CHAPTER III

EXPERIMENTAL SET UP; A TIME-SHARING TWO-BANDWIDTH RADIO POLARIMETER AT 35 MHz

3.1 INTRODUCTION

3.11 SPECIFICATION OF POLARIZATION IN TERMS OF STOKES PARAMETERS

The subject of Stokes polarization parameters and the instrumental techniques for their measurement has been reviewed by Cohen (1958a). Here, we propose to recapitulate briefly some aspects of polarization measurements. The four parameters that specify the state of polarization of a partially polarized radiation are the total intensity $I$, degree of polarization $m$, axial ratio $r$, and orientation angle of the polarization ellipse $\chi$; $r > 0$ means left-handed sense of polarization and $r < 0$ means right-handed (Figure 3.1). In general, partially polarized radiation can be uniquely resolved into unpolarized part $I_u$ and completely polarized part $I_e$ (Chandrasekhar 1955). The Stokes Parameters are a set of four numbers $I, Q, U$ and $V$; $I$ refers to the total intensity, and $Q, U$ and $V$ specify the polarized part of the radiation. They all have dimensions of intensity and, therefore, have an advantage of being closely related to the antenna measurements. They are defined by the following relationships:
Figure 3.1 Coordinate system
I = \bar{I}_e + I_u \quad \ldots (1)

Q = \bar{I}_e \cos 2\beta \cos 2\chi, \quad \ldots (2)

U = \bar{I}_e \cos 2\beta \sin 2\chi, \quad \ldots (3)

V = \bar{I}_e \sin 2\beta. \quad \ldots (4)

where

\beta = \tan^{-1} r. \quad \ldots (5)

The quantities \bar{m}, r, and \chi are given by

\bar{m} = \frac{\bar{I}_e}{I} = (Q^2 + U^2 + V^2)^{1/2}/I, \quad \ldots (6)

\sin 2\beta = V/\bar{I}_e \quad \ldots (7)

and \tan 2\chi = U/Q. \quad \ldots (8)

The values of \chi depend upon the sign of U and Q.

Figure 3.2 shows how \chi can be measured from -90° to +90°. Some of the known techniques involve measurement of Stokes parameters in terms of quantities of different nature like intensity, cross-correlation product and phase angle between the two orthogonal components of radiation. With the progress in digital data processing techniques by computers, digital radio polarimeters have been developed first at Stanford (Bhonsle et al. 1967 and Chin et al. 1971), and later at Boulder (Dodge 1972). Such polarimeters have been used to study the high time and frequency resolution polarization characteristics of type III solar radio bursts. There would be a distinct advantage if all the Stokes parameters would be measured only in terms of
Figure 3.2 Dependence of the orientation angle $\chi$ on the sign of $U$ and $Q$.

- $U$, $Q$ mean $U$, $Q = 0$
- $U^0$, $Q^0$ mean $U$, $Q < 0$
- $U^+$, $Q^+$ mean $U$, $Q > 0$
intensities so that the same receiver can be shared for their measurements.

3.12 POLARIZATION WORK AT AHMEDABAD

The radio polarimeter set up at Ahmedabad has been designed to operate on the so-called "time-sharing principle". The principle of operation of this polarimeter will be described in detail in section 3.2 of this chapter. Important specifications of this polarimeter are given in Table 3.1. The polarimeter has undergone a development in three distinct stages as mentioned below:

Operation of the polarimeter (i) at 25 MHz with a bandwidth of ±10 KHz from July, 1969 to June, 1970; (ii) near 35 MHz at two frequencies separated by 4 KHz with bandwidth of 800 Hz at each frequency, from January to March, 1972, and (iii) two bandwidth (7.5 and 12.5 KHz) polarimeter at 35 MHz from July, 1972 to this date.

3.2 PRINCIPLE OF OPERATION OF A TIME-SHARING POLARIMETER

In this section, we explain the principle of operation of a particular combination of a pair of cross-polarized linear antennas and the fast electronic-switching of the appropriate relative phase delays between them to simulate different polarizations. This fast-switching
### TABLE 3.1

**SPECIFICATIONS OF THE TWO-BANDWIDTH TIME-SHARING 35 MHz RADIO POLARIMETER**

**Antenna:**
- Cross-polarized Yagi
- Cross talk: -32 db
- Gain: 6 ± 1 db (expected)
- Beam width: $40^\circ \times 40^\circ$

**R.F. pre-amplifiers:**

<table>
<thead>
<tr>
<th></th>
<th>Channel A</th>
<th>Channel B</th>
</tr>
</thead>
<tbody>
<tr>
<td>Centre frequency</td>
<td>35 MHz</td>
<td>35 MHz</td>
</tr>
<tr>
<td>Gain (Maximum)</td>
<td>40 db</td>
<td>40 db</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>1 MHz</td>
<td>1 MHz</td>
</tr>
<tr>
<td>Gain control (Manual)</td>
<td>12 db</td>
<td>12 db</td>
</tr>
<tr>
<td>Noise figure</td>
<td>2.7 db</td>
<td>2.5 db</td>
</tr>
</tbody>
</table>

**Receiver:**

- Local oscillator frequency: 45.7 MHz
- Mixer: balanced
- Conversion gain: 12 db
- I.F.: 10.7 MHz
- I.F. gain: 80 db
- I.F. bandwidth: 50 KHz
- I.F. band pass filter bandwidths: 7.5 and 12.5 KHz
- Detector type: Square-law
Loss in square law detector : 3 db
Pen recorder time constant : 1 sec.
D.C. amplifier gain : 17 db
Integration loss : 20 db.
Operational amplifiers
  gain : Unity
Audio amplifier gain : 40 db
Audio amplifier bandwidth : 50 KHz
Power combiner isolation : 35 db
Two-way power divider isolation: 30 db
ON/OFF Ratio of modulator : 48 db
ON/OFF ratio of time demodulator: 30 db
of antennas becomes absolutely necessary because solar burst durations can be very small. All the Stokes parameters which specify the burst radiation must be sampled in a time short compared to the growth-time of the burst.

In order to measure the Stokes parameters, I, Q, U and V, we use the following relationship (Chandrasekhar 1955).

\[ I(\psi, \xi) = \frac{1}{2} \left[ I + Q \cos 2\psi + (U \cos \xi - V \sin \xi) \sin 2\psi \right] \]  ...(9)

where \( \psi \) is the angle between the plane of vibration and the V direction (Figure 3.1) and \( \xi \) is the relative phase delay introduced between the cross-polarized antennas. If \( \psi = \pm 45^\circ \) then equation (9) simplifies to

\[ I(\xi) = I + U \cos \xi - V \sin \xi \]  ...(10)

and \( Q = I_A - I_B \) and \( I = I_A + I_B \)  ...(11)

where \( I_A \) and \( I_B \) are the orthogonal intensities of radiation.

We can obtain two linear polarizations in vertical and horizontal directions, and two circular polarizations with right and left handed sense, by putting relative phase delays between signals induced in the antenna A and B, of 0, \( \pi \), \( + \pi/2 \) and \( -\pi/2 \), respectively. If we denote the powers received with these connections by \( I_V \), \( I_H \), \( I_R \) and \( I_L \), we can write by (10),
on combining equations (12) and (13) and (14) and (15) we get

\[ U = \frac{I_V - I_H}{2} \]  
\[ V = \frac{I_L - I_R}{2} \]  

Thus, the Stokes parameters are obtained by adopting the above-mentioned procedure and the polarization degree, the axial ratio, and the direction of polarization ellipse can be calculated by making use of equations (5) to (8).

3.3 DESCRIPTION OF A TWO-BANDWIDTH POLARIMETER AT 35 MHz

A functional diagram of a time-sharing two-bandwidth polarimeter at 35 MHz is given in Figure 3.3. A pair of equatorially mounted cross-polarized Yagi antennas is oriented at ± 45° angle with the North-South meridian at the local noon in the celestial sphere. The two antenna outputs at 35 MHz, after suitable amplification, are combined in sequence introducing appropriate relative phase delays to generate different polarizations as discussed in Section 3.2. The orthogonal intensities, \( I_A \) and \( I_B \), of radiation are also measured separately. After the amplification of the signal in the I.F. amplifier the signal is
Figure 3.3 Block diagram of Two bandwidth (7.5 and 12.5 KHz) polarimeter at 35 MHz.
divided into two channels by a power divider. The two I.F. signals are filtered separately by means of band-pass filters of 7.5 and 12.5 KHz. This enables us to make simultaneous measurement of the Stokes parameters in the two bandwidths. Thus, we have, after the square law detection, a sequence of six outputs namely, $I_V$, $I_H$, $I_R$, $I_L$, $I_A$ and $I_B$ for each bandwidth for every switching cycle. These intensities are time-demodulated into six separate channels by time-demodulators which are operated synchronously with the modulators. Finally, using operational amplifiers to perform additions and subtractions of various channels we record the Stokes parameters $I$, $Q$, $U$ and $V$ in each bandwidth on two 4-channel chart recorders, as shown in Figure 3.4.

We present more details of the various sub-systems of the polarimeter in the following subsections.

3.31 **ANTENNAS**

A pair of cross-polarized Yagi antennas, each with a folded dipole, two directors and a reflector are equatorially mounted and mechanically driven to track the Sun. Each antenna has been adjusted for the feed point impedance of 50 ohms resistive. The receiver is connected to the antennas through two pieces of RG8/U cables of equal length. The antenna parameters are listed in Table 3.1. The measured cross-talk was - 32 dB. The calculated half power beam width of each antenna was about $40^\circ \times 40^\circ$ and a front to back
Figure 3.4 Recorded Stokes parameters $I$, $Q$, $U$ and $V$ of type III bursts at 35 MHz.
ratio of at least 22 db. The first side lobe was at an angle of 80° and at least 16 db down with respect to the main beam.

3.32 RECEPTOR

3.32.1 35 MHz PREAMPLIFIERS

The two antenna outputs at 35 MHz are amplified by a preamplifier in each channel. The circuit for a preamplifier shown in Figure 3.5 is a 5-stage amplifier, consisting of one stage of a low noise figure transistor (2N3571) followed by four stages of CIL 911. A noise figure of 2.5 db and a gain of 40 db have been achieved. The parameters for both the preamplifiers are listed in Table 3.1. The 3-dB bandwidth of each preamplifier is about 1 MHz. The bandwidths of the preamplifiers are purposely kept little more than what is necessary so as to minimize the relative phase difference suffered by the signals after amplification. Further, phase shifters have been provided in each preamplifier to balance out any residual phase difference before the signal passes on to the modulators.

3.32.2 3-WAY POWER DIVIDER, SWITCHING SCHEME AND MODULATOR

The outputs of each 35 MHz preamplifier are divided equally in four parts by means of two four way power dividers. Since we need only three outputs per
Figure 3.5 35 MHz preamplifier circuit
channel, one output port of each power divider is terminated with a 50 ohm resistor. Power dividers with isolation between any two output ports of better than -30 db and phase equality of ± 1° were commercially available. The isolation among the output ports of the power dividers is important from the point of view of reducing the interaction among different electronic switches which introduce various phase delays by using appropriate lengths of coaxial cables.

An electronic switching system is incorporated to measure polarization parameters of rapidly varying phenomenon, such as bursts. It is assumed that polarization characteristics of the short duration bursts do not vary appreciably within the sampling period.

Figure 3.6 represents the time relationship between the modulators and demodulators in a given switching sequence that yields six pieces of information. They are controlled by common gate pulses each with duration of 4 milliseconds. In the first 4 milliseconds the modulators \( A_1 \) and \( B_1 \) and the demodulator \( V \) are "on". The voltages at 35 MHz obtained from two antennas are combined to give \( V \) component of the polarization at the output of the demodulator \( V \). In the next 4 millisecond interval \( A_1, B_3 \) and \( H \) are "on", and the \( H \) component is measured, and so on. Thus a set of six pieces of information is
Figure 3.6 Switching sequence used in Time-sharing radio polarimeter at 35 MHz at Ahmedabad.
picked up in about 24 milliseconds and the cycle repeats.

The circuit of the r.f. modulator used for introducing phase delays is shown in the Figure 3.7. It uses 1N34 diodes and serves as a solid state SPDT-type switch. This arrangement gives CN-OFF ratio of 46 db. The switches are followed by an amplifier stage to compensate for the insertion loss due to the modulators. The voltage gains of $A_2$ and $B_2$ are adjusted to $(\sqrt{2} - 1)$ times that of the other four so as to equalize overall gains of all the six channels. Thus, the outputs of the six components are all equal when the incident radiation is randomly polarized. After the signal is detected by a square-law detector, the equations (12) to (17), representing relationships between the four Stokes parameters and six intensity measurements, become

\[
I = I_A + I_B \quad \ldots(18)
\]
\[
Q = I_A - I_B \quad \ldots(19)
\]
\[
U = I_V - I_H \quad \ldots(20)
\]
\[
V = I_L - I_R \quad \ldots(21)
\]

The modulator outputs are combined by a power combiner and fed to a mixer as a sequence of pulses at 35 MHz.

3.323 MIXER, LOCAL OSCILLATOR AND I.F. AMPLIFIER

Figure 3.8 shows how the 35 MHz output from the combiner is converted to 10.7 MHz intermediate frequency by means of a mixer. A Clapp-type local oscillator at
Figure 3.7 Modulator circuit.
Figure 3.8 Circuit diagram of oscillator, mixer, I.F. amplifier, Hybrid power divider and square law detector.
45.7 MHz has been used to generate 10.7 MHz I.F. signal. For the sake of frequency stability, a buffer stage has been incorporated between the local oscillator and mixer.

3.324 HYBRID POWER DIVIDER AND BANDPASS FILTERS

Measurement of Stokes parameters in two bandwidths can be done by incorporating two band-pass filters of the desired bandwidths (7.5 and 12.5 KHz) at the two outputs of the two-way power divider which is fed by the signal coming from the I.F. amplifier. A lumped-circuit hybrid (Kurzok 1962) used as a two-way power divider is shown in Figure 3.8b. The lumped-circuit hybrid used in 35 MHz radio polarimeter was preferred to its distributed-circuit equivalent because of the intolerable lengths of transmission lines needed at 35 MHz, and also because the achieved isolation between the two outputs is greater. The isolation of the lumped-circuit hybrid used is frequency sensitive and is about -30 dB at 35 MHz. Since both the bandwidths used are very narrow the frequency sensitivity of the isolation did not pose a problem. Each coil was adjusted to resonate at 35 MHz and were laid out physically at right angles to prevent undesirable electro-magnetic coupling. The outputs are tapped from the output coils and fed to two crystal band-pass filters. Merimac PD-30-55 (Type 6347, MA) and PD-80-55 (Type 6347, MB) crystal band-pass filters and follow-up amplifiers yielded 3 db-bandwidths of 7.5 and
12.5 KHz respectively at 10.7 MHz central frequency. The frequency response as seen at the output of the square-law detectors for each channel is shown in Figure 3.9.

3.325 SQUARE-LAW DETECTORS

Since the Stokes parameters are measured in terms of intensity of the radio waves incident upon the antenna system, a detector with square-law characteristics is desirable in order to maintain a linear input/output relationship of the polarimeter. A solid-state low-output level square-law detector (Ohm and Snell 1963) shown in Figure 3.8b is a network of series and parallel 1N34A diodes. The overall square-law characteristics of the 35 MHz polarimeter (I-channel) is shown in Figure 3.10. It is seen that the output voltage is linearly proportional to the input power until the output at the square-law detector exceeds 8 m.v. So the quantum of post-detector gain required for satisfactory recording on chart is decided by the square-law characteristics of the detector. We operate the radio polarimeter at about 3 m.v. noise level at the output of the square-law detector.

3.326 TIME-DEMODULATORS

The outputs of the square-law detector are first amplified before they are time-demodulated. Two sets of time demodulators are used, one for 7.5 KHz bandwidth channel and the other 12.5 KHz channel. Two 2N995
Figure 3.5. I.F. Bandpass Characteristics of receiver.
Figure 3.10  INPUT/OUTPUT response of 35 MHz receiver.
transistors in an arrangement shown in Figure 3.11 are used to time-demodulate the signal. The synchronous pulses from the control pulse generator are fed inductively at the base of the transistors to operate the demodulators.

3.327 OPERATIONAL AMPLIFIERS

A direct-coupled amplifier, with a large open loop gain and bandwidth, used as an operational amplifier is shown in Figure 3.12. This consists of an emitter follower, two cascaded difference amplifiers which include a constant current generator and common feed back arrangement. The open loop gain was calculated to be about 2400. The frequency response with a resistive feedback was found to be reasonably flat up to 50 KHz. The measured drift per degree centigrade over a 15-minute period was about $20 \mu V$ referred to the input. This is quite good for a direct-coupled amplifier without a chopper. The gain can be adjusted by the ratio of the value of the feedback resistor $R_2$ to that of the input resistor $R_1$. The same circuit can be used as an integrator by using a resistor and a capacitor in parallel in the feedback loop. Analogue operations, involving additions and subtractions, to get $A - B$, $R - L$, and $V - H$ intensities are carried out by the operational amplifiers.

A d.c. amplifier, in which the gain is unity and negative, has been used as an inverter. This is achieved
Figure 3.11 Demodulator circuit.
D.C. AMP. $R_1=5.6\, \text{k}\Omega$, $R_2=470\, \text{k}\Omega$

FOR AUDIO AMP. COUPLE INPUT & OUTPUT THROUGH CAPACITOR

FOR INVERTOR $R_1=R_2$

FOR ADDER $R_1+R_1=R_2$

Figure 3.12 Operational amplifier circuit.
by making \( R_1 = R_2 \), where \( R_1 \) and \( R_2 \) are input and feedback resistors respectively. An arrangement, in which two inputs to a d.c. amplifier are fed through two resistors each equal to \( R_1 \) such that the resistance in the feedback loop \( R_2 = R_1 + R_1 \), is used to obtain an output which is a linear combination of the two inputs. All these operations can be carried out with better than 1 per cent accuracy and are operated at unity gain.

The desired outputs, \( I, Q, U \) and \( V \) (see Figure 3.4) obtained from various time-demodulated signals, after going through the process of electronic analogue operations, are integrated with a time constant of 0.5 S before they are finally recorded on two 4-channel chart recorders, as shown in Figure 3.3.

3.328 CHART RECORDERS

Two 4-channel recorders mechanically coupled are used to record the four Stokes parameters in each bandwidth. On one chart recorder time marks at every minute are also recorded. The chart recorder speed is maintained at 60 cms an hour but can be increased when required.

3.33 CONTROL CIRCUITS

The modulators and demodulators in the polarimeter are controlled by a square wave generator which feeds them the same enabling pulse simultaneously so as to simulate
various polarizations and separate them into six channels as shown in Figure 3.6. Pulse generator, ring counter and OR gates involved in these operations are described in the following subsections.

3.331 PULSE GENERATOR

A free-running emitter-coupled multivibrator has been used as a pulse generator, whose circuit diagram is given in the Figure 3.13. The frequency of the multivibrator is adjusted to 250 Hz. The output is taken from the collector of the second transistor. After differentiation and suitable amplification, the output of the multivibrator is fed to a Darlington pair before being clamped by a P-N-P transistor (AC 128) for getting a negative going pulse of -12 volts pulse height. A Darlington pair is used so that the clamping circuit does not load the free running multivibrator. The output from the pulse generator is taken through an emitter follower.

3.332 RING COUNTER

A chain of binaries in which the output of the first binary is coupled to the second and that of the second to the third, and so on, with the output of the last binary coupled back to the input of the first binary, is called a Ring Counter. In such a counter, each binary device receives its triggering signal directly from the source. This is done so that only one binary responds to the
Figure 3.13 Circuit diagrams of pulse generator and ring counter.
triggering signal. With each successive triggering the response of the binary system shifts from one to the other. This way sequential gating waveforms are obtained. In the system adopted by us, six binaries are used as shown in Figure 3.13 to generate a sequence of six gating waveforms in a period of 25 milliseconds. The last gating waveform obtained from the last binary device also resets the first binary device to make the system ready for the next cycle. The six outputs shown in Figure 3.13 are followed by emitter-coupled monostable multivibrators, shown in Figure 3.14, in which the feedback has been provided through a common resistor. The d.c. voltage derived from a potentiometer controls the collector current which, in turn, controls the gate width of the wave forms. We have adjusted a pulse width of 2 milliseconds and a gap of 2 milliseconds is provided to avoid the cross-talk between different channels.

3.333 "OR" GATES

"OR" gates using 1N66 diodes are incorporated in the control pulse generator to generate sequential gate waveforms for the modulators and time-demodulators. Arrangement of OR gates to work out appropriate relationship between the modulators and time-demodulators is shown in Figure 3.14. The sequence of the gate waveforms in a period of 25 milliseconds, available for different
Figure 3.14 Circuit diagrams of monoshots and "OR" gates.
modulators, is also shown in Figure 3.14. The interaction between the different channels is low because of high reverse-to-forward resistance ratio of 1N66 diodes.

3.4 CALIBRATION PROCEDURE

3.41 PHASE AND GAIN ADJUSTMENTS

The relative phase difference between the two orthogonal components of the incident radiation must remain unchanged at the outputs of the preamplifiers. The phase controls provided in the two preamplifiers were adjusted to equalize the intensities on the R and L channels when the two inputs of the polarimeter are fed from the two outputs of an ISOT connected to a diode noise generator. This arrangement simulates linear polarization in the vertical direction. The same test was repeated after feeding the noise through two equal lengths of co-axial cables which are used to connect the antenna feed points to the inputs of the polarimeter. Further, linear polarization was tested by radiating a c.w. signal at 35 MHz from a $\lambda/8$ test dipole and orienting the Yagi antennas such that the antenna elements are oriented at an angle of $\pm 45^\circ$ to the test dipole (Bhonsle and McNarry 1964a). This test was repeated by radiating a linear polarization from about 300 ft overhead the antenna with the help of a tethered balloon. Any phase inequality between the two channels can be measured by a R.F. vector voltmeter and adjusted
to $\pm 1^\circ$ by the phase shifters. Also the bandwidth of both the preamplifiers is sufficiently wide so that the phase inequality between the two channels does not become a sensitive function of the frequency. The phase adjustment was done first and then the gains of the two preamplifiers were equalized. Two similar noise diode generators are available at the inputs of the two preamplifiers for gain calibration. To match the gains of the two preamplifiers, equal amounts of noise powers are fed to the inputs of the preamplifiers and the gain controls provided in each preamplifier are adjusted to produce equal outputs. The post detector gain of the 7.5 KHz bandwidth channel is greater than that of the 12.5 KHz channel by a factor of 1.6 ($12.5 / 7.5$) so as to obtain equal deflections on both the channel.

3.42 DAILY CALIBRATIONS

The polarimeter recordings are calibrated daily against the output of two similar temperature-limited diode noise generators. Since we are directly measuring the Stokes parameters we simulate some typical polarizations to calibrate the four channels, that is, I, Q, U and V in the two bandwidths. By introducing different phase delays between the noise generators and the inputs of the radio polarimeter we simulate the conditions of completely unpolarized and 100 per cent linearly and circularly polarized
radiation. The phase adjustments are carried out once in a week.

3.5 POTENTIAL SOURCES OF ERROR

The five main sources of errors are: (1) the phase inequality impressed by the instrument between the two orthogonal components of the radiation, (2) the antenna nonzero cross-talk, (3) the nonzero cross-talk as a result of the interaction between the various subsystems of the receiver, (4) the reflections from the local objects and the ground, and (5) the statistical nature. These sources of errors are discussed separately.

3.5.1 ERROR DUE TO RELATIVE PHASE DIFFERENCE BETWEEN THE TWO ORTHOGONAL COMPONENTS

Errors resulting in the relative phase difference between the two channels can arise mainly because of possible non-resistive impedance of antennas and unequal phase shifts introduced by the two preamplifiers and the coaxial cables connecting the antenna feed points to the inputs of the polarimeter. We consider this error under the assumptions of complete isolation between the two antenna outputs and between the various subsystems of the receiver.

Let \( \Delta \) be the error in the relative phase difference between the two orthogonal components. Using relation (10) we obtain the expressions for \( I_H, I_R, \) and \( I_L \) as
\[ I_V = I + U \cos \Delta + V \sin \Delta \] \hspace{1cm} \cdots(22)

\[ I_H = I - U \cos \Delta + V \sin \Delta \] \hspace{1cm} \cdots(23)

\[ I_R = I + U \sin \Delta - V \cos \Delta \] \hspace{1cm} \cdots(24)

\[ I_L = I + U \sin \Delta + V \cos \Delta \] \hspace{1cm} \cdots(25)

Let the apparent Stokes parameters and other polarization parameters in the presence of \( \Delta (\neq 0) \) be represented by primes. From the equations (22) to (25) we obtain

\[ U' = U \cos \Delta + V \sin \Delta \] \hspace{1cm} \cdots(26)

\[ V' = V \cos \Delta + U \sin \Delta \] \hspace{1cm} \cdots(27)

The apparent values of \( I \) and \( Q \) are not in error because any error introduced in the relative phase difference between the two orthogonal components does not change the intensity of two orthogonal components thus

\[ Q' = Q \] \hspace{1cm} \cdots(28)

\[ I' = I \] \hspace{1cm} \cdots(29)

Equations (26) to (28) are used to compute apparent polarized component of the intensity \( I_e \), as

\[ I'_e = (Q' + U'^2 + V'^2)^{1/2} = I_e \] \hspace{1cm} \cdots(30)

so \( m' = m \) \hspace{1cm} \cdots(31)

where in \( m' = \text{degree of polarization when } \Delta \neq 0 \)

As long as the antennas are oppositely polarized, the polarization degree is correctly found and this conclusion does not depend upon the magnitude of \( \Delta \).
The apparent value of the axial ratio $r$ is calculated from the expressions (27) and (30). For $\triangle \ll 1$

$$\sin 2\beta = \frac{V'}{I_e} = \frac{V \cos \alpha + U \sin \Delta}{I_e} = \sin 2\beta \left(1 - \frac{\Delta^2}{2}\right) + \cos 2\beta \sin 2\chi \quad \text{(32)}$$

when the wave is not circular, that is $r \neq 1$, we have

$$\delta r \lessapprox \frac{1}{2} (1 + r^2) \sin 2\chi \quad \text{...(33)}$$

and when the wave is circular ($r=1$)

$$\delta r \lessapprox \Delta \quad \text{...(34)}$$

For noncircular wave $\delta r = 0$ for certain orientations. If $\delta r$ has to be less than .01 then $\Delta$ should in no case exceed .02 radians. The phase controls provided in the preamplifiers are adjusted to equalize the phase in the two channels so that this error is negligible.

3.52 NONZERO CROSSTALK

The crosstalk between the different subsystems of the receiver can be a major source of error. The crosstalk between the two preamplifiers, mainly through the power combiner, is better than -30 db. The crosstalk caused by the leakage at the modulators and demodulators is better than -46 db. The interaction between the two bandwidth channels through a two-way power divider is better than -30 db. Most of the antenna crosstalk is
because of the incidental nonorthogonality of the two crossed dipoles.

Let the crosstalk be a factor of $k$; that is, when one volt is applied to terminal $A$ of Figure 3.1, $k$ volts appear at terminal $B$. The voltages induced on the two antennas can be written in terms of instantaneous voltages in the presence of the crosstalk as

$$E'_1 = E_1 \sin(\omega t) + k E_2 \sin(\omega t - \delta) \quad \ldots (35)$$

$$E'_2 = k E_1 \sin(\omega t) + E_2 \sin(\omega t - \delta)$$

Introducing a phase delay $\xi$ in the second component of the equation (35) we get

$$E'_1 = a \sin(\omega t - \theta_1) \quad \ldots (36)$$

$$E'_2 (\xi) = b \sin(\omega t - \theta_2 - \xi) \quad \ldots (37)$$

where

$$a \cos \theta_1 = E_1 + k E_2 \cos \delta$$

$$a \sin \theta_1 = k E_2 \sin \delta$$

$$b \cos \theta_2 = k E_1 + E_2 \cos \delta$$

$$b \sin \theta_2 = E_2 \sin \delta$$

On combining the two voltages of equations (36) and (37) we get the apparent intensity $I'(\xi)$ after taking the mean of the expression obtained for the intensity of radiation for a period longer than the time period of the
wave. Thus

\[ I'(\xi) = a^2 + b^2 + 2ab \cos (\theta_2 - \theta_1 + \xi) \]

\[ = (1 + k^2) I + 2kU + \left[ 2kI + (1 + k^2) U \right] \cos\xi - (1-k^2) V \sin\xi \]  ... (39)

After going through the procedure of introducing different phase delays we obtain the apparent values of the Stokes parameter denoted by primes as

\[ I' = I \]  ... (40)

\[ U' = 2kI + (1+k^2) U \]  ... (41)

\[ V' = (1-k^2) V \]  ... (42)

\[ Q' = (1-k^2) Q \]  ... (43)

These give

\[ I_e' = 4k^2 I^2 + 4k^2 U^2 + 4k (1+k^2) IU + (1-k^2)^2 I_e \]  ... (44)

or in terms of polarization degree

\[ m'^2 = 4k^2 + 4k^2 U^2 / I^2 + 4k (1+k^2) U / I + (1-k^2)^2 m \]  ... (45)

or

\[ m'^2 = 4k^2 + 4k^2 m^2 \cos^2 \beta \sin^2 \chi + 4mk (1+k^2) \cos 2\beta \]

\[ \sin 2\beta + (1-k^2)^2 m^2 \]  ... (46)

These equations can be interpreted by treating some special cases:

(a) For completely unpolarized wave \( m = 0 \), equation (46) reduces to \( m = 2k \)  ... (47)
In case of the antennas used in the radio polarimeter the measured value of crosstalk between the antennas was better than -30 db \((k=.03)\) and this gives

\[ m' = .06 \] \hspace{1cm} \text{...(48)}

Thus 6 per cent polarization will be read for an unpolarized wave, when the crosstalk is -30 db.

(b) For a completely polarized wave, that is, \(m = 1\), assuming \(k \ll 1\), equation (46) reduces to

\[ m - m = \Delta m = 2k(1-r^2) \sin^2 \chi \] \hspace{1cm} \text{...(49)}

where \(\Delta m = m' - m\).

For a circular wave and certain orientations

\[ \Delta m = 0. \]

For \(r \ll 1\), that is, a highly elliptically polarized radiation excluding certain orientations, \(\Delta m \ll .06\). Thus, the maximum error introduced in the measurement of the polarization degree and crosstalk is less than 6 per cent for a polarized radiation and 6 per cent for an unpolarized radiation.

From equations (42) and (44) we obtain the expression for the apparent value of the axial ratio, \(r'\) as

\[ \sin^2 \beta' = \left[ \frac{m(1-k^2) \sin^2 \beta}{4k^2 + 4k^2 \cos^2 \beta \sin^2 \chi + 4k(1+k^2) \cos^2 \beta \sin^2 \chi + (1-k^2)m^2} \right]^{1/2} \] \hspace{1cm} \text{...(50)}
Again, some special cases are treated.

a) For \( m = 0 \), that is, for a completely unpolarized wave, equation (50) reduces to \( r' = 0 \).

Therefore a completely unpolarized wave is read as 6 per cent linearly polarized radiation (Instrumental polarization). The apparent orientation angle dependent on the intensity of the incident radiation is \( \tan 2 \chi = 2kI \).

b) For \( m \sim 1 \), that is, a a completely polarized radiation, assuming \( k \ll 1 \), equation (50) reduces to

\[
\sin^2 \beta' = \sin^2 \beta / \left[ 1 + 4k \cos^2 \beta \sin^2 \chi \right]^{1/2}
\]

or

\[
\delta r = \left| 2kr \sin \chi \right|
\]

where \( \delta r = r' - r \).

The error introduced in the measurement of the axial ratio depends upon orientations. For a nearly circular wave \( \delta r \sim 0.06 \), and a for highly elliptical polarization, \( \delta r \sim 0.001 \), and \( \delta r \) becomes zero for a linearly polarized radiation. So the antenna errors, either due to nonzero crosstalk or due to the relative phase difference between the two orthogonal components, may not exceed 6 per cent in the measurement of polarization degree and 0.06 in the axial ratio.
3.53 REFLECTIONS FROM LOCAL OBJECTS, ESPECIALLY THE GROUND

Errors due to reflections from local objects, especially the ground, can be quite serious with the antenna systems having a broad beam-width. Thus, this needs proper assessment in order to understand and correct, if necessary, for an undesirable effect due to ground reflection.

The ground can be treated as a semiconductor having a dielectric constant as well as conductivity. The electric properties of the ground depend upon the nature of the terrain and particularly its humidity and slightly on the frequency. The moisture content in the soil drastically changes the electrical properties of the ground. For instance, for wavelengths $>3$ m, the dielectric constant changes from a value of 30 (with respect to vacuum) for a humid soil to a value of 4 when the soil is very dry. Similarly, the soil conductivity may change by a factor of 10. When the soil is very humid, it becomes a very good conductor electrically ($\approx 0.01$ mho/m, David and Voge 1969). This implies that the error due to ground reflections has a seasonal dependence. For longer wavelengths, the conduction current is predominant i.e. the ground can be treated as a good conductor.

For a perfectly plane conducting reflecting surface, the boundary conditions which decide the
characteristics of the reflected wave are simple, that is, the vertical polarization is reflected without any phase change, while the horizontal polarization is totally reflected but with a $180^\circ$ phase change. In general, the phase change is dependent upon the dielectric constant and electrical conductivity. Thus the effective field with its characteristic polarization as seen by the antennas is the sum of the incident field and the reflected field. For specular reflection from the ground, there is always a phase path difference between the directly received incident radiation and the reflected radiation. This phase path difference is a function of the effective antenna height from the ground and the elevation angle of the source. The phenomenon becomes much more complicated if the reflecting surface exhibits irregularities. The specular reflection will disappear if the dimensions of the irregularities are large compared with the wavelength, giving rise to a series of elementary reflections, more or less random in all directions; in other words, to scattering. In addition, obstacles can prove to be nuisance as they may produce undesirable effects like phase changes, attenuation and diffraction.

From the above description it is amply clear that it is very difficult to calculate the exact contribution from the ground. One can do some theoretical iterative
computations and thus make some estimate of the errors (Dodge 1972). But then such estimates very rarely correspond to the real situation. What one can do is to minimize such errors by being very choosy about the site where antennas are to be installed and keep the back and side lobe responses facing the ground at the minimum possible level. We have taken due care in the selection of the site. Further, we restricted our observation period to 2-3 hours before and after the local noon so that the undesirable effects, due to nearby objects and large beam-width of the antennas, are minimized. Also the elevation angle of the sun can reach a minimum of 54° at the meridian transit at our station. This minimizes the errors due to a low elevation angle of the source. The measured front-to-back ratio of the antenna gain is \( \sqrt{20} \) dB. This implies that the reflected wave will be attenuated by \( \sqrt{20} \) dB when the antenna is looking at the zenith.

We estimated a limit on the error due to ground reflections by observing the flux from galactic background. Since the antenna beam-width is about 40° x 40°, the flux coming from the galactic background, which is an extended source, can be assumed to be randomly polarized. This extended source due to the galactic background encounters varying reflecting conditions at the ground. For recording the galactic background the system is to be operated at a
higher gain. It is seen that the error in the polarization degree and the axial ratio does not exceed 10%. The \( \chi \) values were found to be randomly distributed. But it should be noted that the \( \chi \) values of the radiation coming from a narrow source on the sun may not be randomly distributed by the ground reflection because the solar radiation from sources of angular dimension 10' arc (core) at 35 MHz effectively reduces the antenna beam-width to the dimension of the radio sources as viewed from the earth (Dodge 1972). This implies that an antenna will receive the reflected radiation from a smaller reflection area and thereby experience a more uniform ground reflection. Thus, this may sometimes give rise to a clustering of the orientation angles (Chapters IV and V). However, we do not feel that this effect is operative because our two-frequency observations reveal that there is at least one large group of type III bursts which showed far lesser fluctuations in the orientation angles at one frequency than in the other and the ground effects, if operative, are essentially broad-band and thus would not selectively fluctuate position angles only at one frequency.

Further, we made an estimate of the errors due to ground reflection by radiating a linearly polarized radiation at a height of \( \sim 300 \) ft. (by floating a tethered balloon) above the boom of the antenna system when the
antenna system was looking at the zenith. We could measure the error in the degree of polarization and axial ratio caused by the coupling due to reflection from the surrounding objects and the ground. It did not exceed the 10 per cent limit fixed by calculating the error in measurements using galactic background. Thus, we feel that the errors due to ground reflection can be serious but in our case it turns out that the total error is less than 10 per cent and, therefore, should not affect the conclusions reached in Chapters IV and V.

3.54 STATISTICAL ERRORS

Because of the depolarization considerations the I.F. bandwidth $B$ is kept small. The integration time $\tau$ does not exceed 1 S. The receiver output fluctuates because both the burst radiation and the receiver noise have statistical nature. The relative fluctuations in power are $\sqrt{(\beta \gamma)^{-1}}$. With $\beta = 7.5$ KHz and $\gamma = 1$ S, the relative fluctuations in power does not exceed 1 per cent. The error of this kind increases only if any attempt is made to measure the polarization parameters of the radio bursts with intensity smaller than receiver noise (Cohen 1958b).

We did not make any attempt to analyze the bursts which have a low flux. So we consider these errors to be very small.
While scaling the data the errors can creep in, particularly in case of those bursts which have produced amplitudes less than 5 mm on the intensity channel. We have scaled our data at the peaks of the bursts and we have not analyzed those bursts which had caused amplitudes less than 5 mm.