

A high performance operational amplifier housed in an 8-pin d.i.l. plastic package, featuring substantially lower drift characteristics than standard operational amplifiers. Internal pin allocation allows the device to directly replace 14-pin d.i.l. version of the 725 amplifier.
Equivalent to LM725CN.

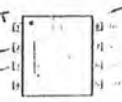




stock no.	price each 1–24	25+
308–168	£3·18	£2·86

741

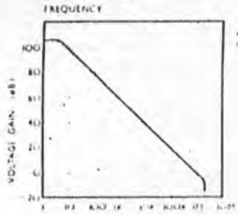
OFFSET NULL
INV. INPUT
NON INV IN.
V−



N.C
V+
OUTPUT
OFFSET

TOP VIEW

Features internal frequency compensation and short circuit protection. 8-pin d.i.l. plastic package. Ready-made p.c.b. 434–065. Refer to the Hardware section.
Equivalent to μA 741CP, etc.



stock no.	price per pk 1–9	10+
305–311	£1·76	£1·58

5 per pk

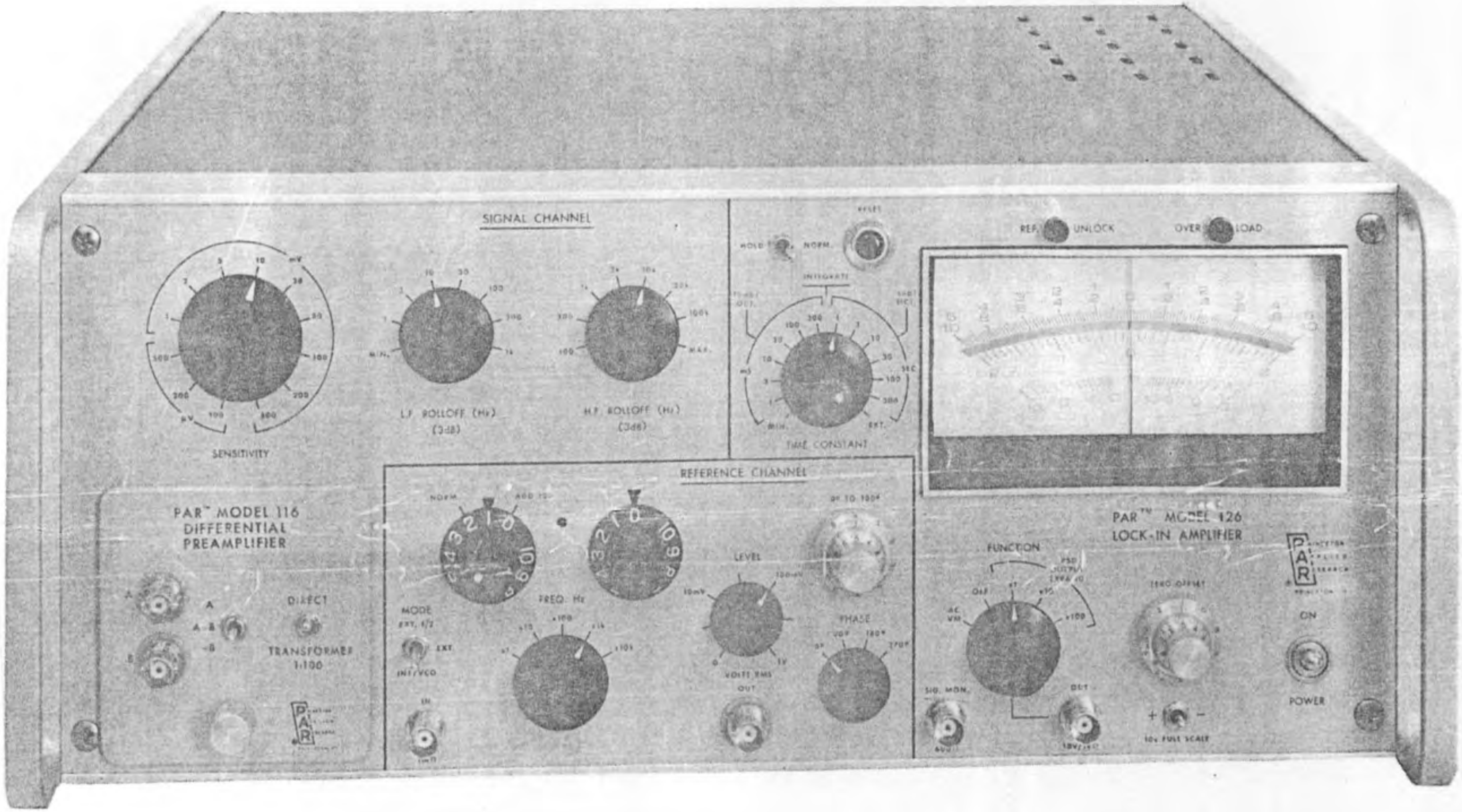
741N

A guaranteed low noise pin compatible replacement for an 8 pin d.i.l. 741. Ready made p.c.b. 434-065. Refer to the Hardware section.

stock no.	price each 1–24	25+
306–308	£1·00	£0·85

APPENDIX B

Lock in Amplifier Model 126



STEP ONE

Choose an appropriate preamplifier (see comparison chart in 2.2 Specifications) and plug it into the Model 126.

STEP TWO

Be sure the rear panel 115/230 V switch is in the proper position. Connect the Model 126 to the power line. Turn the Power switch on.

STEP THREE

Connect the signal source from the experiment to the preamplifier input(s).

STEP FOUR

Select a mode of operation for the Reference Channel. Base your choice on what is to control the frequency.

INT/VCO

Set the frequency by means of the front panel digital dials and range switch for Internal operation. For VCO operation, set up as for Internal except add or subtract to the set frequency by feeding a plus or minus voltage to the VCO Input on the rear panel.

EXTERNAL

Connect the external reference signal source to the Reference Channel Input connector. Set the range switch to a position which will cover the frequency of the external reference according to the following table and wait for the Unlock lamp to extinguish.

BAND	(FREQUENCY RANGE)	MAXIMUM TIME
X1	(0.2 Hz to 21 Hz)	15 min.
X10	(2 Hz to 210 Hz)	2 min.
X100	(20 Hz to 2.1 kHz)	10 sec.
X1K	(200 Hz to 21 kHz)	2 sec.
X10K	(2.1 kHz to 210 kHz)	2 sec.

EXTERNAL f/2

Use this mode to measure the second harmonic of a signal. Connect the external reference signal source to the Reference Channel input connector. The oscillator in the Model 126, and thus the Phase Sensitive Detector, will operate at twice the frequency of the external reference within the limits in the above table after the Unlock lamp has extinguished.

STEP FIVE

Apply the reference drive to the experiment. The reference drive is obtained from the Reference Channel output connector on the front panel when using the INT/VCO mode. In the external modes, the experiment can be driven by either the external reference source or by the Model 126 output, whichever is most suitable.

STEP SIX

Set the two Rolloff controls to bracket the frequency of the signal to be measured.

STEP SEVEN

Preliminary adjustment of the Sensitivity switch: Set the Function switch to ACUM. Set the Time Constant controls to 300 ms, 6 dB/octave, and the small toggle switch to Normal. Be sure the Zero Offset switch is off (center position). Place the Sensitivity switch to the most counterclockwise position which does not cause overload.

STEP EIGHT

Turn the Function switch to the X1 position. Adjust the phase controls for a maximum positive meter deflection. Change the Time Constant switch to as large a setting as required to reduce the meter fluctuations to a tolerable level. If the meter deflection is too small after adjusting the phase controls, attempt to increase the Sensitivity setting without incurring overload. If the meter indication is still too low, use the Function switch to increase the output gain to either X10 or X100 as necessary.

STEP NINE

Assuming a Type 116, 117, or 119 Preamplifier is installed, operating in the Direct mode, the amplitude of the signal can now be read on the meter with full-scale deflection equal to the Sensitivity switch setting divided by the output gain, e.g., 100 mV and X10 settings yield a full-scale reading with a 10 mV input signal. When using the Type 118 Preamplifier, the Lock-In Amplifier is ten times more sensitive so that full-scale corresponds to 1 mV. The phase of the signal may be read directly from the phase controls.

To increase the resolution of the meter to small amplitude variations, the Zero Offset controls can be employed to suppress the deflection allowing the output gain to be increased without overload. The amplitude of the signal can then be read directly from the Zero Offset controls providing they have been set to obtain a null on the meter.

Hold/Normal switch ------------------- NORM.
Time Constant switch ---------------- Red knob, 6 dB; Black knob, 300 ms.
Function switch --------------------- ACVM
Zero Offset ------------------------- 0 (fully ccw).
10X Full Scale switch --------------- OFF (center position).
(Zero Offset toggle switch)

(6) Turn the Power switch ON and wait five minutes for warmup.

(7) Connect a cable between the Reference OUTPUT connector and the preamplifier's A input.

(8) Adjust the Reference Level potentiometer for a full-scale reading to the right. The resulting position of the Level control should be between the 10 mV and 100 mV marks on the front panel.

(9) Change the L.F. Rolloff switch setting to 1 kHz. The meter should read between 30% and 50% of full-scale. Return the switch to MIN.

(10) Change the H.F. Rolloff switch to 100 Hz. The meter should read between 15% and 40% of full-scale. Return the switch to MAX.

(11) Change the Time Constant switch setting to 10 sec. The meter should read full-scale.

(12) Place the Hold/Normal switch to HOLD. The meter should still indicate full-scale.

(13) Change the red Time Constant switch to INTEGRATE.

(14) Press the Reset button momentarily. The meter should read zero.

(15) Place the toggle switch to NORM. for exactly five seconds, and then return to HOLD. The meter should read approximately half-scale.

(16) Set the Time Constant switches to 300 ms and 12 dB/OCTAVE. Set the toggle switch to NORM. The meter should read full-scale.

(17) Change the Function switch to X1. The meter should again read full-scale.

(18) Set the Function switch to X10 and readjust the Reference Level control to obtain full-scale.

(19) Set the Zero Offset toggle switch to +, and the turns counting dial fully clockwise. The meter should be against the left-hand limit and the Overload lamp should be on.

(20) Set the Function switch to X100. The meter should be somewhere on scale, either positive or negative, and the Overload lamp should be off.

(21) Return the Function switch to X10 and the Zero Offset switch to OFF

SECTION III
INITIAL CHECKS

1 INTRODUCTION

ne following procedure is provided to facilitate initial performance check-
ng of the Model 126. In general, the procedure should be performed after
nspecting the instrument for shipping damage (any noted to be reported to
ne carrier and to Princeton Applied Research Corporation), but before using
t experimentally. Should any difficulty be encountered in carrying out
nese checks, contact the factory or one of its representatives, a list of
nom appears at the end of the manual.

.2 EQUIPMENT NEEDED

(1) General-purpose oscilloscope.

(2) Signal generator, having any 1 kHz repetitive waveshape that crosses its mean exactly twice each cycle and having a peak-to-peak voltage of at least 100 mV.

(3) Assorted BNC cables.

.3 PROCEDURE

)TE: This procedure must be performed in sequence.

(1) Install a preamplifier if not already installed.

(2) Check the rear panel 115/230 switch. Be sure the number showing in the window corresponds to the line voltage to be used.

(3) Turn the front panel Power switch OFF.

(4) Plug the line cord into the rear panel and wall receptacles.

(5) Set the front panel controls as follows.

Meter	Check mechanical zero. Adjust if necessary.
Preamplifier	Select A input and DIRECT (if applicable).
Sensitivity switch	50 mV (if using a Model 118 Preamplifier, set the Sensitivity switch to 500 mV).
L.F. Rolloff	MIN.
H.F. Rolloff	MAX.
Frequency Mode switch	INT/VCO
Reference Frequency dials	Red knob, NORMAL; Digits, 4.0
Reference Frequency Range	X100
Phase potentiometer	0° (fully ccw).
Phase switch	0°

(center position). The meter should read full-scale.

(22) Place the Phase switch to 90°. The meter should go to zero ±20%.

(23) Set the Phase switch to 180°. The meter should read negative full-scale (to the left).

(24) Set the Phase switch to 270°. The meter should read zero ±20%.

(25) Turn the Phase potentiometer nine full turns clockwise (90°). The meter should read full-scale to the right.

(26) Connect the signal generator's output to the Input connector in the Reference Channel section of the front panel. The peak-to-peak voltage can be anywhere between 100 mV and several volts. Set the frequency to approximately 1 kHz.

(27) Disconnect the cable between the Reference Channel output connector and the preamplifier.

(28) Advance the Level control to the 1 V setting.

(29) Monitor the Reference Channel Output connector with the oscilloscope. The waveform should be a sine wave, 2.8 V (±0.3 V) peak-to-peak. (If the Power Amplifier Option has been installed, the amplitude will be a factor of ten greater.)

(30) Place the Reference Mode switch to EXT. The Unlock lamp should glow momentarily. During this time the frequency of the oscilloscope waveform should have increased to where it is equal to the frequency of the signal generator.

(31) Note the frequency on the oscilloscope. Then place the Mode switch to EXT. f/2; the frequency should double.

This completes the Initial Checks. If the instrument performed as indicated, one can be reasonably sure that it is operating properly.

CHARACTERISTICS

2.1 INTRODUCTION

The PAR™ * Model 126 Lock-In Amplifier is the basis of a high-quality system for the precision measurement of signals contaminated by noise, power line interference, frequency drift, etc. The measurement is accomplished by means of an extremely narrow bandwidth phase sensitive detector which has its center frequency locked (referenced) to the frequency of the signal to be measured. This results in considerable improvement of the signal to noise ratio, permitting accurate amplitude and phase measurements.

A few of the outstanding characteristics include a series of plug-in preamplifiers providing optimum low noise figure for practically any range of signal frequency and source resistance. To compliment the standard line of plug-in preamplifiers, the Model 184 Photometric Preamplifier has a virtual ground input which connects directly and efficiently to photodetector outputs.

The succeeding signal amplifier in the Model 126 has independently adjustable high and low frequency rolloffs providing a means of rejecting a great deal of noise before phase sensitive detection. The band of frequencies which do enter the phase sensitive detector is converted to sum and difference frequencies with respect to the reference. A selectable single or double section low pass filter rejects all frequency components except the dc component due to a signal present at the reference frequency. Thus noise or other interference is eliminated, leaving only a dc voltage proportional to the signal. Calibrated phase controls indicate the phase of the incoming signal while the amplitude is displayed on a standard panel meter.

Certain design features have been incorporated to facilitate practical applications. The gain of the ac signal amplifiers before the detector and the gain of the dc amplifiers after the detector can be separately determined to obtain the greatest degree of output stability for a given incoming signal-to-noise ratio. The reference circuitry, in addition to operating as a tracking oscillator locked to either the external frequency or twice the external frequency, can also function as a VCO or as a standard laboratory oscillator with its own digital frequency dials. It is possible to operate over certain frequency ranges of 100 to 1 without touching a single control, including phase. The output filter can be switched to a true integrator mode with hold and reset capabilities.

A wide selection of accessories enhances the versatility of the Model 126. A two-phase accessory, the Model 127, enables simultaneous measurement of both in-phase and quadrature components while the Model 126 independently detects at any selected third phase or operates as a standard ac voltmeter. Other ancillary equipment includes light choppers, photomultiplier tube housings, teletype, and computer interface modules.

*PAR is a trademark of Princeton Applied Research Corporation.

SPECIFICATIONS

ERAL

FREQUENCY RANGE: 0.2 Hz to 210 kHz.

SENSITIVITY: 18 full-scale ranges in a 1-2-5 sequence combining both the Sensitivity control and Output Gain control ranges. Actual full-scale voltages are determined by the choice of preamplifier. Refer to the preamplifier chart for detailed information.

AUXILIARY POWER OUTPUT: Regulated ±24 V at 150 mA capability is available at a rear panel connector.

NAL CHANNEL

PREAMPLIFIERS:

IFICATION		Model 116	Model 117	Model 118	Model 119
nput Z		Selected by front panel switch: Direct: 100 meg. SE/DE Transformer[a]: Low Z SE/DE	100 megohms SE/DE	>10 kilohms SE/DE	Selected by front panel switch: Direct: 100 meg. SE/DE Transformer: Low Z SE/DE
ndwidth		Direct: 0.2 Hz – 210 kHz Transformer[b]: 1.5 Hz – 10 kHz	0.2 Hz – 210 kHz	0.2 Hz – 210 kHz	Direct: 0.2 Hz – 210 kHz Transformer[b]: 1 kHz – 210 kHz
Figure ontour)	R_S	Direct: 4 kilohms – 20 megohms Transformer: 15 ohms – 100 ohms	4 kilohms – 20 megohms	200 ohms – 2.5 kilohms	Direct: 4 kilohms – 20 megohms Transformer: 1–10 ohms
	f	Direct: 1 Hz – 210 kHz Transformer: 20 Hz – 10 kHz	1 Hz – 210 kHz	500 Hz – 210 kHz	Direct: 1 Hz – 210 kHz Transformer: 1.5 kHz – 150 kHz
mon Mode ction Ratio		Direct: 120 dB at 60 Hz Transformer: 140 dB at 60 Hz	120 dB at 60 Hz	110 dB at 60 Hz	Direct: 120 dB at 60 Hz Transformer: 120 dB at 60 Hz
ull Scale nsitivity		Direct: 1 μV Transformer: 10 nV	1 μV	100 nV	Direct: 1 μV Transformer: 10 nV
aximum Input Voltage		Direct: ±200 Vdc Transformer: ±200 Vdc[c]	±200 Vdc	±5 V	Direct: ±200 Vdc Transformer: ±200 Vdc[c]

NOTES: a. Input Z is complex but approaches 0.25 H at low frequencies. May be wired for 1:50 to 1:350 turns ratio. Standard is 1:100.

b. Varies with source impedance.

c. Common mode only.

Table II-1. PREAMPLIFIER COMPARISON CHART

BANDWIDTH: Dynamic range of the Lock-In Amplifier is increased by limiting the bandwidth of the signal channel. Independent low and high frequency rolloff controls (6 dB/octave) set the −3 dB bandwidth. The range of the Low Frequency Rolloff is 1 Hz to 1 kHz in a 1-3-10 sequence plus MIN. The High Frequency Rolloff covers a range of 100 Hz to 100 kHz in a 1-3-10 sequence plus MAX. A relatively narrow passband ($Q=\frac{1}{2}$) can be obtained by setting both controls to the same frequency. When operated with maximum bandwidth, response is flat within ±1% from 10 Hz to 110 kHz, ±2% from 110 kHz to 210 kHz, and +0, −30% below 10 Hz.

SIGNAL MONITOR CONNECTOR: Allows monitoring of signal just ahead of demodulation. 100 mV rms sine wave corresponds to full-scale in X1 setting of the Function (Output Gain) switch, 10 mV in X10, and 1 mV in X100. Output impedance is 600 ohms.

Line frequency pickup due to internal power supply is less than 20 nV rms (referred to the Direct inputs of a Type 116 Preamplifier).

REFERENCE CHANNEL

OUTPUT: Sine wave at the reference oscillator frequency. Amplitude is continuously adjustable from 0 to 1 V rms by means of the Level control. Total harmonic distortion is less than 2%. An optional power amplifier is available which increases the maximum output level to 10 V rms (adjustable from 0 V to 10 V) and fixes the output impedance at 600 ohms.

OPERATING MODES:

INT/VCO: Frequency of the internal reference oscillator is set by means of the front panel digital dials and/or the rear panel VCO Input control voltage. Frequency dial accuracy is ±2% or 0.05 Hz, whichever is greater. VCO control voltage of 0 to ±10 V corresponds to the full frequency range of the set band. VCO input impedance is 10 kilohms.

EXTERNAL: The internal reference oscillator will lock in both frequency and phase to virtually any externally generated signal crossing its mean only twice each cycle. Minimum time required on either side of the mean is 100 ns. Amplitude excursion must be at least 50 mV on each side of the mean. Input impedance is 1 megohm.

Once locked-on, the reference oscillator will track the external signal over a frequency range of 100:1 within the range of the set band of frequencies. Maximum frequency acquisition (lock-on) times for each frequency band are given in the following table:

BAND	(FREQUENCY RANGE)	MAXIMUM TIME*
X1	(0.2 Hz to 21 Hz)	15 min.
X10	(2 Hz to 210 Hz)	2 min.
X100	(20 Hz to 2.1 kHz)	10 sec.
X1K	(200 Hz to 21 kHz)	2 sec.
X10K	(2.1 kHz to 210 kHz)	2 sec.

Table II-2. MAXIMUM LOCK-ON TIME FOR EACH BAND

*Time can be shortened appreciably by momentarily switching to INT/VCO mode and manually setting the oscillator to the proper frequency.

When the frequency has locked, the phase will track at the typical rates shown in Figure II-1.

EXTERNAL f/2: Essentially the same as the External mode, except the internal reference oscillator locks to twice the frequency of the external control signal. Maximum input frequency is 105 kHz.

PHASE SHIFTER: Phase accuracy is ±2° from 0.2 Hz to 21 kHz, and ±5° from 21 kHz to 210 kHz.

EMODULATOR

The Function switch allows the Demodulator to operate as a Phase Sensitive Detector with selectable output gain or as a conventional ACVM.

DC OUTPUT GAIN/ZERO STABILITY: The Function switch permits selection of the optimum gain versus stability trade-off for the characteristics of the signal being measured. The following table lists both the zero stability and overload capabilities at each switch setting.

OUTPUT GAIN	OUTPUT STABILITY	OVERLOAD CAPABILITY*
X1	15 ppm/°C	3 x full-scale
X10	100 ppm/°C	30 x full-scale
X100	1000 ppm/°C	300 x full-scale

Table II-3. OUTPUT STABILITY AND OVERLOAD AS A FUNCTION OF OUTPUT GAIN

*Defined as the ratio, at the input of the lock-in amplifier, of the maximum pk-pk non-coherent signal which can be applied without overloading the Model 126 to the pk-pk coherent signal required to yield full-scale Model 126 output meter deflection. Note that, expressed as the ratio of the pk-pk non-coherent signal to the rms value of the coherent signal required for full-scale output, this number can be as great as 1000 (maximum input signal: 3 V pk-pk).

UTPUT FILTER OPERATING MODES

6 dB: Single section low pass filter with Time Constant switch selections of 1 ms to 300 s in a 1-3-10 sequence plus minimum (MIN.) and external (EXT.). Rolloff rate is 6 dB/octave.

12 dB: Double section low pass filter with the same range of settings as for 6 dB/octave. Rolloff rate is 12 dB/octave.

INTEGRATE: A true integrate mode may be selected at the front panel. Build-up rate is such that a full scale signal will cause the output to reach the usual full scale level after a time interval equal to the time constant selected by the Time Constant switch has elapsed. The Integrate mode is often used to measure signals which are not continuously present at the input. Accuracy is ±20%.

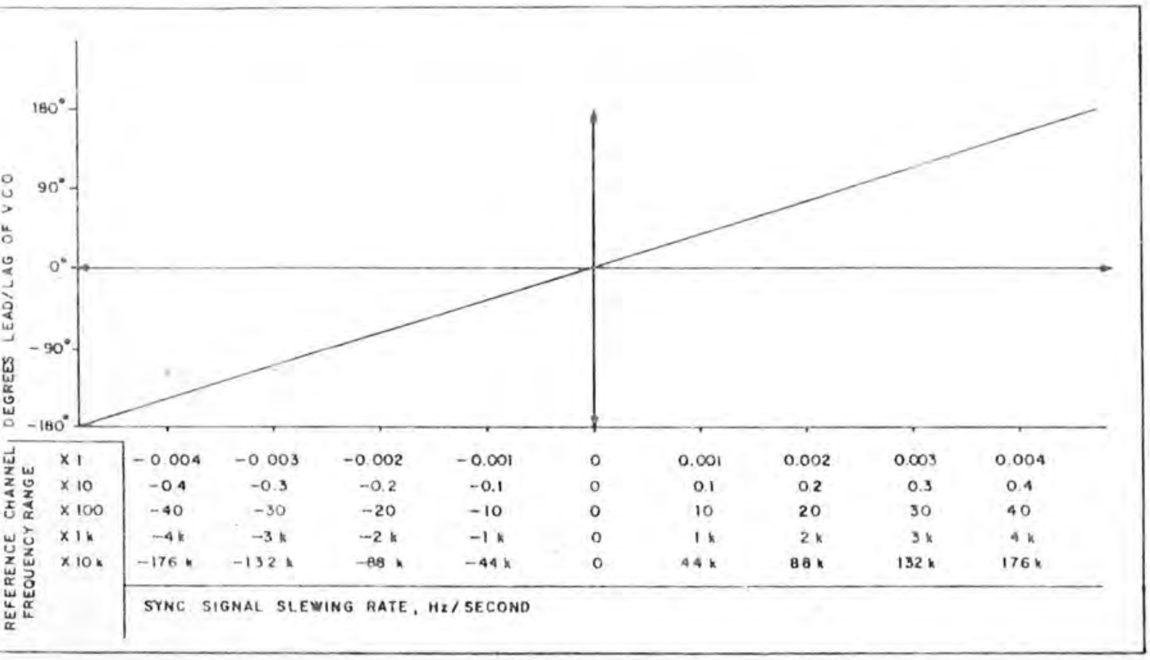


Figure II-1. SYNC. SIGNAL SLEWING RATE
AS A FUNCTION OF RANGE

HOLD/NORMAL AND RESET: A set of two switches can be used for holding or resetting the output level in any of the three Output Filter operating modes. The Hold/Normal switch can be used to determine the integrating time limits. Output stability when holding is typically 200 ppm of full-scale per time constant interval.

EQUIVALENT NOISE BANDWIDTH: 416 μHz minimum (300 seconds maximum internal time constant, 12 dB/octave).

ZERO SUPPRESS: Calibrated control permits off-setting zero by ±1000% on the X10 and X100 ranges, and ±300% on the X1 range of the Function switch.

METERED OUTPUTS: Standard panel meter is 0.5% linear. Full-scale corresponds to ±10 V at the front panel Function Out connector (1 kilohm impedance) and at the recorder output binding posts (1 kilohm impedance).

ACVM ACCURACY: ±1% from 10 Hz to 20 kHz, increasing to ±10% at 210 kHz; +0, -30% below 10 Hz. Average responding.

TEMPERATURE RANGE: 15°C to 45°C.

POWER REQUIREMENTS: 105-130 V or 210-260 V; 50-60 Hz; 30 watts.

SIZE: 6-31/32" H x 17-1/8" W x 18-1/4" D (17.8 cm H x 43.6 cm W x 46.5 cm D).

MOUNTING: Brackets included for mounting in standard 19" rack.

SECTION IV
OPERATING INSTRUCTIONS

4.1 INTRODUCTION

The PAR™ Model 126 Lock-In Amplifier is a precision instrument crafted to permit detailed analysis of signals buried in noise or other interference. This section describes the block diagram operation, full consideration of the function of each control, and the interfacing of the Model 126 with both the experiment and ancillary equipment. For a cursory method of setting the controls, at the expense of overlooking the reasoning behind certain steps, see the Condensed Operating Instructions, Section I. For more specific information on the operation within one of the circuit blocks, refer to the Theory and Circuit Description section. Beginning with an overall assessment of the instrument and following with full disclosures of each operational trait, this section is confined to mastering the controls of the instrument and connecting it to an experiment in the best manner.

4.2 BLOCK DIAGRAM AND FRONT PANEL LAYOUT

When operating as a Lock-In Amplifier, the Model 126 comprises three distinct sections, the Signal Channel, Reference Channel, and Phase Sensitive Detector along with its output circuitry; each is clearly separated at the front panel by heavy black lines. A typical situation is shown in Figure IV-1. To improve clarity the interference which might be present at the output of the experiment is not shown. The signal being measured, combined with any interference, is amplified by the Signal Channel which has a variable bandwidth capable of rejecting a large amount of the interference before submitting the signal to the Detector. The Reference Channel has the twofold function of driving the experiment so that the frequency of its signal output is known, and switching the Detector at exactly the same frequency but with an adjustable phase. Therefore, the Detector receives both a signal with its interference and a frequency reference of known phase. From these two inputs, the Detector produces sum and difference frequencies. The difference between the signal and the reference is, of course, zero frequency or simply dc, with an amplitude proportional component of the signal at the set phase. All interference and noise not at the reference frequency appear at the Detector's output at frequencies greater than zero. By sending the total Detector output through a lowpass filter, only the dc is present to deflect the meter a calibrated amount.

4.3 SIGNAL CHANNEL

PREAMPLIFIERS

The first decision to be made when placing the Model 126 in operation is the choice of plug-in preamplifier. The choice is based on consideration of the noise figure, common mode rejection ratio, and distortion, for the particular signal source characteristics. Generally, for very low signal levels, the selected preamplifier model yields the lowest noise figure for the source resistance and frequency of the signal being measured. Four models are presently available for voltage amplification age conversion (often used with photomultiplier tubes).

Figure IV-2 shows the area in which the noise contributed by the preamplifier is insignificant compared to that which must be present in the source resistance due to thermal agitation. The Model 117 is well suited for signals from relatively high source resistances. The Models 116 and 119 are identical in operation from high source resistances but have an input impedance matching transformer which can be switched in to allow low noise performance

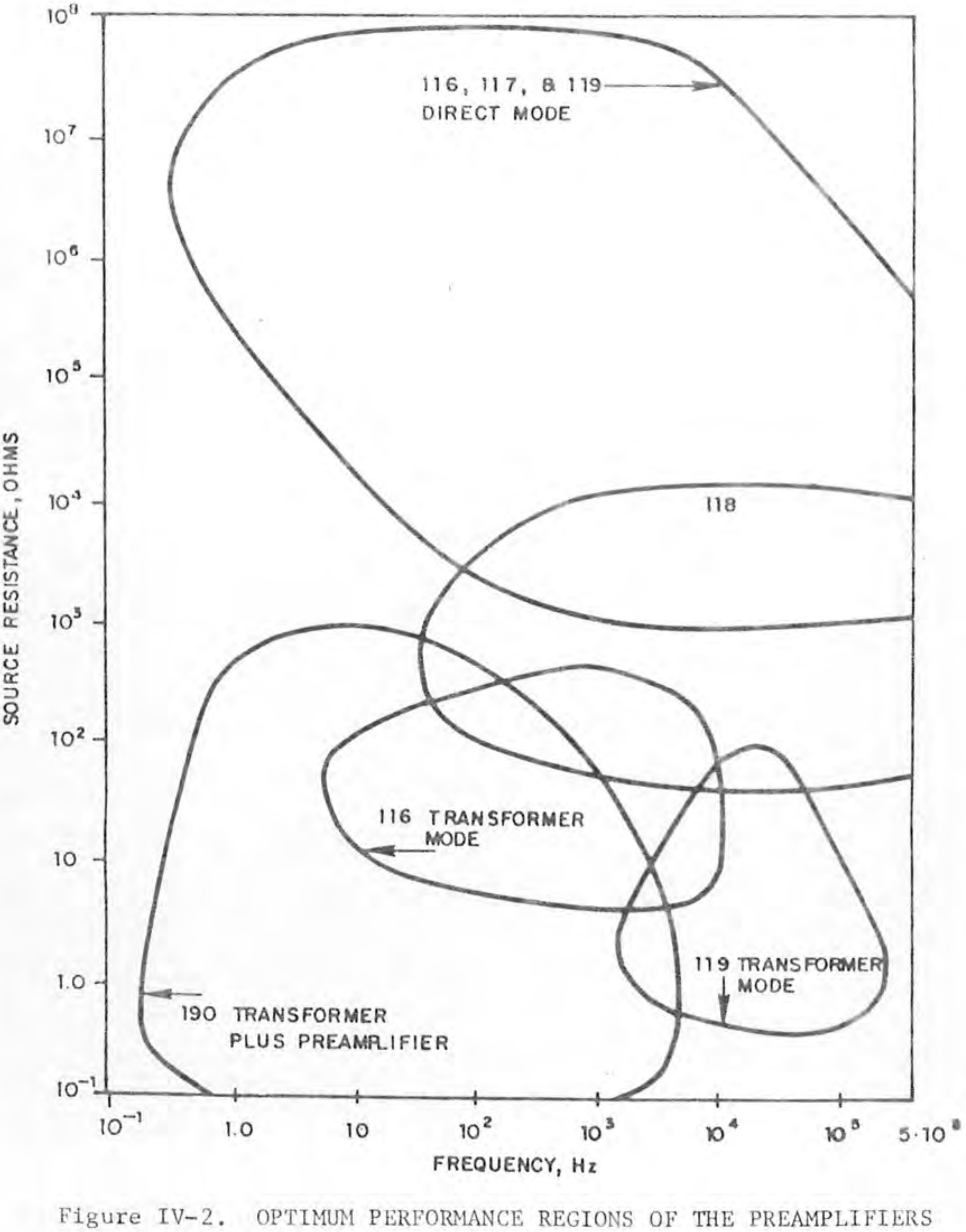


Figure IV-2. OPTIMUM PERFORMANCE REGIONS OF THE PREAMPLIFIERS rom very low source resistance values. The difference between the two is ne optimum frequency range for the transformer type used in each model. The odel 118 performs well with intermediate values of source resistance.

lthough not actually a preamplifier, the Model 190 Low Noise Transformer hould be considered at this time. This transformer, when placed ahead of a igh input impedance preamplifier such as the Model 116, 117, or 119, forms a ystem which has an optimum noise figure contour as shown in Figure IV-2. oth 1:100 and 1:1000 turns-ratios are provided.

detailed discussion of noise figures is given in later paragraphs under Use of Noise Figure Contours".

he Model 184 Photometric Preamplifier provides an input calibrated in terms f current ranging from 10^{-5} amperes per volt to 10^{-9} amperes per volt, peritting easy, accurate measurements of photodetector current, or any other inute current. Overall system sensitivity is the ampere/volt setting of the odel 184 multiplied by the Sensitivity and Output Gain settings of the Model 26, permitting measurement of currents as small as 10^{-15} ampere full scale. or typical noise figure information for the Model 184, refer to the instrucion manual for that preamplifier.

nother factor in selecting a preamplifier is its ability to amplify without istortion. If an experiment requires measurement of low level harmonics in he presence of a high level fundamental, it is important that the preamplifier does not add significant harmonic signals by non-linear amplification of he fundamental. With the single exception of the Model 118, P.A.R. preamlifiers operating in the Direct mode distort a signal so slightly as to be nmeasurable (much less than 0.01%) using conventional methods. The Model 18 can, at high signal input levels, cause appreciable distortion as shown n Figure IV-3. When using the Models 116 or 119 in the transformer mode, oth phase shift and distortion can occur and should be measured for each inividual operating condition.

ommon mode rejection ratio is an important consideration when the signal beng measured has a high level of interference common to both terminals across hich the signal appears. This type of interference is rejected by the preamplifier itself so that the signal to noise ratio "improvement" is in addiion to the improvement made by the lock-in process. Figure IV-4 depicts the ypical common mode rejection curves of the preamplifiers in both the direct nd transformer modes operating differentially (A - B).

he full scale sensitivity of the Model 126 Lock-In Amplifier can be affected y the choice of preamplifier. Refer to the paragraphs on Full Scale Sensiivity on page IV-14.

SE OF NOISE FIGURE CONTOURS

f every precaution is taken to insure that the signal being measured is not ontaminated by such things as line frequency interference and ground loops, here is still a residual noise, the Johnson noise, which is present due to he thermal effects on the electrons in the source resistance at any temperaure above absolute zero. This thermal noise can be calculated from the folowing equation:

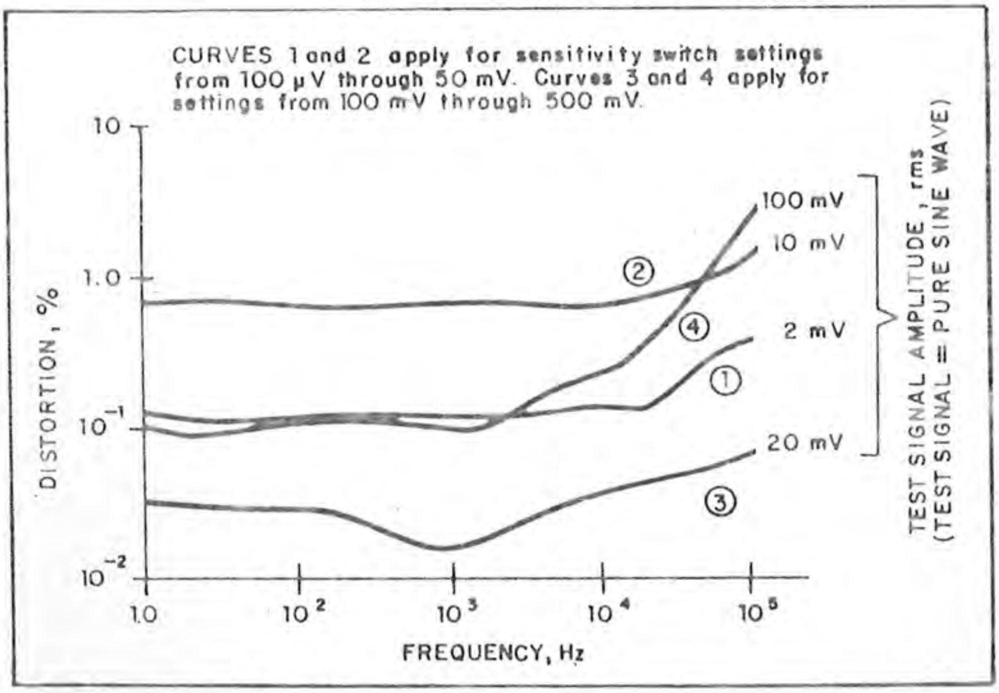


Figure IV-3. DISTORTION v. FREQUENCY AND AMPLITUDE, MODEL 118 ONLY

$$E_n = \sqrt{4KTR_S B_N} \qquad (1)$$

where:

E_n = rms noise voltage within a bandwidth of measurement.
K = Boltzmann's constant = 1.38×10^{-23} joules/Kelvin.
T = Absolute temperature in Kelvins.
R_S = Resistive component of the impedance across which the voltage is measured.
B_N = Bandwidth across which the noise voltage is measured.

Since room temperature is the most common situation for signal sources, the graph shown in Figure IV-5 which relates the Johnson noise per root-Hertz for a range of source resistances at room temperature has been provided.

A second source of noise must also be considered. Every amplifier of any type or design also has a certain level of internally generated noise which appears at its output. It can be seen that whenever any signal is passed through an amplifier there will be a certain amount of noise added to that already inherent in the signal source. One way of evaluating the practical importance of these two types of noise is to relate them by determining the "Noise Figure" for a preamplifier according to the following formula:

Noise Figure =

$$20 \log_{10} \frac{\text{total rms noise voltage referred to the amplifier input}}{\text{source thermal rms noise voltage}} \text{ dB} \qquad (2)$$

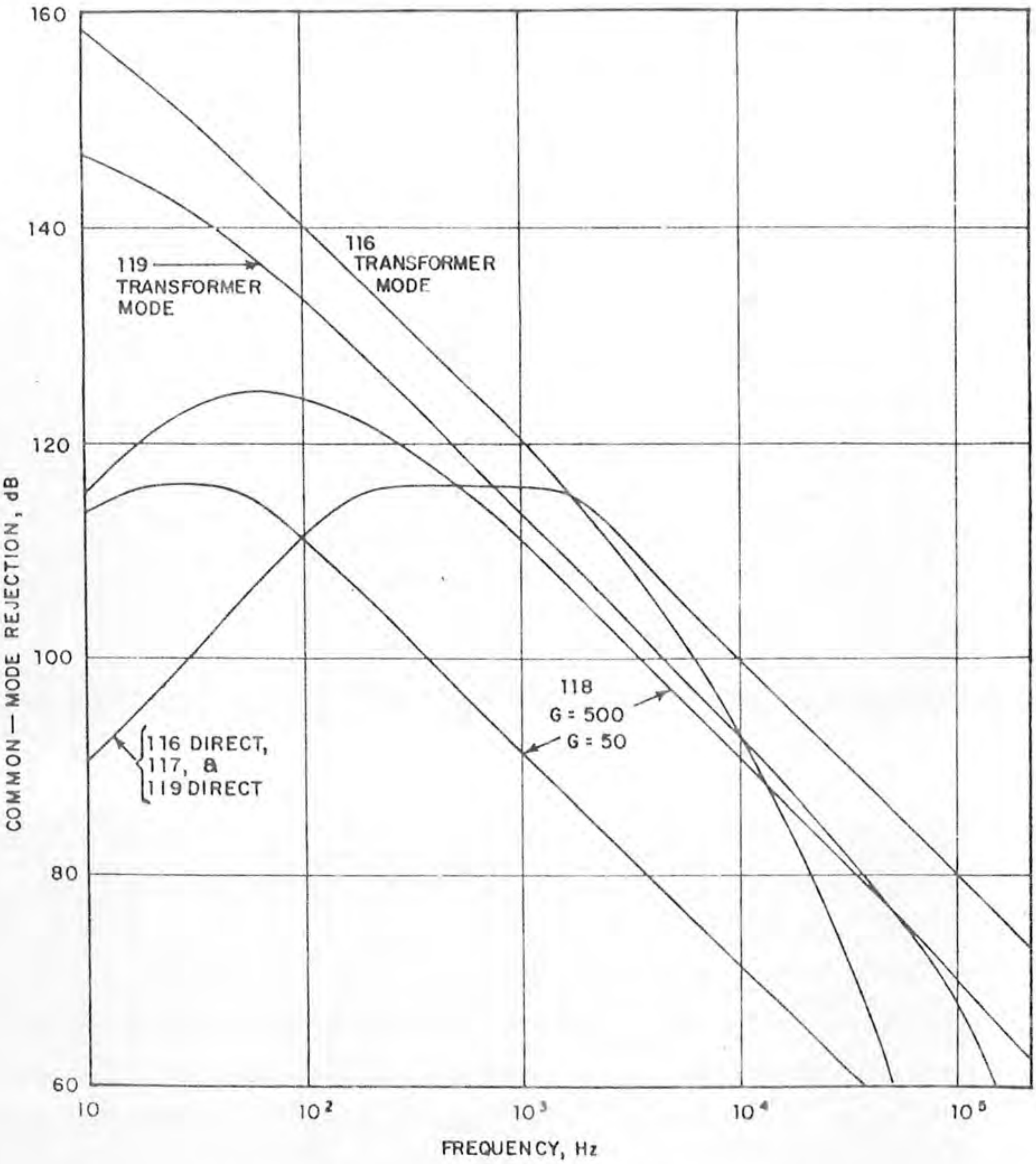


Figure IV-4. TYPICAL COMMON-MODE REJECTION

Since the noise contributed by a given preamplifier depends on the particular resistance connected to its input and also on the frequency of interest, the noise figure for that preamplifier can be plotted as a set of contours against frequency and resistance axes. Contours for the various plug-in preamplifiers are given in Figures IV-6 through IV-9. Additional information pertaining to the Model 190 is included in Figures IV-10 through IV-12. The inner areas with the lower noise figure values are located where the preamplifiers contribute a negligible amount of noise compared to the noise due to the source resistance. If the characteristics of the signal source are

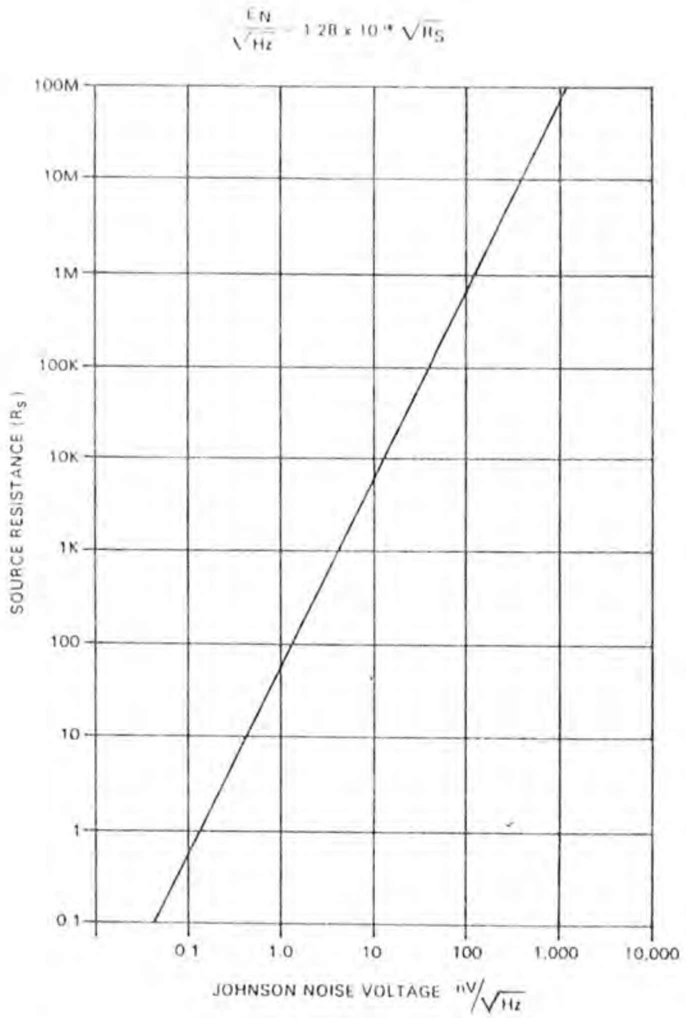


Figure IV-5. JOHNSON NOISE VOLTAGE FROM A RESISTANCE AT ROOM TEMPERATURE (290 K)

own, it is a simple matter to determine which preamplifier is most appro-iate and whether it should be operated with direct or transformer coupled puts.

is important to note that if the source resistance of the signal is lower an that which would be optimum for a given preamplifier, the source resis-nce should not be changed by merely adding a series resistor to it. To do ould actually result in a degraded signal-to-noise ratio although the ise figure has improved (only because more thermal noise has been intro-ced at the input without a change in signal level). A similar result oc-rs if a shunt resistor is added to a source which is higher than the opti-n value for a preamp; in this case, the effect is to reduce the signal lev-more rapidly than the thermal noise. A transformer can often be used to vantage since, over certain ranges, the apparent source resistance as seen the preamplifier can be changed to optimum without degrading the signal to se ratio.

pose that the Model 126 is to be connected to an experiment that presents a source resistance of 100 ohms and a signal frequency of 1 kHz. In Figure IV-6, it can be seen that the noise figure would be approximately 10 dB when using either a Model 117 or a 116 or 119 in the Direct mode. If the Model 116 were operated with the transformer switched in, a noise figure of less than 1 dB (Figure IV-7A) would be realized. In addition, the Model 118 (Figure IV-9) would be slightly higher than 1 dB under these signal conditions. Strictly from a noise figure standpoint, the Model 116 operated in the Transformer mode would be the best choice. Nevertheless, the frequency response curves for this transformer, Figure IV-7B, should be checked. It can be seen that the calibration will be affected, and this should be taken into account.

It should be emphasized that noise figure is not the only criterion in choosing a preamplifier, especially if the signal level is not small compared to the noise. A flat frequency response curve such as the Model 118 may be important if the operating frequency of the Model 126 must be scanned over a large range.

So far, actual noise voltages have not been discussed. This is because the voltage as given in Equation (1) depends on the bandwidth, which, under normal use of the Model 126 as a Lock-In Amplifier, is controlled by the setting of the Time Constant switch. The Noise voltage and the bandwidth is described in some detail in the "Output Filter Modes" paragraph in subsection 4.5.

FULL SCALE SENSITIVITY

Effect of Preamplifier Choice: The Sensitivity switch on the Model 126 has settings ranging from 100 μV to 500 mV full scale. The actual full-scale sensitivity is equal to the setting when using the Model 117 or either the Model 116 or 119 operated in the Direct mode. Use of the Transformer mode on the latter two will increase sensitivity by a factor of 100 so that the Sensitivity switch covers a range of 1 μV to 5 mV. The Model 118 boosts sensitivity by a factor of 10 so that a range of 10 μV to 50 mV results.

All of the foregoing assumes that the Output Gain (see PSD and Output Amplifiers) is set to X1. Increasing the Output Gain setting also affects the actual full scale sensitivity. Using the maximum setting of X100 in conjunction with the Model 116 or 119 Preamplifier in the Transformer mode yields a full scale sensitivity of 10 nV.

Optimum Setting: In general, regardless of what preamplifier or Output Gain is used, set the Sensitivity to the lowest full-scale voltage which does not cause overload as indicated by the lamp.

ROLLOFFS

The prime function of the low and high frequency rolloff circuits is to eliminate as much interference and noise as possible while having minimal effect on the signal of interest. Under most conditions, this increases the dynamic range of the instrument and reduces the fluctuations in the final dc output of the Model 126. As shown in the Block Diagram, Figure IV-1, the bandwidth of the signal channel can be set by individual High and Low Frequency Rolloff switches.

Use of the MIN. and MAX. positions simultaneously provides a flat response curve over the full operating range of the Model 126 (refer to the specifications, subsection 2.2, Bandwidth). A flat response is useful for operation of the instrument as an ordinary ac voltmeter or as a broadband, low noise signal amplifier. When operating as a Lock-In, the flat response is recommended when the signal changes frequency by large factors during the course of the measurement.

Whenever possible, the bandwidth should be minimized so that the control settings closely bracket the signal frequency. While this can greatly reduce noise and interference which are not close to the signal frequency, bear in mind that there is also some effect on the amplitude and phase of the signal itself which could affect the apparent calibration of the phase controls and full-scale sensitivity. Figures IV-13 and IV-14 show the characteristics of each filter where E_{in} and E_{out} refer to the input and output voltage of each filter. E_{in} may or may not be the same for the two filters. To find the combined effect of the filters, multiply the Amplitude transfer fractions and add the Phase transfers.

Suppose the signal frequency is 1 kHz and that both line frequency interference (50 or 60 Hz) and wideband noise are present and large compared to the signal. By setting the LF Rolloff to 100 Hz and the HF Rolloff to 10 kHz, it can be seen that the line frequency pickup is reduced to less than 0.5 times the level at the input of the filter, using an f/f_{3dB} value of 60 Hz/100 Hz in Figure IV-13. The bandwidth of the wideband noise has been reduced from the full bandwidth of the Model 126 to approximately 10 kHz. Wideband noise fed into the input of the amplifier produces a noise voltage at the output not only proportional to gain, of course, but also proportional to the square root of the bandwidth. In this case, the noise voltage at the output of the Signal Channel has been reduced by a factor of $\sqrt{210\text{ kHz}/10\text{ kHz}}$ or about 4.4.

Now what about the amplitude and phase shifts of the signal itself? For the LF Rolloff Phase Transfer, f/f_{3dB} is 1 kHz/100 Hz or simply 10. As shown in Figure IV-13, there is about a 7° lead imposed on the signal and negligible amplitude loss by the LF Rolloff under these conditions. The f/f_{3dB} value for the HF Rolloff is 0.1 which produces a similar effect except that the signal lags 7°. Although not necessary, separating both the LF and HF Rolloffs by the same factor with respect to the signal frequency cancels out the total phase shift through the filters.

In some cases, maximum signal-to-noise ratio improvement can be obtained by operating the two filters at the same frequency and causing the signal to appear at that frequency. Again, the total phase shift is nominally zero, but in this situation, the signal amplitude is reduced by 6 dB or 0.5. Nevertheless, in the example of the preceding paragraph, the line interference is reduced to about 0.07 times the level at the input of the filter so that compared to the signal it is now much smaller. The amplitude reduction of the signal itself should be taken into account when considering full scale sensitivity.

OVERLOAD

The Overload lamp monitors two points in the Signal Channel, at the preamplifier output just ahead of the signal rolloffs, and at the final output just ahead of the Phase Sensitive Detector. Not only is the possibility of saturating the last signal amplifier monitored, but overloads in the early stages which might not pass through the rolloffs are also checked. The same lamp additionally indicates overload occurring in the output stages after the Phase Sensitive Detector so that all possible locations and types of overload are monitored.

If the meter reading is off-scale when the Overload lamp is on, the nature of the overload is obvious and the gain of the instrument should be reduced. The Output Gain should be reduced a step at a time until the light goes out. If the light still glows after the X1 position is reached, the Sensitivity switch should be rotated clockwise.

Output overload can also occur with an on-scale meter reading. This can happen if the Time Constant is set too low to allow sufficient filtering of peak voltages from the Phase Sensitive Detector. The remedy is to simply increase the Time Constant setting, and/or switch from 6 dB/octave to 12 dB/octave.

Input overloads ahead of the signal rolloffs can only be eliminated by using a higher setting of the Sensitivity switch. Overloads at the input of the Mixer can be eliminated by either changing sensitivity and/or setting the signal rolloffs to obtain a higher degree of rejection of unwanted signals.

Overload recovery time can be significantly reduced by avoiding the use of the MIN. setting for the Low Frequency Rolloff switch.

SPECIAL CONSIDERATIONS

Grounding: The preamplifiers all have two means of rejecting unwanted signals due to ground-loop currents. The most effective property of the preamplifiers is their extremely high common mode rejection when using the A-B input mode; that is, almost total rejection of unwanted signals which appear at both inputs at the same phase and amplitude. The other property is a unique arrangement of the grounding of the preamplifier within the Model 126 mainframe.

A ground loop is a conduction path (Figure IV-15) around the signal ground and chassis ground. Current in the loop is caused, usually, by environmental electric and magnetic fields that cut the loop, the most notorious being at the power-line frequency. In addition to induced current, other sources of current are galvanic activity, slight differences of potential between the chassis grounds of the signal source and preamplifier, etc. This ground loop current develops voltage in the signal ground resistance and this voltage can be amplified along with the signal if steps are not taken to prevent this.

P.A.R. preamplifiers can greatly reject ground loop voltages even when operated with a single-ended input (A only, for instance). The preamplifiers have a 10 ohm resistor between the signal input BNC connectors' outer shells and chassis ground across which nearly the entire ground loop voltages appear. Amplification is with respect to signal ground (which is very nearly identical at each end of the shield of the cable because of this scheme) and ground loop voltages are thereby excluded to a large degree.

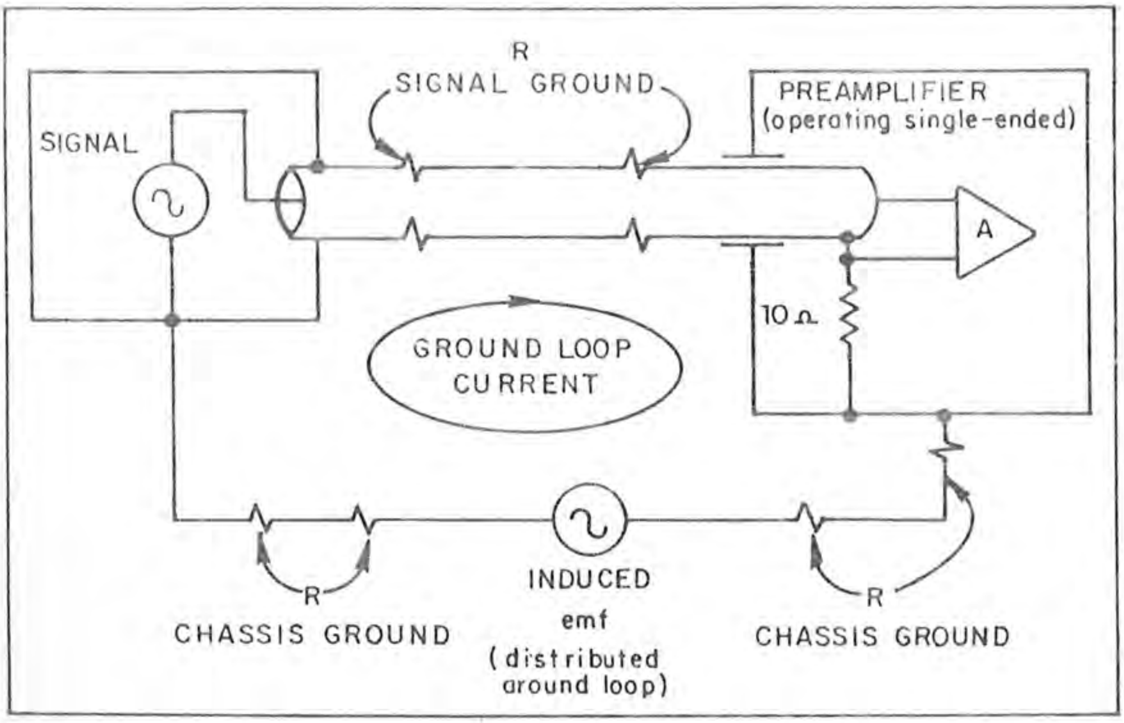


Figure IV-15. GROUND LOOP VOLTAGE REJECTION USING SINGLE ENDED INPUT

Figure IV-16 represents a typical situation using the differential inputs (A - B) of a preamplifier. With this connection, even the remainder of the ground loop voltage, which appears at equal amplitudes at both signal terminals, is rejected. The differential inputs are such that A is amplified with respect to B and not to ground. Since both inputs are very high impedance compared to the wires connecting them to the signal, there can be no current loops to cause the voltage at the preamplifier inputs to be any different than the voltage at the signal source.

Power Line Interference: Operation at the power line frequency is not recommended for low level measurements. Although much care has been taken with the internal design of the Model 126 and the preamplifiers (line frequency voltage due to internal power supply is less than 20 nV rms referred to the Direct inputs of a Model 116 Preamplifier) it is difficult to adequately shield the experiment's signal source to maintain such a low pick-up level. Operation at any harmonic frequency of the power line, from about the second through tenth, also requires more than usual care in shielding.

4.4 REFERENCE CHANNEL

MODES OF FREQUENCY CONTROL

INT/VCO: The frequency of the internal reference oscillator is set by a combination of the digital dials and the range switch. The digital dials may normally be set anywhere from 0.2 (which yields 20 Hz on the X100 band) to

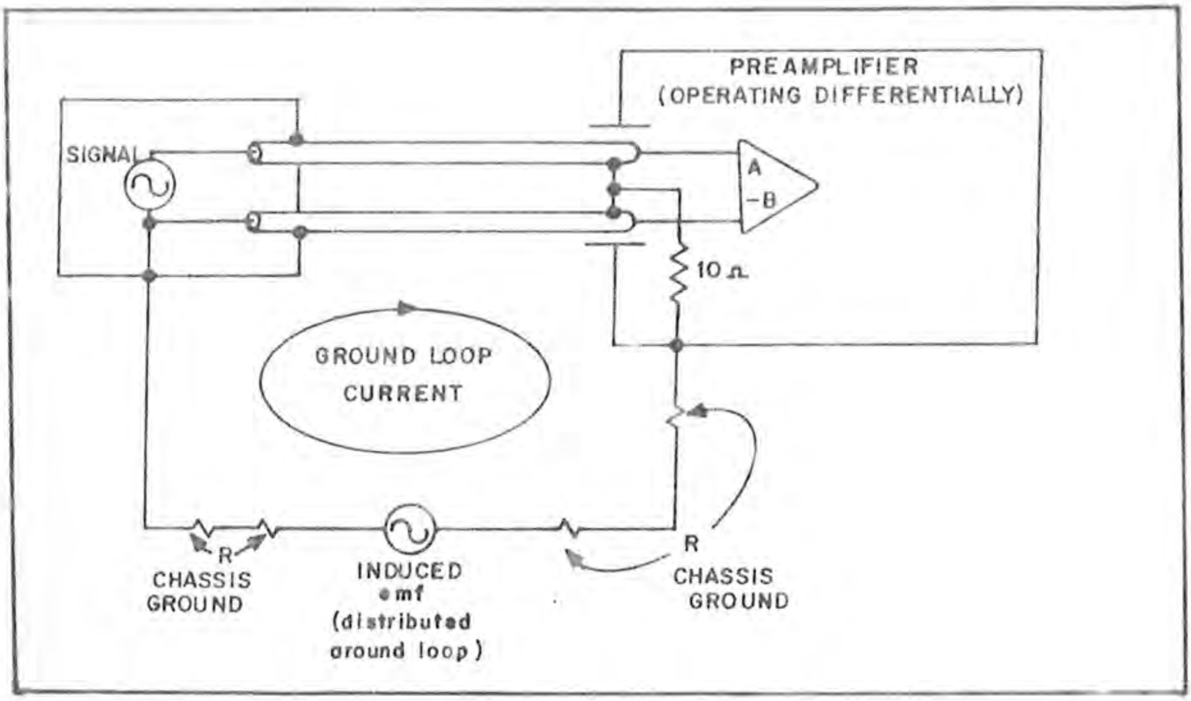


Figure IV-16. GROUND LOOP VOLTAGE
REJECTION USING DIFFERENTIAL INPUTS

10.0 (giving 1 kHz). In addition, an ADD 10 control which is concentric on the same band with the units dial can be used with the digital dials to effectively extend the range from 10.0 through 20.0. 1.5 kHz on the X100 band, for instance, is produced by setting the controls for 5.0 and ADD 10. Alternately, the X 1k band could be used with digital settings of 1.5 (NORM./ADD 10 switch set to NORM.).

The VCO method is operated in conjunction with the digital dials according to the following relation:

frequency = (Digital Dial Setting + 2 x VCO voltage) x Band Setting (3)

The nomogram shown in Figure IV-17 can also be used to determine the effective dial setting which can then be multiplied by the band setting to arrive at the frequency.

Suppose it is desired to have the oscillator scan from 500 Hz to 1.5 kHz and back repetitively at a uniform rate of 10 Hz per second, what dial settings and input voltage would be required? One possibility would be to set the dials to 5.0, X100 and feed in a triangle wave which has peaks at 0 V and +5 V. When the triangle wave is at 0 V, a frequency of 500 Hz is obtained; when the triangle wave is at +5 V, it can be seen from Equation (3) that 1.5 kHz is produced. Since a 5 V change in the VCO input voltage causes a 1 kHz frequency change on this band, this change should be produced over a time of

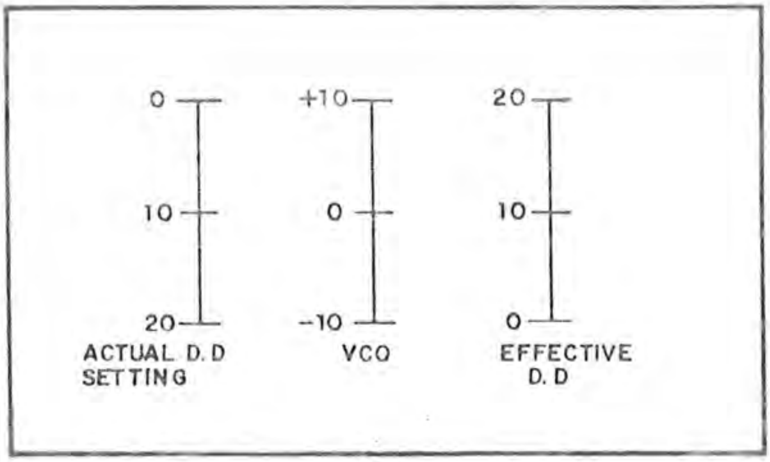


Figure IV-17. EFFECTIVE DIGITAL DIAL SETTING NOMOGRAM

100 seconds in order to achieve a 10 Hz per second slew rate so that the total period for the complete triangle wave should be 200 seconds.

EXT: A sophisticated tracking circuit automatically locks the internal reference oscillator to both the frequency and phase of an external waveform fed into the Reference Channel IN connector (see subsection 2.2, Specifications). Two common situations exist which demand operation of the Model 126 in the EXT. mode: when an external frequency reference must be used but does not have a waveshape suitable for exciting the experiment, e.g., a train of narrow pulses. And when the sine wave available from the Model 126 is not appropriate to drive certain experiments such as those requiring a great deal of power so that external reference drive equipment having the necessary characteristics is employed. The Model 126 should then be locked to the external equipment, unless, of course, the external equipment itself can be locked to the Model 126 operating in the INT/VCO mode.

The Reference Unlock lamp indicates that the tracking circuitry is not locked to the external frequency. Acquisition times vary according to the band setting and according to the difference between the external frequency and the frequency of the internal oscillator. Table II-2 in the Specifications section shows both the frequency range over which the tracking circuitry will perform well for each band and the maximum time to lock to the external reference. By operating in the INT/VCO mode for a moment at the approximate frequency of the external reference, the internal oscillator can be preset, greatly reducing the acquisition time when operating on the lower bands.

The frequency of the external generator must not change more rapidly than the Model 126 can track it (see Figure II-1). Provided the Reference Channel maintains a fixed frequency and phase relationship with the signal, detection is essentially the same as for fixed frequency signals.

It can be seen that there is considerable overlap of the frequency ranges of the bands. This often makes it possible to operate on either of two, or even three bands for a given frequency, e.g., 2.1 kHz on the X100, X1k, or X10k band. If the external frequency changes rapidly, it may be necessary to operate on the highest band which covers that frequency since the slewing rate is very much greater on the higher bands.

In either of the external modes, there is no offset produced by a change in symmetry of the synchronizing input waveform no matter how great that change might be. Response of the Phase Sensitive Detector to harmonics is given in Table IV-1 on page IV-25 and is also not affected by any change in symmetry.

EXT. f/2: Operation of the Reference Channel in this mode is essentially the same as for the EXT. mode except the internal oscillator locks to twice the frequency of the external reference. The maximum input frequency in this mode is 105 kHz.

This mode makes it possible to immediately take direct measurements of signals at the second harmonic frequency of the external reference.

Special Considerations: Very high speed circuitry is employed in the external modes of operation in order to reduce the phase difference between the internal oscillator and the external reference to an absolute minimum even at the maximum operating frequency of 210 kHz. For this reason, some care should be taken to insure that the "frequency" actually fed into the Reference Channel is the desired frequency and not something higher. One way of inadvertently introducing additional crossings of the mean is to connect a waveform with very fast rise-fall times to the Reference Channel input without properly terminating the interconnecting cable. Improper termination could produce ringing which would be interpreted by the Model 124 as a higher frequency than the operator expected. Another way of introducing additional crossings of the mean is to allow interference from motors, light switches, or any spark generating electrical equipment to contaminate the external reference. Use of shielded cable to the Reference Channel input should prove helpful. Whenever possible, the level of the external reference should be several volts so as to be large compared to any unwanted interference.

PHASE CONTROLS

The high resolution potentiometer covering a range of 0 to 100 degrees in conjunction with the Quadrant switch determines the phase of the synchronous detection process with respect to the Reference Channel output (and input when using the external modes).

The Phase Sensitive Detector produces a dc output proportional to the component of the signal at the phase set on the controls. When the controls are adjusted for maximum meter indication, the phase of the signal can be read from the controls while the amplitude of the signal can be read on the meter.

Quite often, meter fluctuations due to noise cause difficulty in finding the setting which gives a maximum reading. When this is the case, it is normally more accurate to adjust for a null output which occurs when the controls are set at either plus or minus 90 degrees with respect to the phase of the signal. It is then a simple matter to switch to the adjacent quadrant which

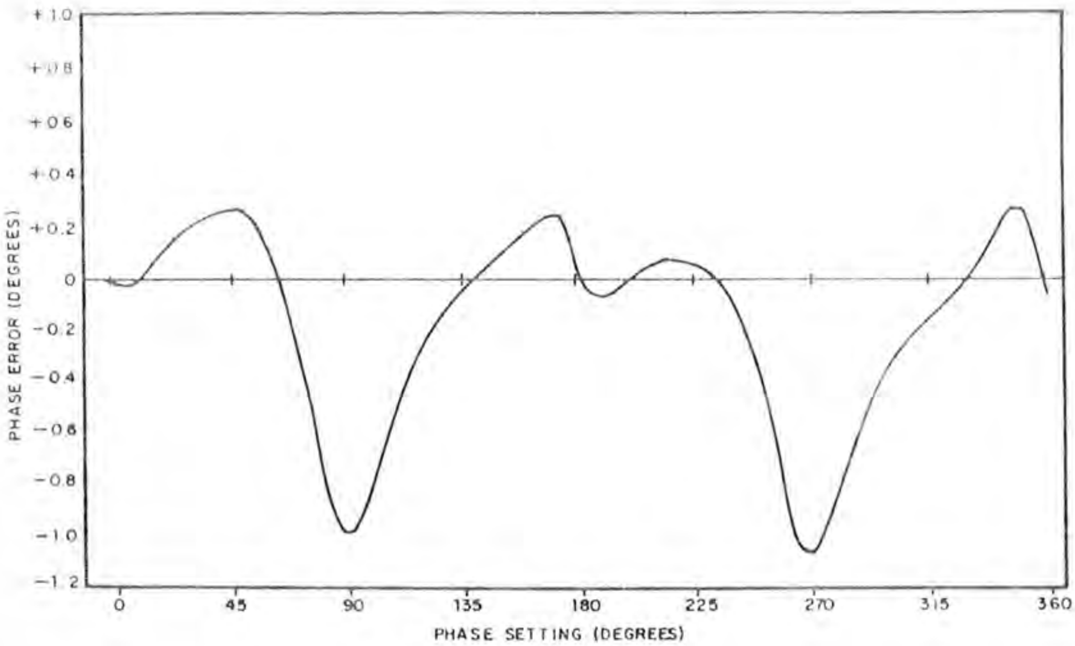


Figure IV-18. TYPICAL ACCURACY OF THE PHASE CONTROLS (20 Hz to 20 kHz)

yields the positive deflection on the meter and read the phase from the controls.

Absolute accuracy and resolution of the phase controls are described in the Specifications, subsection 2.2. Typical accuracy of the phase controls is shown in Figure IV-18. Beyond the frequency range given (20 Hz to 20 kHz) in the figure, the phase controls may not indicate the actual phase of the signal because of phase shifts introduced in the Signal Channel at extreme frequencies. Even at middle frequencies, it should be noted that the settings of the Low and High frequency Rolloff controls can have an effect on the phase of the signal. A calibration can easily be made at any frequency by connecting the preamplifier input to the reference drive at the experiment and adjusting the phase controls for a maximum output. The phase controls then indicate the net phase shift in both the Signal and Reference Channels. This shift is then subtracted from any experimental phase readings taken while operating at that particular frequency.

OUTPUTS

The Reference Channel OUT connector provides a sine wave output at the internal oscillator frequency. The amplitude can be varied from "zero" (actually a few microvolts minimum) to 1 volt rms by adjusting the logarithmically tapered Level potentiometer. The settings corresponding to various rms output amplitudes are marked on the front panel. The output impedance varies from about 1 kilohm to 1.6 kilohm depending on the setting.

Instruments with the Power Amplifier Option installed produce an output level of a factor of ten greater than that shown on the front panel, i.e., 0 to 10 volts rms. The output impedance is fixed at 600 ohms regardless of the setting of the Level control. A standard Model 126 can be converted to include the Power Amplifier Option by merely changing the Power Supply Board to one which has the Option installed on it. No chassis wiring changes are necessary.

4.5 PHASE SENSITIVE DETECTOR AND OUTPUT STAGES

FUNCTION-MODES OF OPERATION OF THE DETECTOR

Phase Sensitive Detector (PSD): When the Function switch is in any of the three PSD positions, the Model 126 operates as a Lock-In Amplifier. As stated in subsection 4.3, the Block Diagram and Front Panel Layout discussion, the detector convolutes the signal frequency in such a way that the output of the detector is the sum and difference frequencies of the signal and the reference. It can be seen then, that if the signal and the reference are at the same frequency, which is normal, one of the output frequencies of the detector is zero or simply dc. Noise or other interference is normally not exactly at the same frequency as the signal and therefore cannot produce zero frequency at the output of the detector. The dc which is produced is proportional to the amplitude of the in-phase component of the signal and is sent through the low pass filters of the Output Stages to drive the meter and the various output connectors.

There is no dc output produced by the component of the signal which is in quadrature phase with respect to the reference. The dc output due to a sine wave signal of amplitude A and with a phase difference Ø with respect to the reference is related by the following:

$$\text{DC VOLTAGE} = A \cos \phi .$$

Since the phase of the reference drive to the detector can be independently varied by the front panel Phase controls, Ø can be adjusted to any value.

To be more precise, the reference input to the detector is actually a square wave composed of a certain amplitude of fundamental frequency f, and lesser amplitudes of all the odd harmonics - third harmonic at 1/3 the amplitude, fifth harmonic at 1/5 the amplitude, seventh harmonic at 1/7 the amplitude, etc. Therefore, if the incoming signal has any odd harmonic components, the detector will also convert them to dc at its output, since their difference with respect to the corresponding harmonics of the reference is zero. However, because the detector's output is actually the product of its two inputs, the dc produced by harmonic signals is less than that produced by the fundamental. In fact, the response of the detector to the odd harmonic signals is only 1/3 for third harmonic signals and similarly smaller for the higher ordered harmonics.

Great care has been taken to minimize the detector's response to even harmonics. The typical harmonic response when operating with maximum Signal Channel bandwidth is shown in Table IV-1. Reducing the High Frequency Rolloff setting can significantly diminish the response to harmonics.

ACVM: When the Function switch is in the ACVM position, the Model 126 oper-

	PHASE SETTING			
Harmonic	0°	90°	180°	270°
2nd	0.15%	0.5%	0.2%	0.5%
3rd	35%	35%	35%	35%
4th	0.13%	0.55%	0.25%	0.7%
5th	15%	15%	15%	15%

Table IV-1. TYPICAL HARMONIC RESPONSE
OPERATING WITH MAXIMUM SIGNAL BANDWIDTH

ates as an ac voltmeter with certain advantages over a conventional acvm. For one, the voltmeter can be more frequency selective by adjusting the bandwidth of the Signal Channel Rolloff controls. Also, the meter response time can be easily adjusted over a wide range by use of the Time Constant switch permitting accurate measurements of fluctuating signals. The meter circuit has a "crest factor" (ratio of peak amplitude to rms amplitude) greater than five, allowing measurement of signals having small average and large peak amplitudes without overloading. The accuracy of the ACVM is ±1% from 2 Hz to 20 kHz, increasing to ±10% at 210 kHz; +0, -30% below 2 Hz.

Phase Meter Option: The Phase Meter Option, which may be switched on by a toggle switch on the rear panel, enables direct readout of the phase of any symmetrical input signal. The meter deflection is a maximum of plus or minus 90% of full scale on the X10 position of the Function switch, corresponding to the signal being exactly in-phase or exactly out-of-phase with the setting of the front panel Phase controls. For instance, if the Phase control is set to 90°, the meter will read the phase of the signal with respect to the Reference Channel Output with 0-to-90° lead presented linearly as 0-to-90% on the positive meter scale, and 0-to-90° lag presented on the negative meter scale.

Below 10 kHz, the Sensitivity switch can be set to 100 μV for all signal levels from a few microvolts to the overload point (which is monitored by the Overload Lamp) at approximately 70 mV. At higher frequencies, better results can be obtained by setting the Sensitivity switch to the approximate level of the input signal. In any case, the meter indication is relatively unaffected by amplitude. At midrange frequencies, for example, the phase indication on the 10 mV setting of the Sensitivity switch is nearly constant with input levels ranging from about 50 μV to 70 mV, assuming, of course, that the phase of the signal is held constant at all levels.

OUTPUT GAIN/NONCOHERENT SIGNAL OVERLOAD

When operating the Model 126 as a Lock-In Amplifier, i.e., in one of the PSD positions of the Function switch, the least amount of Output Gain should be employed. The actual full scale Sensitivity is affected by the Output Gain

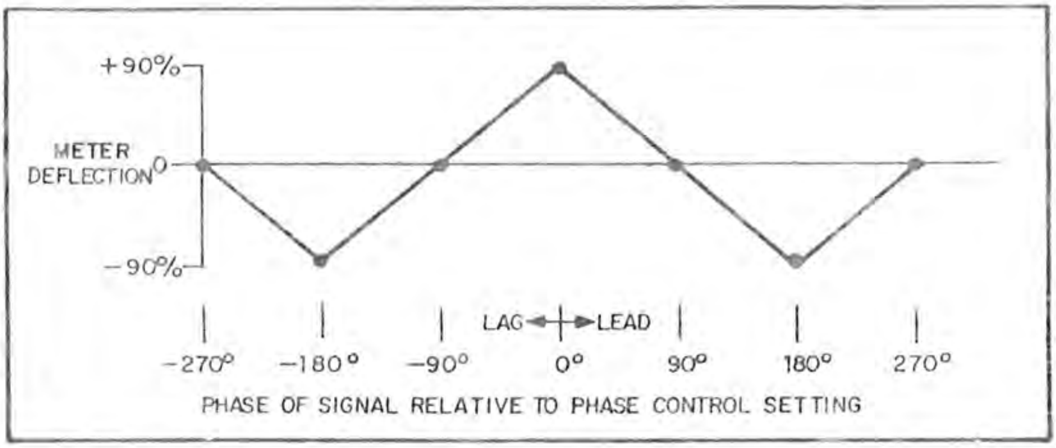


Figure IV-19. METER INDICATION AS A FUNCTION OF PHASE DIFFERENCE BETWEEN INPUT SIGNAL AND PHASE CONTROL SETTING

setting so that, for example, 1 mV full scale can be obtained by setting the Full Scale Sensitivity switch to 1 mV and the Output Gain to X1, or by setting the Full Scale Sensitivity switch to 10 mV and the Output Gain to X10, or finally, by using settings of 100 mV and X100 (all settings based on use of a Model 117 Preamplifier).

This ability to manipulate the distributed gain in the instrument has been incorporated in the design so that the performance can be made optimum for various signal-to-noise ratio conditions of the incoming signal. For relatively clean signals, where the noise and other noncoherent voltages are not greater than three times the required full scale sensitivity, the X1 Output Gain setting can be used. This results in an extremely stable output zero since the temperature dependent drift of the dc amplifiers has minimal amplification. If the signal is very small compared to the noncoherent voltages, lowering the Full Scale Sensitivity switch so as to obtain a greater meter deflection may result in overload of the ac amplifiers in the Signal Channel. In this case, the use of a higher Output Gain will boost the dc output of the instrument and, thus, the meter indication, without causing the Signal Channel to overload. The drift in the dc amplifiers will be magnified, however, so that the output zero will not be as stable under this condition. The relation between Output Gain, stability, and overload capability is given in subsection 2.2, Specifications.

The Model 126 tolerates noncoherent signals unusually well. The presence of large nonsynchronous signals produces only a slight shift in the output zero of the Lock-In Amplifier as shown in Figure IV-20.

OUTPUT FILTER MODES - TIME CONSTANTS AND INTEGRATOR

The conventional function of the Output Stages is to operate as a low pass filter, rejecting all components at the output of the Detector except the dc voltage (which is proportional to the in-phase component of the signal of interest). The Output Stages of the Model 126 have the additional capability of performing a true integration of the Detector's output. A Normal/Hold switch and a pushbutton Reset switch provide full control of the Output Filter in either the Low Pass Filter or the True Integrate mode.

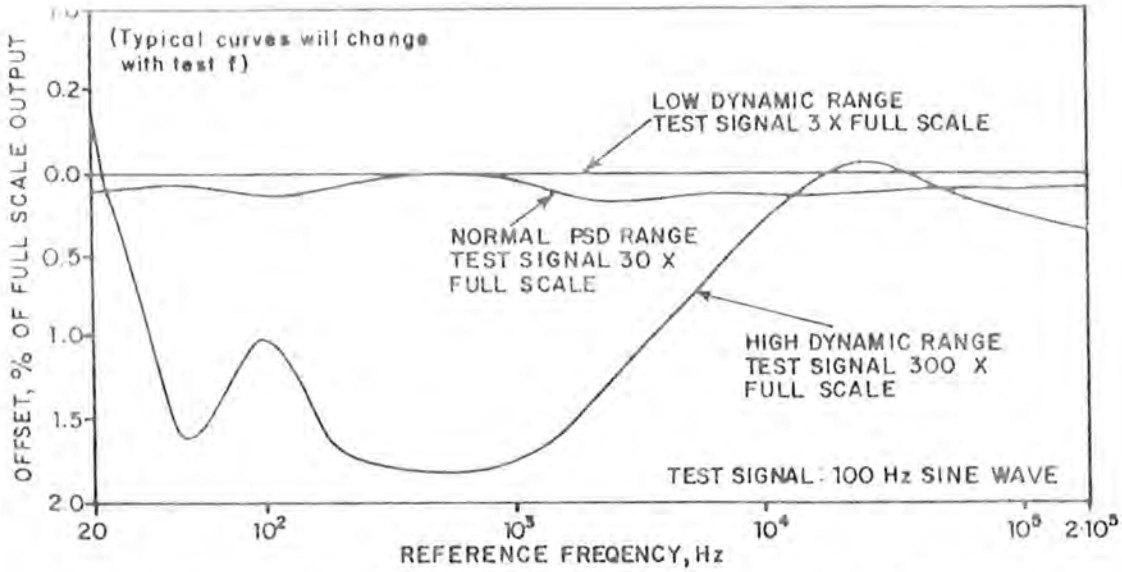


Figure IV-20. OUTPUT OFFSET DUE TO NONSYNCHRONOUS SIGNALS

Low Pass Filter: The filter can be operated either single section with a 6 dB/octave rolloff rate, or double section with a 12 dB/octave rolloff rate as selected by the inner concentric knob on the Time Constant switch. The outer knob sets the time constant of each section simultaneously with a range of 1 ms to 300 s plus MIN. (slightly less than 1 ms) and EXT. which permits the use of external capacitors to determine the time constant.

The Time Constant switch should be set so that the output, as read on the meter or on an external monitor, is as noise-free as required. In general, signals which are contaminated by large amounts of noise near the signal frequency require larger time constants. In many cases, where the characteristics of the noise are known for frequencies near the signal frequency, the noise voltage referred to the input can be determined for any given time constant by one of the following relations:

$$E_n = e_n\sqrt{\frac{1}{4TC}} \qquad \text{for 6 dB/octave}$$

$$E_n = e_n\sqrt{\frac{1}{8TC}} \qquad \text{for 12 dB/octave}$$

where e_n is the noise voltage at input in volts/$Hz^{\frac{1}{2}}$,
TC is the Time Constant in seconds, and
E_n is the output noise in volts referred to the input.

The term inside each radical is referred to as the "equivalent noise bandwidth" or ENBW of the Low Pass Filters.

For example, suppose that a 1 mV signal is being measured with the controls set for a 1 mV full scale sensitivity. Also suppose that 10 mV of noise, occurring over a frequency range which includes the signal frequency, is present at the input. If it is known that the noise has a characteristic value of, say, 100 $\mu V/Hz^{\frac{1}{2}}$ in the region near the signal frequency, the effect of this can be readily calculated. Using a typical setting of 6 dB/octave and a time constant of 1 s, the above equation gives the following:

$$E_n = \frac{100\ \mu V}{\sqrt{Hz}} \sqrt{\frac{1}{4(1s)}} = 50\ \mu V\ rms$$

A considerable improvement in the signal to noise ratio is clearly evident. The signal to noise ratio at the input is, of course, 1 mV:10 mV or 1:10. After processing, the noise is reduced to a level of 50 µV referred to the input for an effective signal to noise ratio of 1 mV:50 µV or 20:1. The improvement is a factor of 200 by simply introducing a one second time constant in the signal path!

As shown in Figure IV-21, the 6 dB/octave filter exhibits a 3 dB attenuation at $1/2\pi TC$ on the frequency axis. On the time axis, it has a step-function response of $1 - e^{-t/TC}$ seconds. The rise time from 10% of full amplitude to 90% of full amplitude is 2.2 TC, and from 0% to 95% is 3 TC.

Also as shown in Figure IV-21, the 12 dB/octave filter has a 6 dB attenuation at a frequency equal to $1/2\pi TC$. On the time axis, it has a step-function response of $1 - (1 + t/TC)e^{-t/TC}$. The rise time from 10% to 90% of full amplitude is 3.3 TC, and from 0% to 95% is 4.8 TC.

External Time Constant: For special intermediate values of time constant or for values in excess of 300 seconds, place the Time Constant switch in the EXT. position and connect a pair of capacitors of equal value between pins 8-9 and 10-11 of the rear panel octal socket. To determine the time constant for this external mode, multiply the single capacitor value (in Farads) by 30 megohms. The capacitors should be low-leakage film types (mylar, polycarbonate, polystyrene, or teflon, but not electrolytic or tantalum) rated at 25 volts or higher.

Integrator: Placing the inner concentric Time Constant control to the center position enables the Output Stages to integrate the output voltage from the Detector with respect to time with the following characteristics:

$$E_{out} = \frac{1}{TC}\int_o^t E\ dt = \frac{1}{TC}\ Et,$$

where E is the voltage which would be obtained if the Lock-In Amplifier were used in the more conventional mode with the output stages operating as simple dc amplifiers rather than as an integrator. That is,

$$E = \frac{\text{input signal voltage x cos } \phi}{\text{Full Scale Sensitivity}}$$

ϕ is the phase of the signal with respect to the phase of the reference drive to the detector.

In plain language, the equations indicate that if the Model 126 is operated as a conventional Lock-In Amplifier measuring a signal which is say, 10% of full scale, the same output indication, 10% of full scale, would be obtained by operating in the Integrate mode for a period of time equal to one time constant. If the integration time were made equal to 5 time constants, then the output indication would be 50% of full scale. The integration time can be controlled by releasing the Reset pushbutton at time t = 0 and then placing the Normal/Hold switch to Hold at the end of the desired time. If the input amplitude varies during the integration time the final voltage is based on the average signal amplitude multiplied by the number of elapsed time constants.

OUTPUTS

The final dc output voltage, after all processing, is available at a front panel BNC connector located near the Function switch. Plus or minus 10 volts dc corresponds to plus or minus full scale; the output impedance is 1 kilohm. The same voltage, through a separate 1 kilohm impedance, is also available at the rear panel Recorder binding posts. Loading either output will not affect the voltage at the other, nor will it affect the meter indication.

REFERENCE