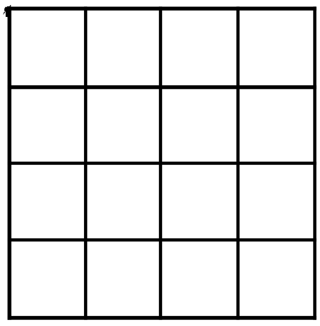
CHAPTER 2

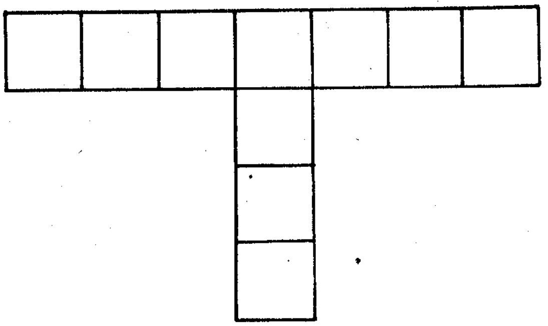
DESIGN CRITERIA

2 INTERFEROMETRY AND SKELETON TELESCOPES

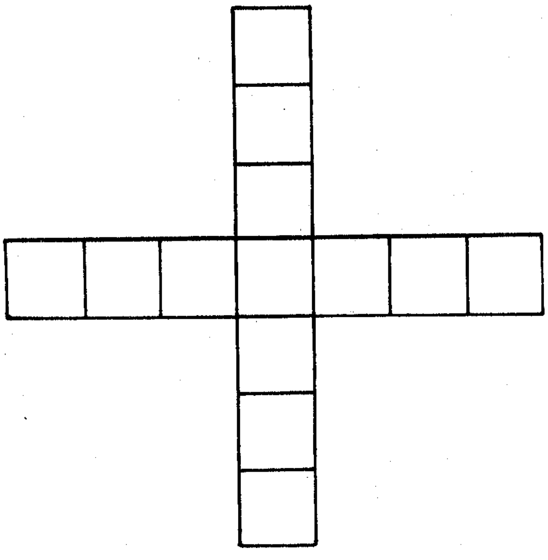
The resolution of a telescope has a diffraction limit λ/d, where λ is the observed wavelength and d is the .aperture diameter. One of the major problems of decametric radio-telescopes is, that even to obtain a resolution of the order of 1°, one needs an aperture d 573 mts. The sensitivity of continuum observation at decametric wavelengths is more often limited by confusion than by receiver noise. Thus it is essential to have higher resolution to achieve better sensitivity. It is of little use to increase the sensitivity beyond the point where more than one source is likely to be within the antenna beam. The resolution obtainable with a given number of receiving elements depends upon the placement and the way in 'which they are connected. The speed at which a large aperture can be synthesized depends upon the arrangement, the number and the mobility of the elements (Fomalont and Wright, 1974). consider a square aperture of figure 2.1. At the receiver the contribution from each element will produce separately



FILLEO APERTURE



T ANTENNA



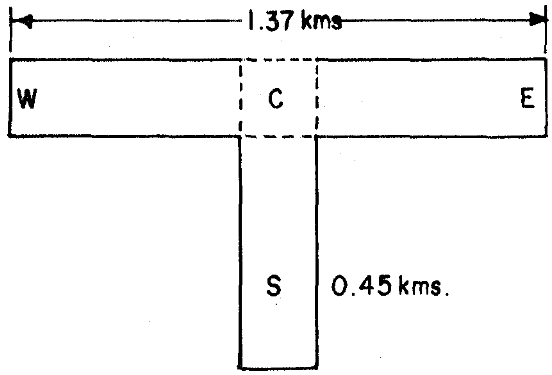
CROSS ANTENNA

FIG. 2.1 FILLED APERTURE, **T ANTENNA** ***8*** CROSS **ANTENNA.**

an interference pattern with each of the other N-1 elements. 'Thus the total number of cross products are N(N-1)/2. In these cross products many occur with redundancy. Highest redundancy occurs at closest separations. The "Tee" array is a skeleton array which contains all the relative positions of a filled aperture. For example, only 10 receiving elements will be required to span a filled rectangular aperture containing 16 elements. Generally, if we divide a square aperture into nxn elemental areas the corresponding "Tee" array contains only 3n - 2 elements. Another possible arrangement of these elements is in the form of a cross. For the same resolution a cross needs more number of receiving elements than a Tee. But a cross can be combined to form a correlation telescope which is less susceptible to errors in the phase adjustment. A cross also has a better sensitivity than a Tee (if sensitivity is not confusion limited).

At Gauribidanur, 1000 dipoles are used to form the T' array. The basic dipole is shown in Figure 2.2. Their characteristics and arrangement are also shown in the Figure.

In the study of radio sources whose emission is not time dependent, it is not necessary that the entire telescope aperture be present at the same time (Ryle and Hewish, 1960). With just two elementary areas, one of them being movable one can sequentially measure all the cross products. The fourier transform of these cross products can



T ANTENNA AT GAURIBIDANUR

1. LOCATION — LATITUDE, 13° 36' 12"N.
LONGITUDE _ 77° 26' 07"E.
INSTRUMENTAL ZENITH _ 14° 6'N.

2. ARRANGEMENT — ALL DIPOLES ARE ORIENTED ALONG E-W DIRECTION.

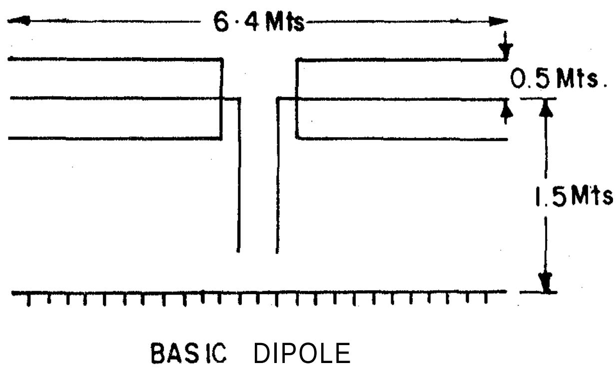
3. E-W ARRAY — 4 ROWS OF 160 DIPOLES EACH.

4. SPACING — 8.6 mtrs IN E-W DIRECTION.
5 mts IN N-S DIRECTION.

5. N-SARRAY — 90 ROWS OF 4 DIPOLES EACH.

BASIC DIPOLE CHARACTERISTICS

IMPEDANCE — 600 Ω. BANDWIDTH - 10 MHz WIDE AROUND 32 MHz.
V.S.W.R - 1.1 TO 1.5. POLARISATION - LINEAR IN E-W DIRECTION



BASIC DIPOLE

FIG. 2·2 BASIC DIPOLE, T ANTENNA AT GAURIBIDANUR.

be performed later. This is the principle of aperture synthesis. In spite of the simplicity and economy of the above antenna system, it is not used in Gauribidanur since it has a few disadvantages, which are listed below.

1. Surveying sensitivity is reduced.

2. It is difficult to re-examine any small portion of the sky.

3. To get a map of the sky one has to wait until all the observations are complete.

4. Aperture synthesis telescopes are useful only to study radio sources whose emission is not time dependent. Thus studies of solar bursts, scintillations and pulsars are not possible with telescopes where the aperture is synthesized using a small number of elements.

5. Ionospheric refraction and scintillation are more pronounced at decametric wavelengths. This makes the maps made using cross products obtained on differfent days less reliable.

2.2 IMAGE FORMATION

A directional antenna can be regarded as an energy rejecting device. If the voltages from dipoles are simply added, the energy received from a large part of the sky will be rejected. This is the response of the antenna. The antenna field pattern is the fourier transform of the grading. On the other hand, if the contributions from dipoles are amplified separately and combined in different ways, one can generate multiple independent beams simultaneously. In principle, an 'N' element array can generate N independent beams. Multiple beams allow parallel operation and a higher data rate. A single-beam. instrument can use only part of the total available time to observe each beamwidth in the sky, while a multiple-beam or image forming instrument of the same effective area will not divide its time between measurements in different directions.

2.2.1 Image Forming Techniques

A simple linear array which generates a single beam can be converted to a multiple beam antenna by attaching phase shifters to the output of each element. Each beam to be formed requires one additional phase shifter per element. This is known as a Blass network.

At Gauribidanur there are 1000 dipoles. To form multiple beams using them, one requires a very large number of phase shifters, which is impractical. For the Gauribidanur telescope, the sensitivity is limited by confusion. Thus it is not necessary to form multiple beams in the East-West direction to obtain increased observation time. For continuum observations, the East-West array is used in the transit mode, and the earth's rotation is used for covering the whole sky. Observation of pulsars and scintillation are not confusion limited. For such studies, individual phase shifters for groups of dipoles can be used to provide tracking facility to increase the observation time. In the existing system, the East-West array is divided into ten major groups, 5 in the Eastern arm and 5 in the Western arm of the array (Fig.2.3). Each major group contains 4 minor groups of dipoles between which phase shifters are provided for tracking. Each minor group contains four elementary dipoles in the E-W direction which are combined without any phase shifters. This re'stricts the tracking angle to about 15° of the sky.

Since the earth's rotation is from West to East, a similar arrangement will not suffice in the N-S direction. There are 360 dipoles in the Southern arm. These are arranged in 4 E-W rows each containing 90 dipoles (Fig. 2.4 A,B). Since multiple beams are required in the N-S direction, it is appropriate to combine the dipoles in the E-W direction and to use 90 groups in the Southern arm. The primary beam of each group has an extent of 15° in the E-W direction and covers $\pm 60^{\circ}$ in the N-S direction. With 90 groups in the Southern arm, 90 independent beams can be formed. To form 90 independent beams with the phase shifters, one requires 90 X 90 = 8100 phase shifters which is economically a large requirement. The efficiency of beam forming can be improved by using a "Butler Matrix" (Butler, 1961). This is a lossless network which utilises 3-dB directional couplers along with fixed phase shifters. It

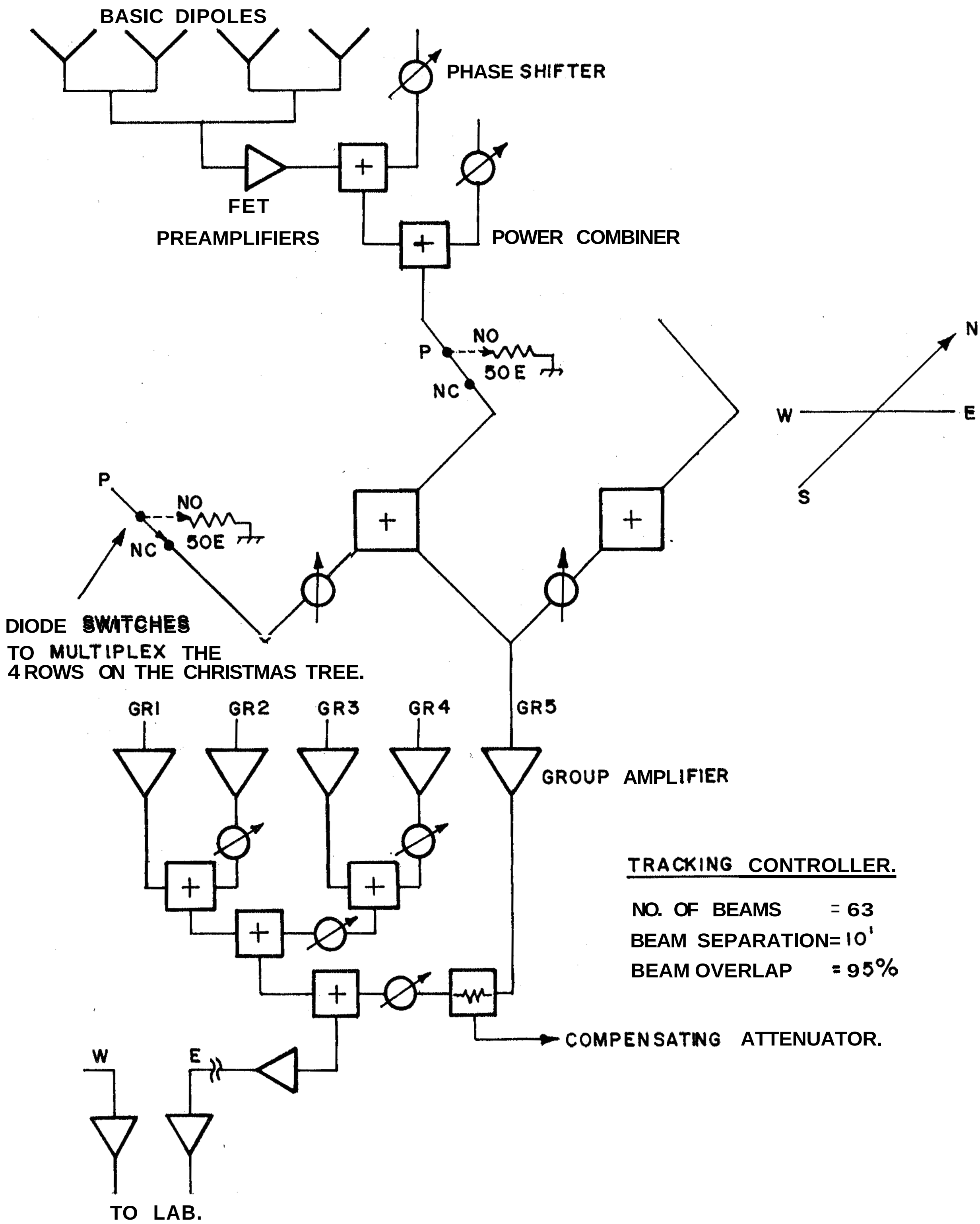


FIG. 2·3 SCHEMATIC DIAGRAM OF THE E-W ARRAY.

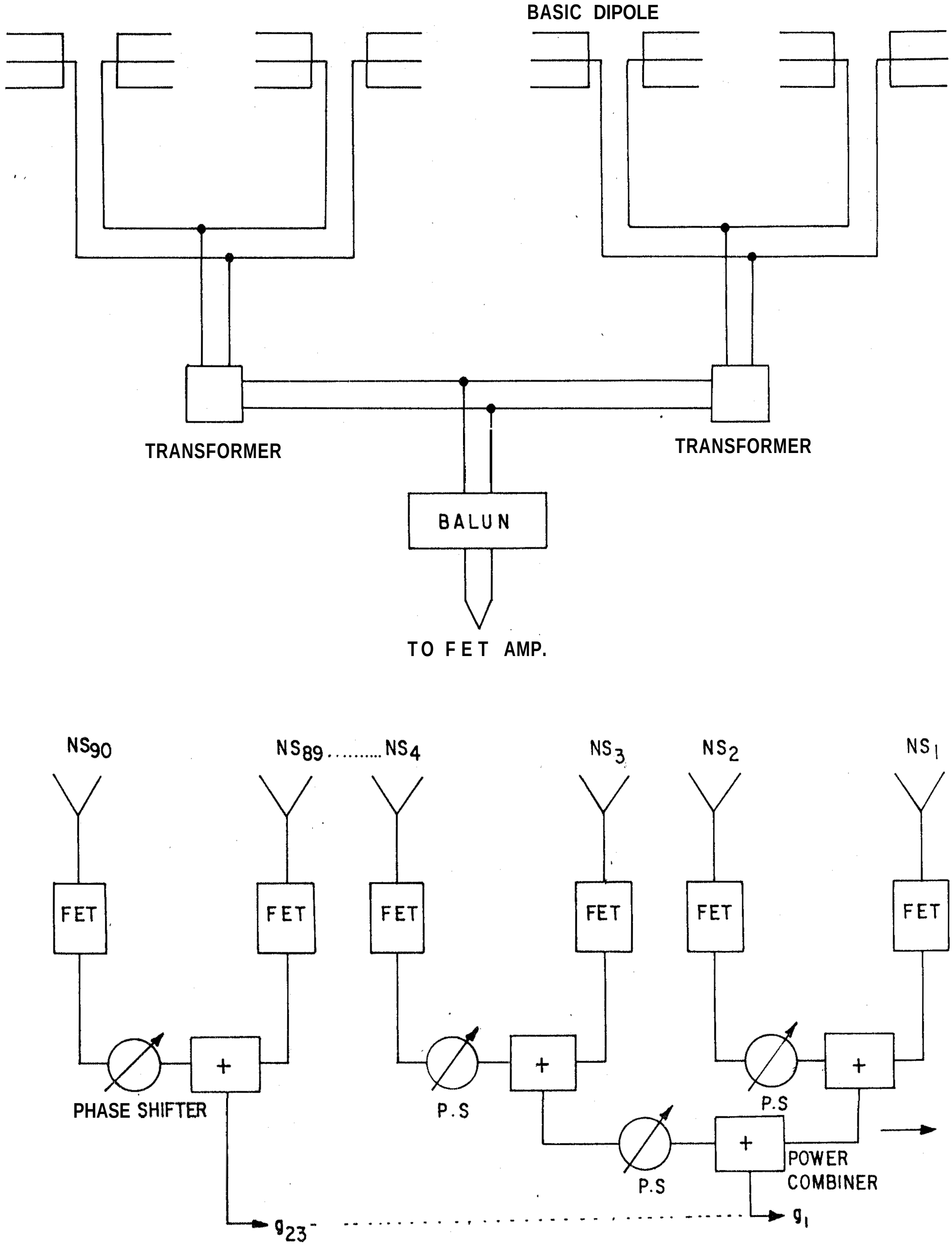
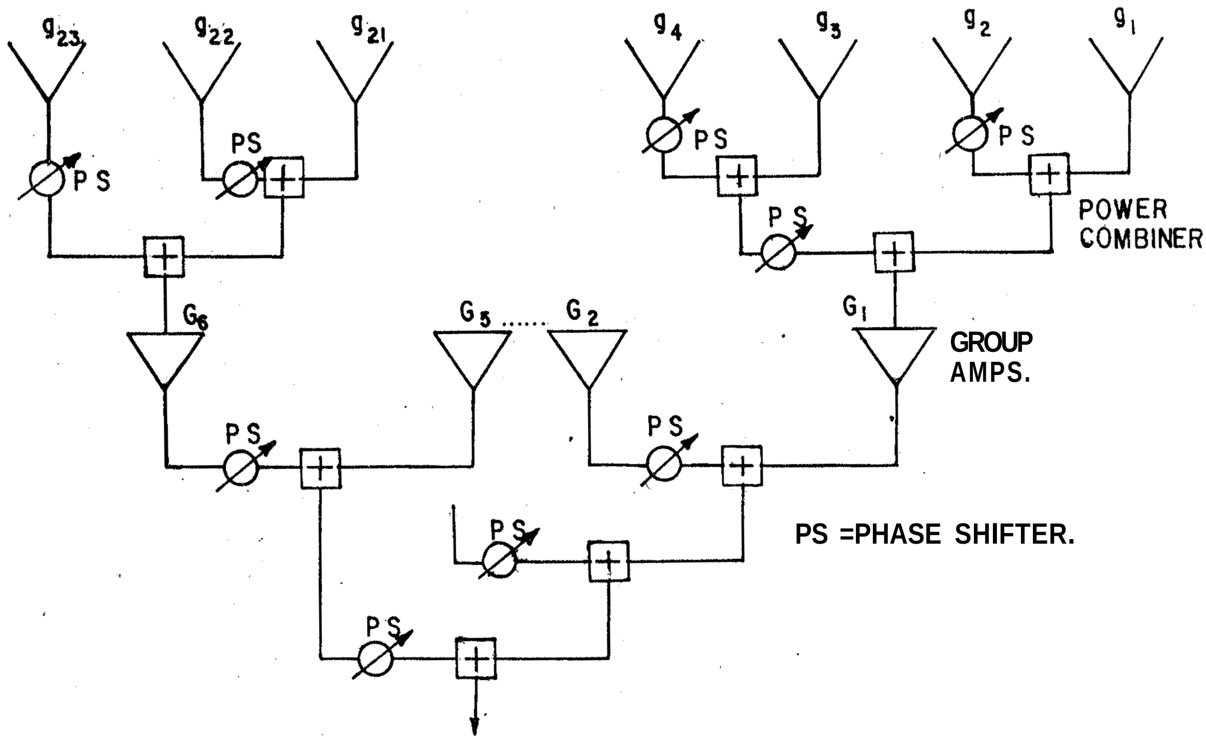


FIG. 2·4A SCHEMATIC DIAGRAM OF THE N-S ARRAY.



N.S SCANNING

MAXIMUM NO. OF DECLINATIONS THAT CAN BE SCANNED = 16

MINIMUM SELECTABLE STEP – $0.^{\circ}2$.

CYCLE TIME = 0 . 8 Secs.

MAXIMUM INTERCHANNEL LEAKAGE = 2.5%

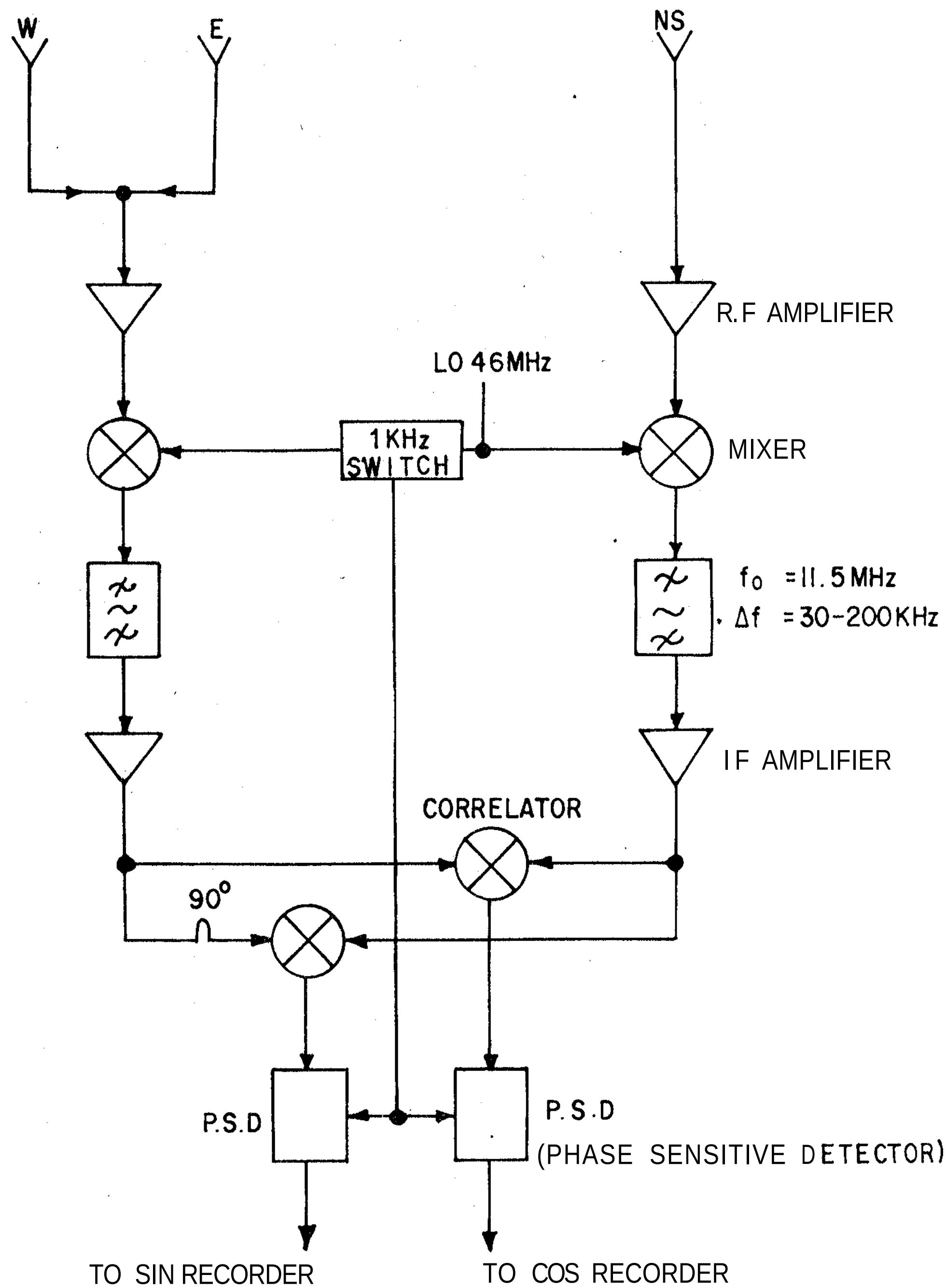
MAXIMUM SKY COVERAGE/SCAN = $6^{\circ}.4$.

FIG. 2·4B SCHEMATIC DIAGRAM OF THE N-S ARRAY.

can form N contiguous beams from an N element array. A Butler matrix for an N element array needs N/2 LogN/2 hybrids and N/2 (LogN/2 - 1) fixed phase shifters. It is of interest to note the relation of a Butler network to Fast Fourier Transform (FFT). A Blass network required N^2 couplers for N inputs and N outputs while Discrete Fourier Transform (DFT) also requires N^2 multiplications for an N point transform. The Butler network utilises N/2 LogN/2 hybrids just as FFT uses N/2 LogN/2 multiplications for an N point transform. The complexity of the Butler network increases with the number of elements. For example, a 64 element array requires 192 directional couplers and 160 fixed phase shifters. This leads to constructional difficulties and besides, the matrix will not be lossless. This makes the Butler Matrix not suitable for large arrays. All systems with only phase shifters suffer from the effect of "Band width decorrelation" unless one uses "Delay Shifters" in each arm. In a system using a Butler matrix it is difficult if not impossible to introduce differential delays to eliminate bandwidth decorrelation. This arises an account of the signal flow (similar to FFT) in the matrix. Further for post processing (like removing the point source response) it is preferable to oversample the sky brightness distribution. This results in increasing the hardware required for multibeam forming using passive networks.

The effect of multiple beams can also be obtained by rapidly steering a single beam (time division multiplexing). The existing receiver at Gauribidanur uses this technique. The response of the beam in the N-S direction can be switched in steps of $0\overset{\circ}{.}2$ and 16 beams can be formed in the time-multiplexed mode (Fig. 2.4B). Switching is done at the rate of 0.8 sec/cycle. The beams so formed are correlated with the E-W arm (Fig.2.5). The width of the cosine pencil beam is 26′E-W/40′ sec ZA N-S. This system has a receiver noise $\sqrt{N}$ times higher than the one in which N beams are formed by N independent elements. It also has a limited viewing angle of $6\overset{\circ}{.}4$. A few HII regions and the Sun have been mapped using this system (Sastry, 1981; Dwarakanatb, 1982; Deshpande, 1984).

Multiple beams can also be .formed by obtaining the interference patterns (both Sine and Cosine) from pairs of component antennas (Christiansen, 1966). To obtain the interference patterns the outputs of the component antennas are amplified separately and are multiplied in Cosine and Sine correlators. The output of each pair measures a fourier component of the brightness distribution of the sky that is seen by it. (sec. 1.3). Multiple beams can then be formed off-line by fourier synthesis of the measured correlation values. This method of image formation has a number of advantages.



ANTENNA PARAMETERS

RESOLUTION E-W ARRAY - 21'/30° RA/ DEC.

N-S ARRAY - 25°/69' RA/ DEC.

AFTER CORRELATION RESOLUTION
OF COSINE PENCIL BEAM=26'/40' SEC. Z A

THEORETICAL BEAM SHAPE — SINC FUNCTION

MINIMUM DETECTABLE FLUX FOR 10 Sec INTEGRATION TIME 8 200KHz BANDWIDTH } ≈ 30 JANSKY

FIG. 2·5 SINGLE BEAM-FORMING RECEIVER

a) Data in the visibility domain can be easily manipulated to obtain maps with different gradings, to have finer sampling of the radio brightness and to excise interference.

b) Instrumental phase information can be obtained using strong point sources in the sky

c) Delays between the signals can be easily compensated, thus reducing the bandwidth decorrelation.

2.2.2 Arrangement Of The Gauribidanur 'T' Array For One Dimensional Synthesis

The aim of the present work is to obtain multiple beams in the **N-S** direction only. This requires the measurement of the correlations between pairs of component antennas having different **N-S** spacings:

As mentioned earlier there are 4 rows in the E-W array and 90 rows in the **N-S** array. One can multiply each of the four rows of the E-W array with the 90 rows of the N-S array to obtain the required fourier components for synthesis. Such a system requires 94 SHR (superheterodyne receivers) and 360 correlators. With such a system the entire visible sky can be synthesised in one day (ignoring the effects of bandwidth decorrelation).This scheme uses the entire collecting area of the array and has the maximum surveying sensitivity. But it has a few disadvantages.

a) Even though 360 products are obtained, only 90 of them are independent fourier components.

b) In the existing system, only one output is available from the E-W array. Thus to obtain four outputs from the E-W array the Christmas tree has to be reconfigured,

Two ways in which one can avoid any redundant North-South spacings are: 1. To multiply only one row of the E-W array with each of the 90 rows of the N-S array. 2. To multiply each of the four rows of the E-W array with every fourth row of the N-S array. Both configurations need rearrangement of the E-W christmas tree, have the same collecting area and require the same number of correlators. Each has a sensitivity only 1/2 of the maximum system sensitivity attainable by using the entire array. This may not be a serious disadvantage when confusion and not the system temperature limits the sensitivity. The 2nd configuration needs only 27 front ends compared to 94 front ends that configuration 1 requires. It is also possible to use the second configuration without completely reconfiguring the E-W array. Instead of obtaining the outputs of the four rows independently, one can obtain the outputs on the existing christmas tree in a serial mode. This will require additional diode switches which are relatively inexpensive compared to the cost of the cable required to reconfigure a 1.4 km long array. Since time division multiplexing of the four rows is involved, this system will take more time (a factor of 4 compared to the other configurations) to measure all the required fourier components, and further reduce the sensitivity.

The objective of the present work was to test as soon as possible the multibeam forming receiver and to obtain radio maps to assess its suitability for a large scale survey. For this, the N-S array was divided into 23 yroups, and the entire E-W array was multiplied with these groups to obtain the fourier components. Such a scheme needed no reconfiguration of the existing N-S and E-W arrays and is shown in Fig. 2.6. The primary beam of each element of the interferometer was 15° at the zenith, and thus restricted the viewing angle (section 4.2). To map different regions of the sky, the direction of the primary beam response of the interferometers was changed using the phase shifters. Chapter 4 discusses the observations and the limitations of the maps made using such a configuration.

2.3 RECEIVER SYSTEMS AND DELAY COMPENSATION

The difference. in the time (τ) of arrival of a wavefront at the E-W and different N-S groups increases with the spacing and zenith angle. When τ becomes comparable to the inverse of the receiver bandwidth, decorrelation becomes large. To overcome this, one introduces artificially a delay τ_D in the signal path from one of the antennas to the receiver. Figs. 2.7 and 2.8 describe a Single Sideband (SSB) and a Double Sideband (DSB) receiver. In an SSB system, the local oscillator frequency is such that image

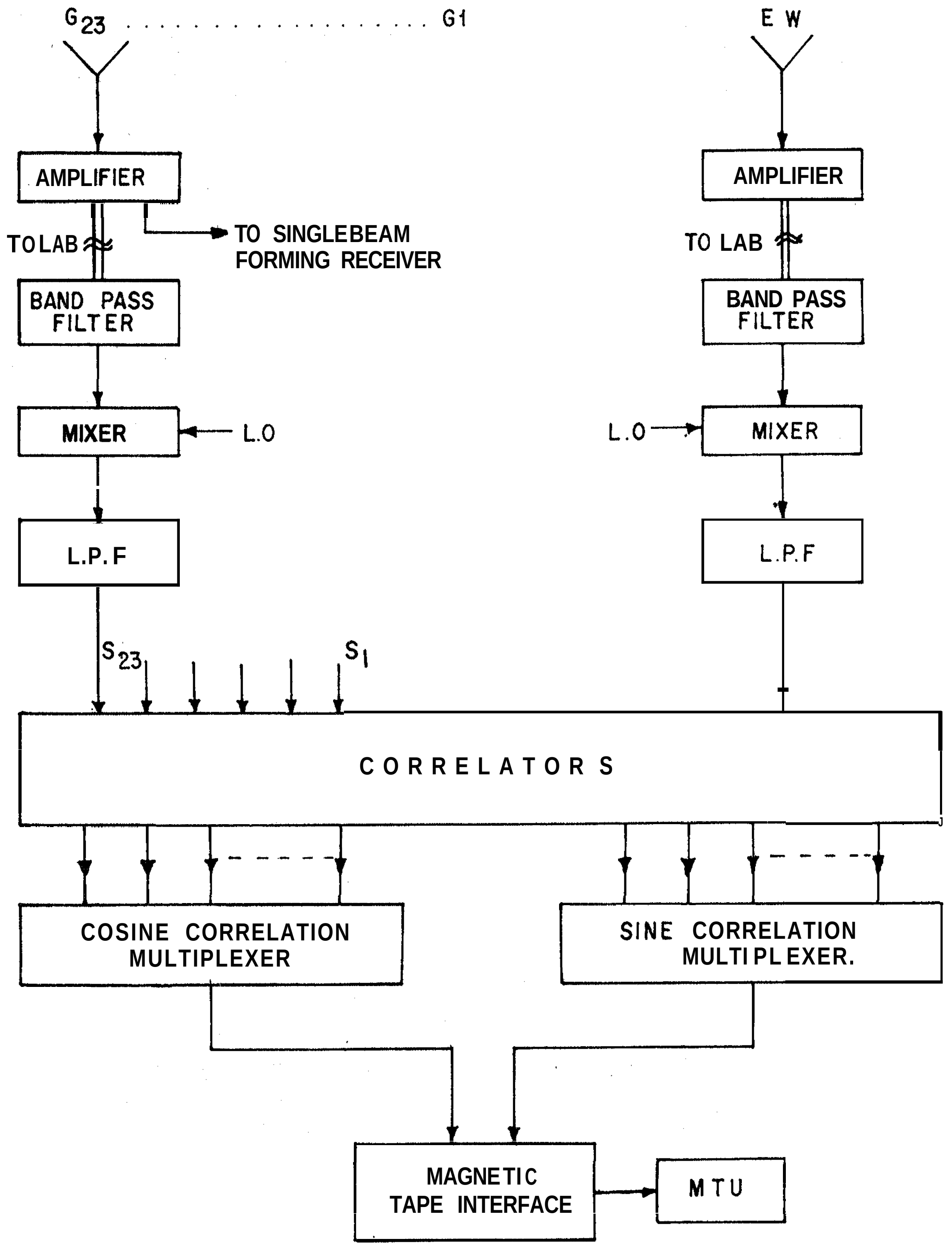


FIG. 2·6 ANTENNA AND RECEIVER CONFIGURATION FOR THE DIGITAL MULTIBEAM FORMING RECEIVER.

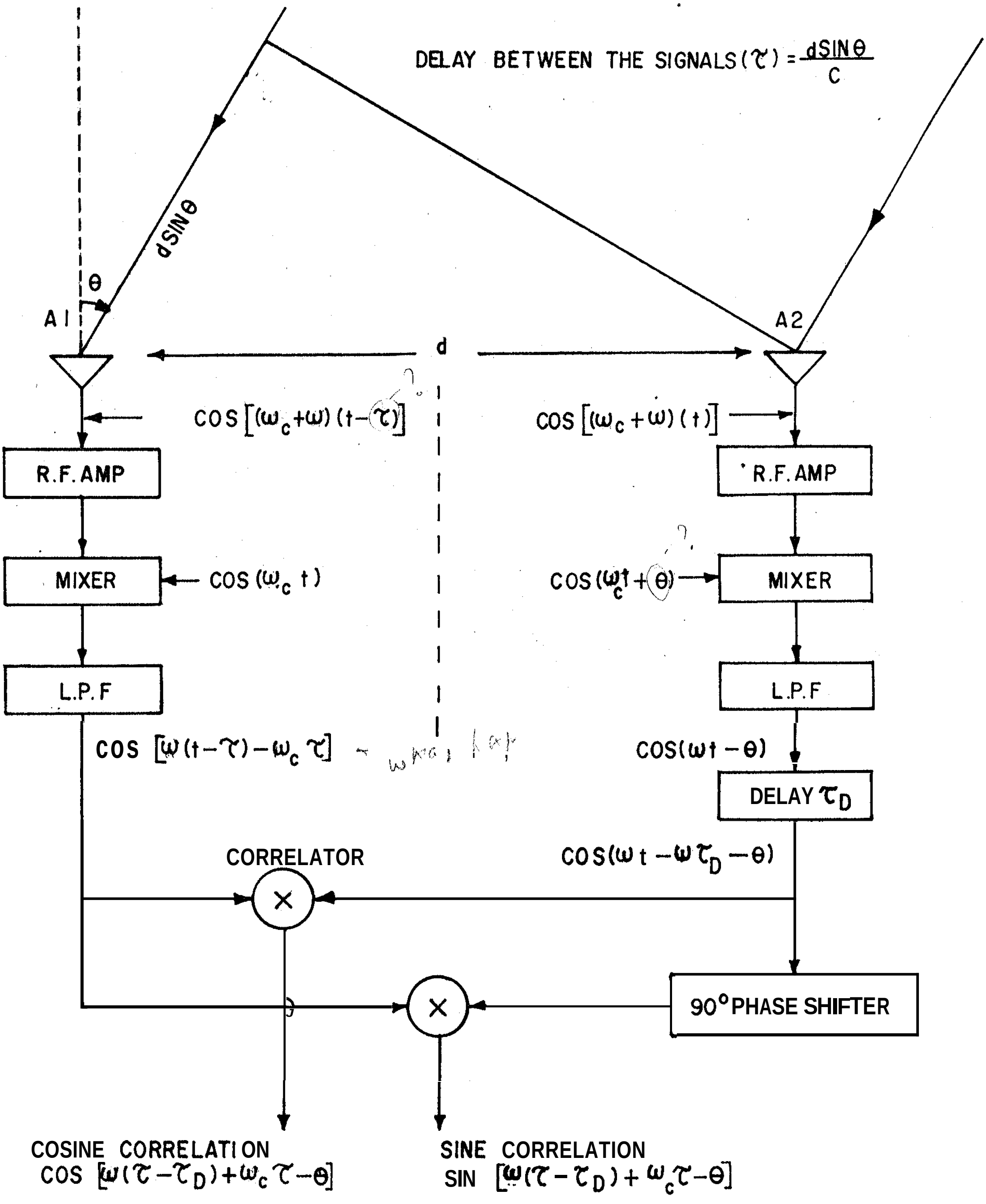


FIG. 2.7 CONVENTIONAL SINGLE SIDE-BAND RECEIVER.

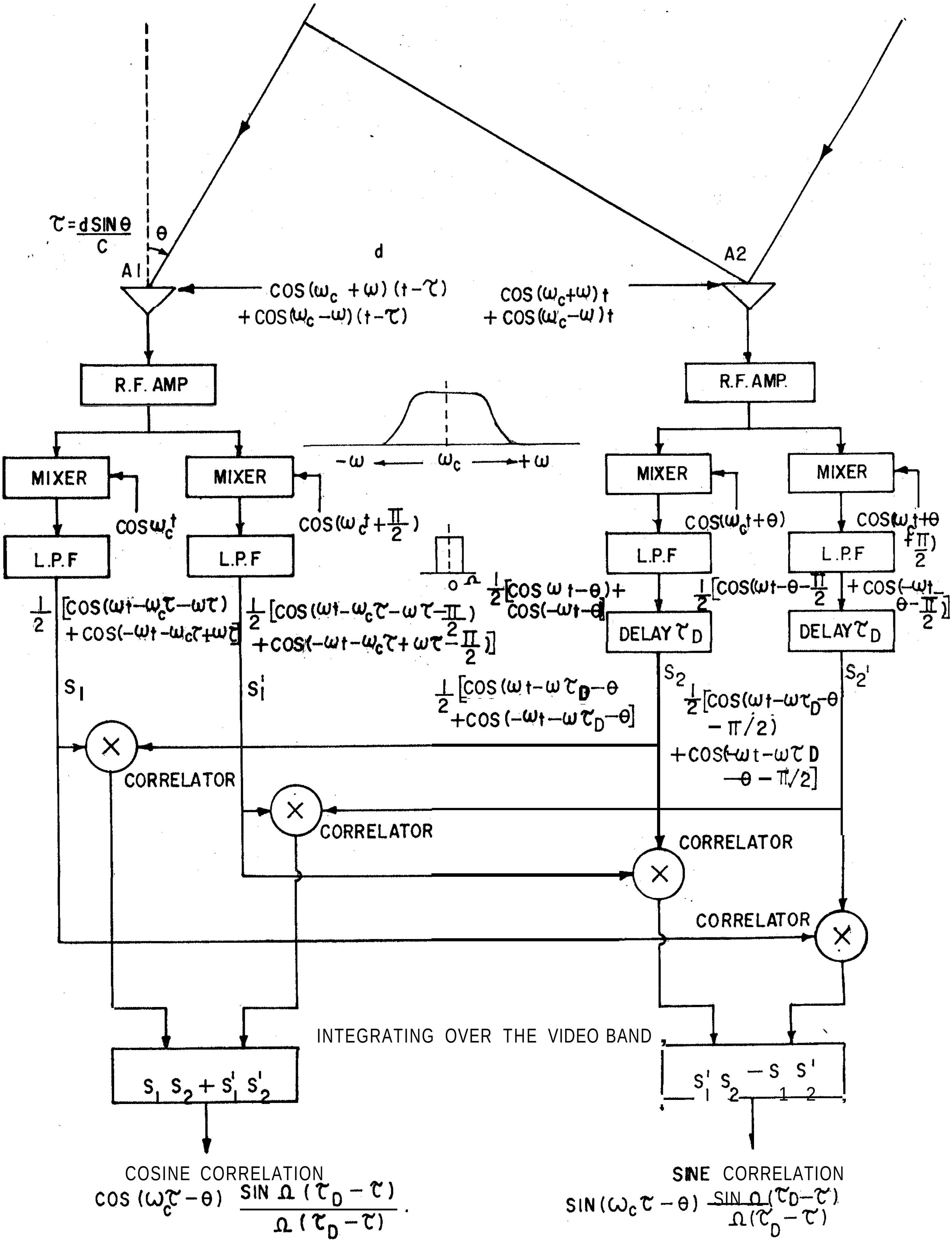


FIG. 2·8 CONVENTIONAL DOUBLE SIDE-BAND RECEIVER.

frequencies are rejected while heterodyning. This needs sharp cut-off filters. In a DSB system, image frequencies are not rejected. Here the LO will be within the RF band. A DSB operation has certain advantages over SSB in receivers. The following conclusions can be drawn from the equations shown in figs. 2.7 and 2.8.

2.3.1 <u>SSB System</u> (Fomalont And Wright, 1974)

1. For a given direction, correlation can be maximised only for a single value of ω by adjusting either ϕ or τ_D. In order to maximise correlation for any value of both ϕ and τ_D must be adjusted such that $\tau_D = \tau$ and $\phi = \omega_c \tau$. Hence to maximise the signal two adjustments are required.

2. Fine and continuous delays are difficult to achieve. Simple hardware can give only coarse steps in τ_D. Uncompensated delay $\Delta\tau = \tau - \tau_D$ should be much smaller than the reciprocal bandwidth.

3. Coarse steps in τ_D introduce phase jumps equal to $\omega_{IF}\tau_D$. Thus τ_D must be instepsof $2\pi/\omega_{IF}$ to ensure that all phase jumps are 2π at the centre of the IF band. However significant losses occur unless bandwidth is $\ll \omega$ IF.

4. Alternatively, τ_D can be changed in very small steps resulting in small phase jumps. These can be compensated by a phase adding device in the IF line.

2.3.2 DSB System (Read, 1963 Radhakrishnan, 1972, Fomalont 1974)

1. A DSB interferometer acts like a single frequency interferometer. This frequency is the LO frequency. This helps to fix the interferometer baseline in a wideband' interferometer without knowing the centre of gravity of the amplifier's bandshape.

2. The response of a DSB system is doubly periodic. One frequency is proportional to ω_c and other to' ω . The phase of the rapidly varying component is no longer dependent on τ_D.

3. τ need not be compensated accurately. Coarse steps of τ_D are enough to keep the response near the maximum.

4. Phase jumps due to τ_D for the two sidebands are in the opposite directions. This ensures the phase calibration to be independent of τ_D .

Because of the above advantages, a DSB system was chosen for the present receiver.

In order to achieve phase calibration independent of the delay settings, the DSB system shown in Fig. 2.8 uses four, correlators. The phase part could have been obtained with only two correlators. This is true because $\overline{S_1 S_2}$ will be numerically equal to $\overline{S_1' S_2'}$ since the LO phase cannot change the correlation coefficient between the two signals .

One could have determined the cosine correlation by using only one correlator. The same is true of the sine correlation as $\overline{S'_1 S_2} = -\overline{S_1 S'_2}$. But the use of only two correlators to determine the correlation coefficients reduces signal to noise ratio (SNR). Eventhough $\overline{S_1 S_2} = \overline{S'_1 S'_2}$ the noise on them are not the same. S_1 (S_2) and S'_1 (S'_2) are samples of the same signal which are independent. Hence $\overline{S_1 S_2}$ and $\overline{S'_1 S'_2}$ will be two independent determinations of the same correlation coefficient. Thus their addition improves the SNR ($\overline{S_1 S_2} = \overline{S'_1 S'_2}$ and $\overline{S_1 S'_2} = \overline{S'_1 S_2}$ is efficiently made use of in the present receiver system). The video bandwidth of S_1 and S'_1 is only half of the RF input bandwidth. This means that one needs to sample S_1 and S'_1 at half the rate required in a SSB system. A DSB system requires sampling of both S_1 and S'_1 to get an SNR equal to that of a SSB system. Thus the number of independent samples in the two systems a're equal.

2.4 CHOICE OF CORRELATORS

For one dimensional synthesis mapping at Gauribidanur, the cross correlation of signals from E-W array and N-S array are determined. The cross correlation function of any two real signals $V_E(t)$ and $V_S(t)$ is

$$\rho(0) = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} V_E(t)\, V_S(t)\, dt \qquad (2.1)$$

But the above limiting operation cannot be performed in practice. An estimate of $\rho(0)$ can be obtained by a correlator that computes

$$\rho_M(0) = \frac{1}{T}\int_0^T V_E(t)\, V_S(t)\, dt \qquad (2.2)$$

In modern telescopes correlators are implemented digitally for the following reasons.

1. Digital operations are precisely defined and are repeatable with 'zero'error.

2. Due to availability.of **LSI** chips, the physical size of correlators will be small and multichannel systems can be built at low cost.

3. Component tolerances are not critical. Component drift and spurious environmental signals within limits have no influence on the system.

4. For the bandwidths presently used digital circuits can perform multiplication and integration.

5. Delays required to reduce bandwidth decorrelation can be obtained by simple shift registers.

6. Integration time can be easily programmed by using variable length counters.

Sampling and quantization are required for a digital correlator. The sampling frequency must be at least twice the bandwidth (Nyquist sampling rate). Such samples taken at an interval $\Delta t = 1/2\Delta\nu$ where $\Delta\nu$ is the bandwidth fully describe a band-limited signal after sampling. When a signal of bandwidth $\Delta\nu$ is quantised before sampling, the quantised signal will have a bandwidth larger than $\Delta\nu$ depending on the number of bits used. If the signal is then sampled at a rate $2\Delta\nu$ there will be a corresponding loss of information. The Gauribidanur receiver uses one-bit quantisation. The one bit method of finding cross correlation of two signals is shown in Fig. 2.9. This method uses a theorem derived by Van Vleck in 1943 (also by Sheppard in 1898) . It can be stated as follows.

If X(t) is a sample function of a Gaussian random process with zero mean and Y(t) is the function formed by infinite clipping of X(t). Then the normalised autocorrelation'functions of X(t) and Y(t) are related by

$$\rho_x = \sin\left[\frac{\pi}{2}\rho_y\right] \qquad (2.3)$$

This theorem also holds good for cross correlation of two signals which are jointly Gaussian random processes. For sampled waveform eq 2.2 reduces to

$$\rho_M(0) = \frac{1}{K}\sum_{n=0}^{K-1} V_E(n\Delta t)\, V_S(n\Delta t) \qquad (2.4)$$

Where Δt is the sampling interval. If V_{SC} and V_{EC} are obtained by infinitely clipping V_S and V_E , one-bit correlator obtains normalised correlation coefficient by performing the operation

$$\rho_C(0) = \frac{1}{K}\cdot\sum^{K-1} V_{EC}(n\Delta t)\, V_{SC}(n\Delta t) \qquad (2.5)$$

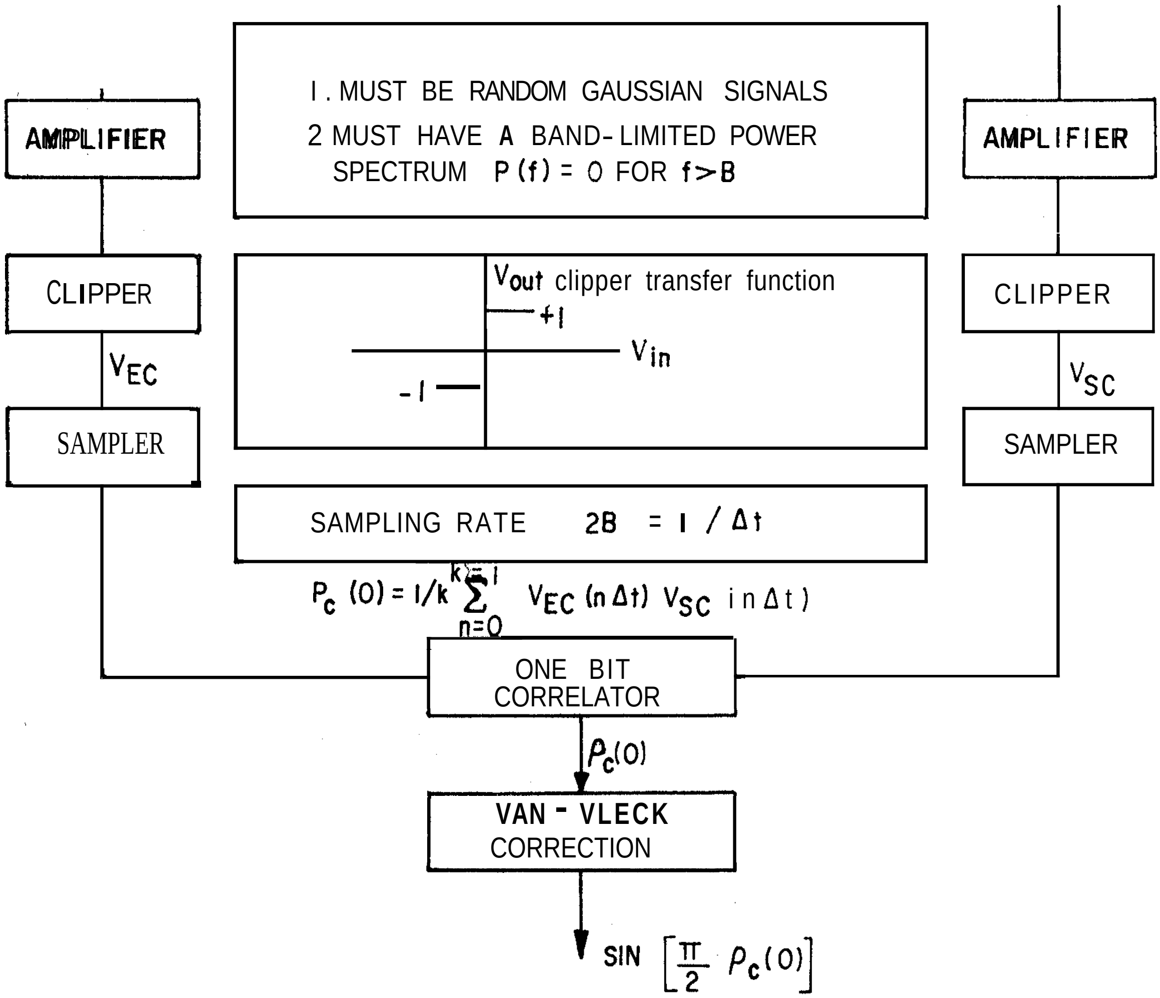


FIG. 2·9 **ONE** BIT CROSS-CORRELATOR.

$\rho_M(0)$ is determined by using the Van Vleck relation.

$$\rho_M(0) = \sin\left[\frac{\pi}{2}\,\rho_c(0)\right] \tag{2.6}$$

The reasons for the choice of one-bit correlators become evident by the following arguments.

1. As mentioned earlier quantisation and sampling results in loss of SNR compared to that of an unquantised correlator. One can define' the degradation factor of a digital correlator as

$$D = \frac{\text{Output SNR of an analog correlator}}{\text{Output SNR of a digital correlator}} . \tag{2.7}$$

D is a function of the number of quantising levels and sampling rate used. The degradation factor of a one-bit correlator is 1.56. (Weinreb, 1963). This is true when the sampling is done at Nyquist rate. Since the clipped signal is not band limited, the degradation factor can be reduced by over sampling. Degradation factor versus normalised sampling rate is shown in Fig. 2.10. It also shows the degradation factor for correlators with various numbers of quantised levels. The figure which is due to Bowers and Klinger (1974) clearly indicates that by over sampling, the D of a one-bit correlator can be reduced. In the present receiver the IF bandwidth is 400 KHz. In the DSB technique this is further reduced by a factor of two. The sampling frequency used is 2 MHz. Thus the system employs a sampling

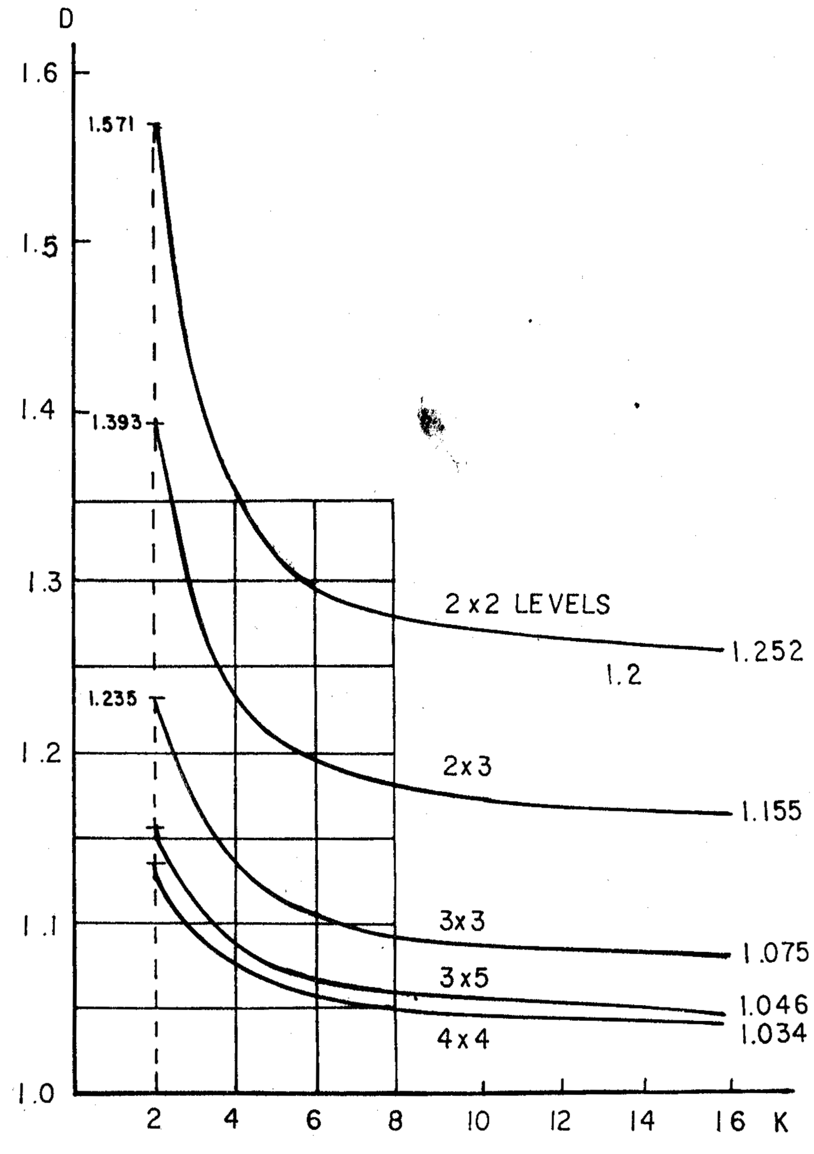


FIG. 2·10 DEGRADATION FACTOR VS NORMALISED SAMPLING RATE FOR CORRELATORS WITH VARIOUS NUMBERS OF QUANTISED LEVELS. (BOWERS 8 KLINGER, 1974.)

rate 10 times the bandwidth. The value of D for such high sampling rates is 1.252 which is very close to a two-bit correlator's D of 1.23 for samples at the Nyquist rate. Though the sampling rate required in a one-bit correlator to attain values of D close to that of a two-bit correlator looks very high, the overall complexity of the digital hardware required for a one-bit correlator is much less compared to that required for a two-bit correlator.

2. In the one-bit correlator the signal is infinitely clipped. Thus the gain of the receiver is of no consequence. On the contrary, one needs stable gain amplifiers for a two-bit correlator. In an aperture synthesis application, this requires equal gain amplifiers in the path of all interferometers. Gain adjustments can shift with time and such readjustments can be time consuming.

3. The SNR of a two bit correlator depends on the ratio of input noise level (σ of noise) to the thresholds used (V_{TH}) for quantising. This restricts the dynamic range of the input signal over which two bit correlators can function satisfactorily (or requires automatic gain controlled amplifiers). A one-bit correlator simply requires that the input signal to the zero cross detector be very high compared to the minimum overdrive above zero required for the zero cross detector (ZCD) to saturate. It is required that the input DC error should be small and independent of the input amplitude. This does not set any

limitation on the dynamic range of the input signal. For any correlator the bandwidth of the quantiser must be much greater than the signal bandwidth. Thus the one-bit correlator imposes minimum restrictions on the RF part of the receiver system.

4. A DSB system requires four correlators to obtain the same SNR as that of a SSB system. It is easy to implement the addition and subtraction required for a DSB system in one-bit correlators. Addition before making the Van Vleck correction (which is non linear) is permitted, since the terms added in the DSB system are of the same value except for the noise. A simple technique using phased clocks and upcounters is used in the present receiver system.

There are however, a few disadvantages in the one-bit correlator system. In one-bit correlation the amplitude information of the signal is completely lost. This makes absolute brightness calibration of the sky impossible. To measure the absolute brightness of the sky, one has to measure the total power in each channel. This information is not obtainable in one-bit correlators. This thesis describes a novel way of using the one-bit correlators to measure the total power of the signal, thus making it possible to obtain the absolute brightness calibration.

Since the gain of the RF and IF amplifiers do not play any part in the one-bit correlators, it is generally believed that the grading of the elements used for the synthesis is always uniform. In one-bit correlator the bandwidth of the IF amplifiers affect the grading of the elements. Two bands which are not identical cannot give unity correlation coefficient after quantisation, even if the signals are completely coherent. In this thesis we discuss the SNR of one bit correlators when the two signals have different bandwidths.

2.5 HYBRID TECHNIQUE TO OBTAIN QUADRATURE SAMPLES OF A BANDPASS SIGNAL

A DSB system uses four correlators. These are used to obtain four cross products between the two orthogonal components of each input signal. These two orthogonal components contain independent information about the RF input band. We shall call one of the components the inphase component, and the other the quadrature component.

A conventional technique to obtain the quadrature components is shown in Fig. 2.11. Heterodyning is done in a receiver to translate the RF band of interest to a convenient operating frequency. In the GBD telescope the RF is centred around 34.5 MHz, which is not too high to be processed by available logic circuits. For example, Emitter coupled logic (ECL) circuits can toggle at over a 100 MHz rate. Even the latest Fairchild Advanced Schottky

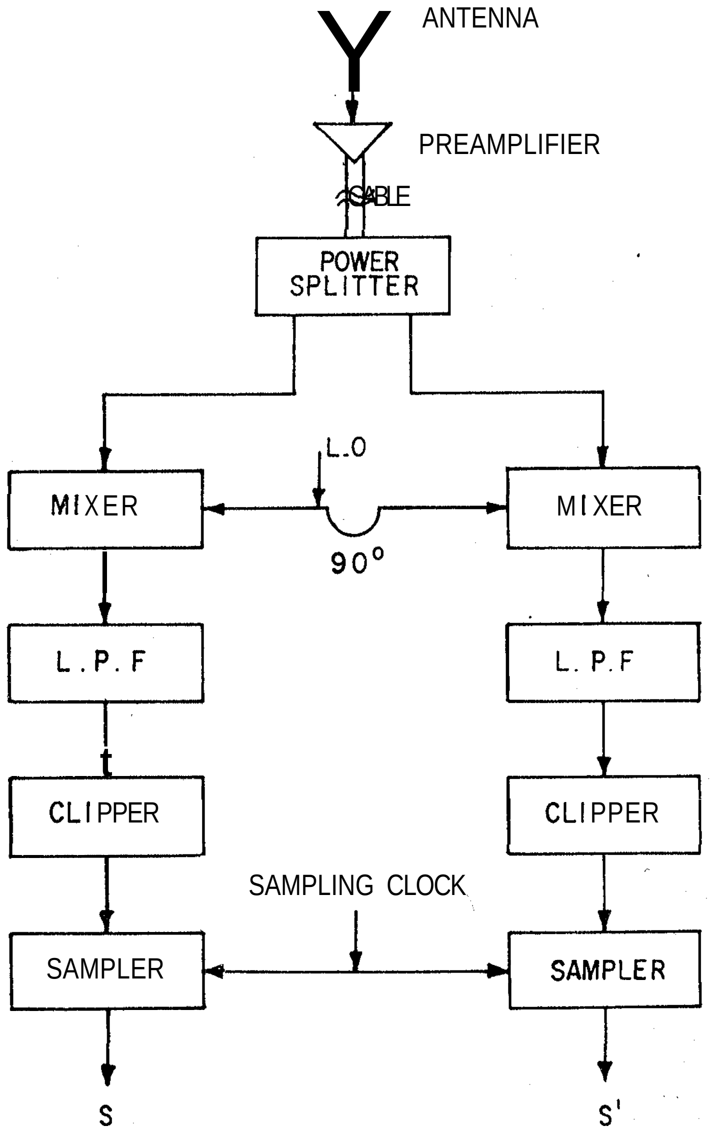


FIG. 2·11 CONVENTIONAL TECHNIQUE TO OBTAIN THE QUADRATURE SAMPLES OF A BAND-PASS SIGNAL.

transistor (FAST) logic devices can toggle at rates above 100 MHz. Thus, in a low frequency telescope heterodyning is not essential. Fig. 2.12A shows a technique to obtain quadrature samples of a bandpass signal. This technique utilises two pulse trains which are separated in time equal to 1/4th the period of the centre frequency of the band. (equivalent to a 90^{o} phase shift of the centre frequency). A shift T in the time domain changes the phase spectrum by an amount $\omega_c T$, which in this case is 90^{o}. .This interlaced sampling is simply equivalent to two mixers using LOs shifted in phase by 90^{o}. If the LO frequency is equal to the centre frequency of the RF band, it produces spectral overlap of the USB and LSB. An effect equivalent to this can be obtained by sampling at a frequency equal to a subharmonic of the centre frequency. In order to satisfy the Nyquist theorem, the total number of samples/second obtained by the two pulse trains should be greater than twice the RF bandwidth. These two pulse trains result in independent samples of the same RF band. This can be seen by looking at the autocorrelation function of the RF band (refer Fig. 2.12B). The centre frequency of the GBD telescope is 34.5MHz. So one needs pulse trains separated in time by 7.246 nanoseconds. This' requires devices fast enough to be operated at 138 MHz (4 X 34.5). It is not impossible to design an ECL based system operating at such frequencies. The component count of such a system is very small compared to the conventional system. But such a system design has several disadvantages.

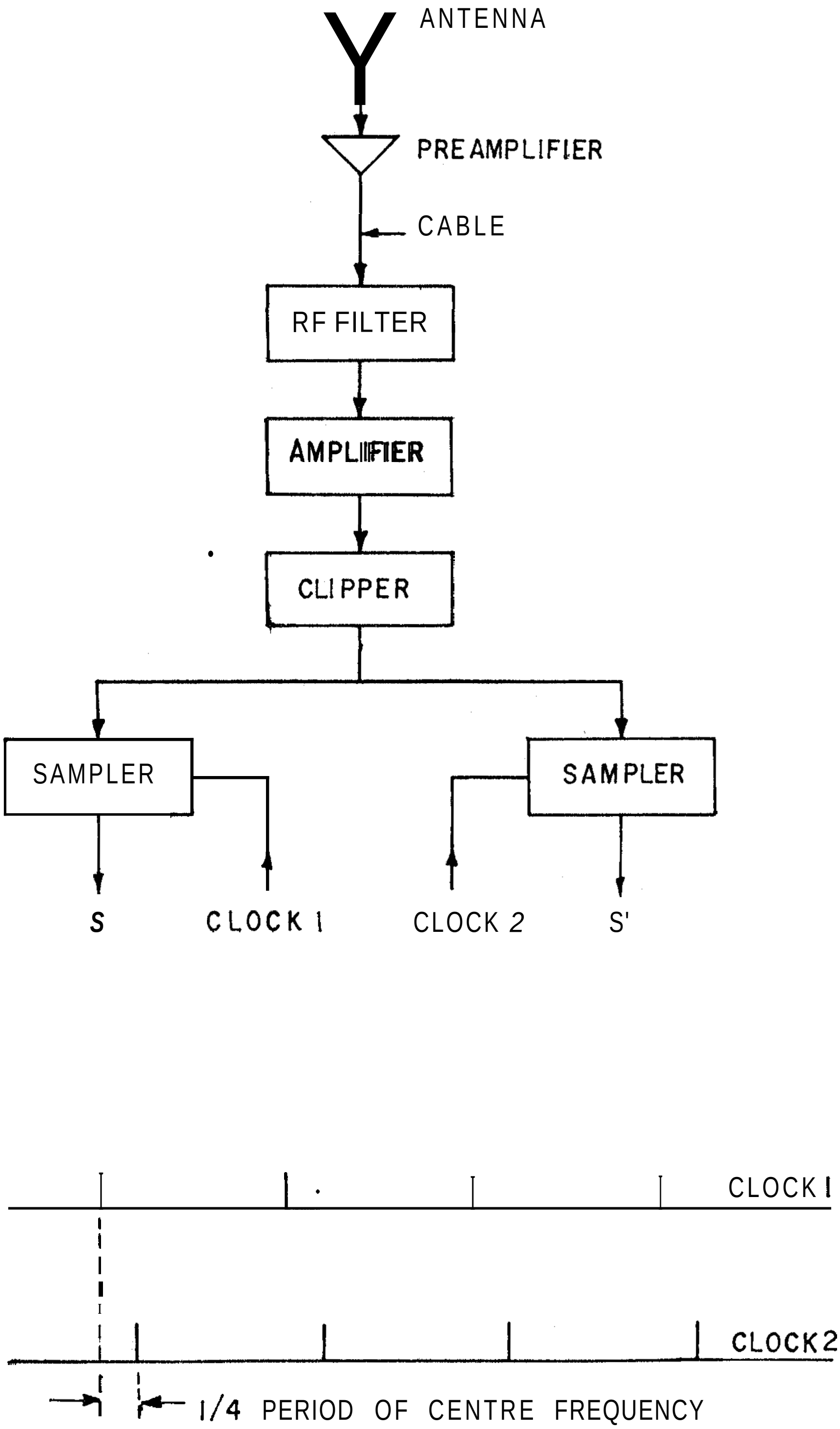


FIG. 2·12A. FAST SAMPLING TECHNIQUE TO OBTAIN THE QUADRATURE SAMPLES OF A BAND-PASS SIGNAL.

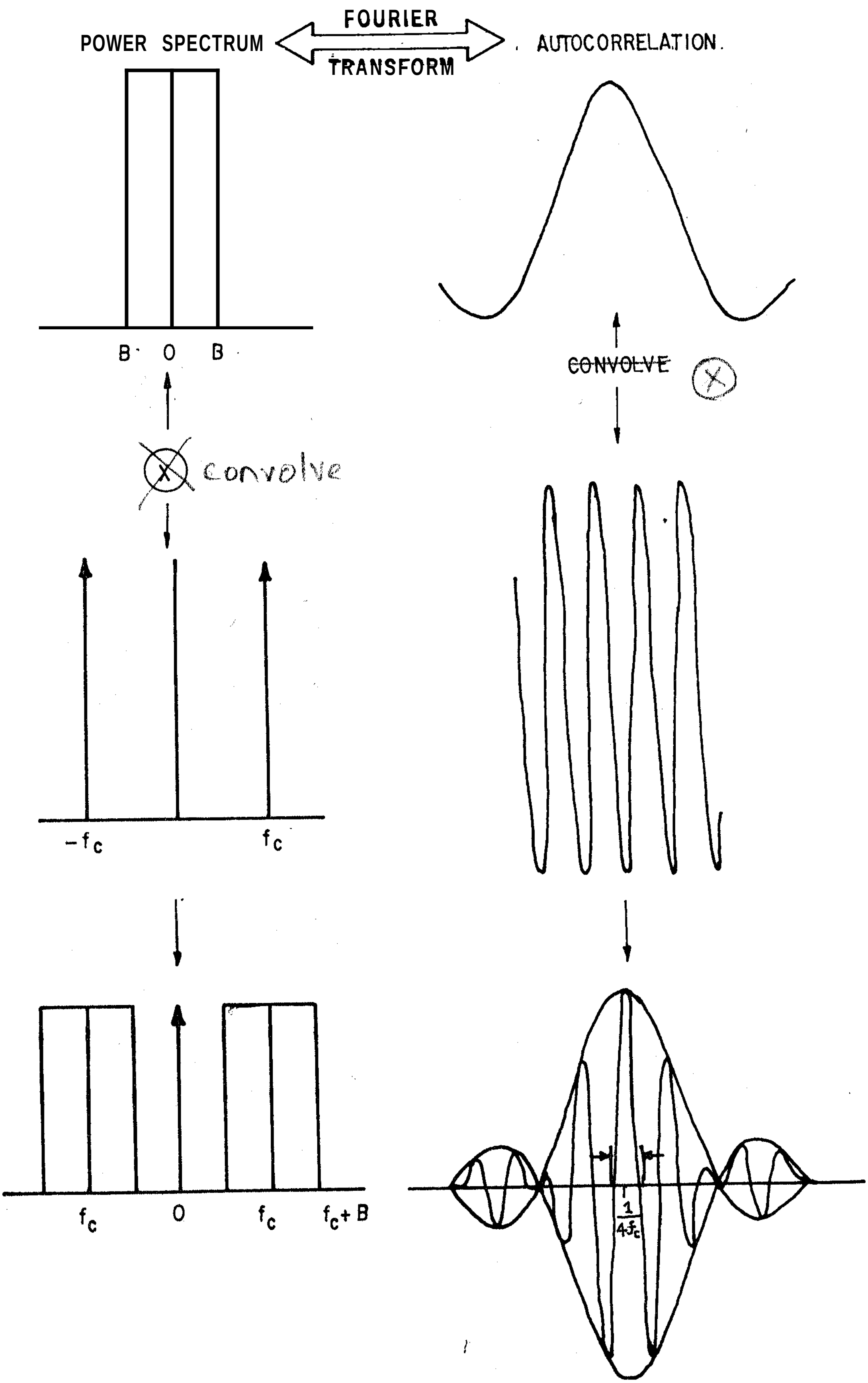


FIG. 2·12B AUTOCORRELATION FUNCTION OF A BAND-PASS SIGNAL.

1. Some of the disadvantages of the ECL systems are (a) System noise margin is low. (b) Requires very careful layout on PC cards (c) Difficult to interface with the other logic circuits. (d) LSI devices are not available in ECL. Thus part count tend to be excessive.

2. Even a one nanosecond jitter in the sampling clock causes phase errors of the order of 15°.

3. It is difficult to change the operating bandwidth and particularly make them narrower, since this calls for very high 'Q' coils.

4. One has to change the shift between the pulse trains with observing frequency.

A technique which overcomes most of the above disadvantages is shown in Fig. 2.13. This uses the best of both the bandpass sampling technique and the conventional technique, and may be called a Hybrid Technique. The GBD receiver uses an IF of 4 MHz and a sampling frequency of 2 MHz which is a subharmonic of 4 MHz. Advantages of the hybrid technique are:

1. Table 2.1 clearly indicates that the device count is less than that for a system using the conventional technique.

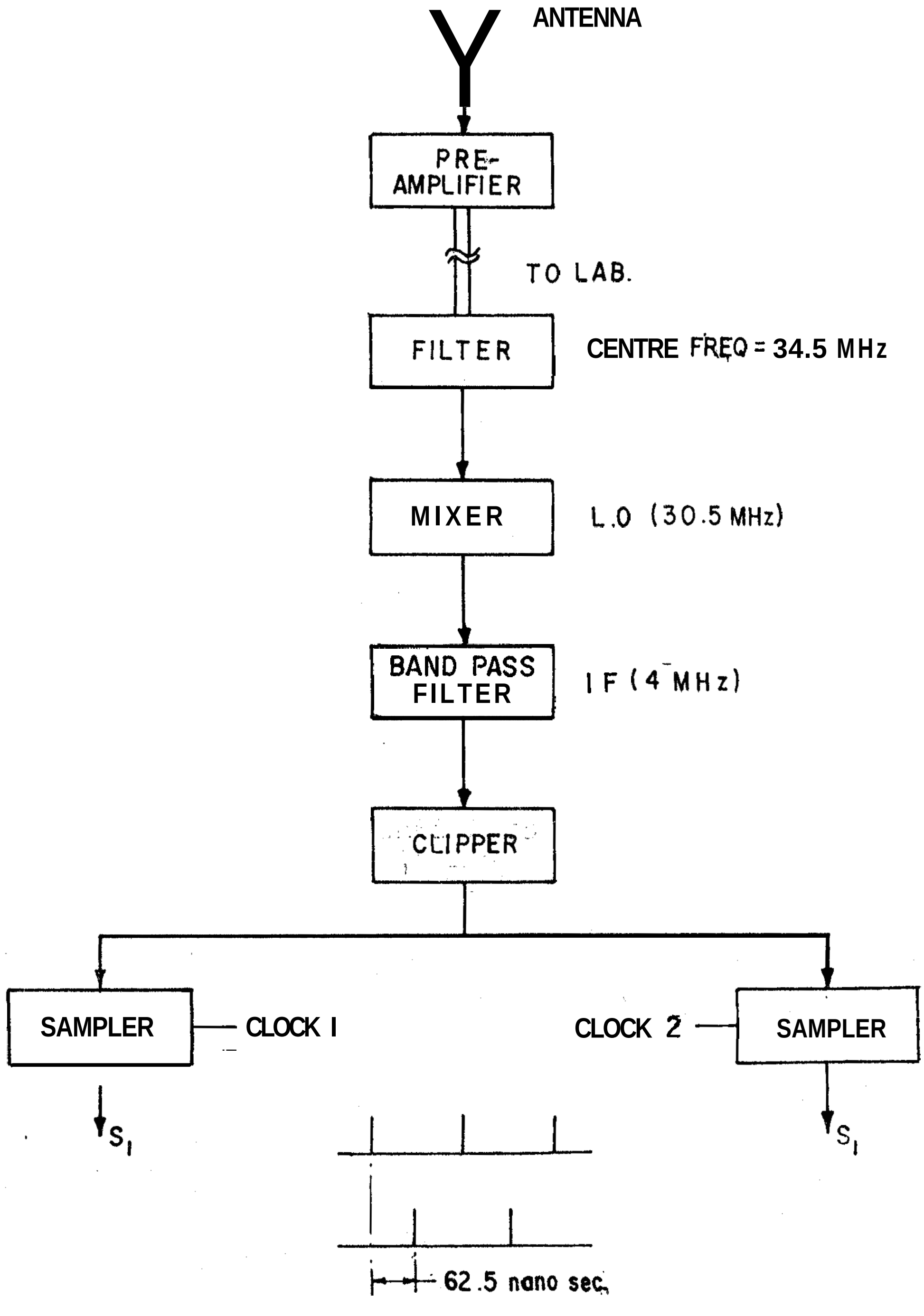


FIG. 2·13 HYBRID TECHNIQUE TO OBTAIN THE QUADRATURE SAMPLES OF A BAND-PASS SIGNAL.

FOR A 32-CHANNEL SYSTEM

TECHNIQUE	NO. OF R.F. FILTERS	MIXERS	VIDEO FILTERS	Z.C.D	SAMPLERS
CONVENTIONAL	NOT ESSENTIAL	64	64	64	64
BAND PASS SAMPLING	32 (SHARP CUT-OFF FILTERS)	NO MIXERS	—	32	64
HYBRID	32 (FAIRLY SHARP)	32	32 (BAND PASS FILTERS)	32	64

TABLE 2.1

COMPARISON OF HARDWARE REQUIREMENTS FOR DIFFERENT TECHNIQUES OF OBTAINING QUADRATURE SAMPLES OF A BAND-PASS SIGNAL.

2. Reduced number of zero cross detectors and filters reduces phase errors.

3. It is easy to build the RF filters needed to cut off the image band which is 8 MHz away.

4. Operating frequency can be changed by changing the LO.

5. Since the IF is at 4 MHz, it is easy to make variable narrow band filters for interference supression.

6. TTL compatible zero cross detectors and samplers can be used. The two pulse trains required for sampling are now separated by 62.5 nanoseconds. FAST devices which can toggle over 100 MHz can. be safely used since they don't introduce any appreciable phase errors.

The only limitation of this technique stems from the fact that the products $\overline{S_1 S_2'}$ and $\overline{S_1' S_2}$ are not formed on the samples of signals obtained at the same instant of time. This introduces "Bandwidth decorrelation" and limits the usable bandwidth. In low frequency telescopes however, man made interference generally limits the usable bandwidth and hence this disadvantage does not pose any serious problem. For example, due to interference considerations the usable bandwidth at GBD is around 400 KHz. After folding this reduces to 200 KHz. Correlation on samples of a 200 KHz noise band separated in time by 62.5 nsecs. results in negligible bandwidth decorrelation. This reason and the design ease of the R.F. filters led to the choice of 4 MHz I.F. A simple experiment was set up to confirm that the conventional technique and the hybrid technique were exactly equivalent even though the sampling was done after quantisation. (Refer Fig. 2.14). The sampling clock and LO used are coherent. To compensate for the group delay introduced by the mixer and the IF filter,,a delay generator consisting of Schottky inverting buffers was used in the path of the sampling clock. The correlation between S_1, S_1' and S_2, S_2' which are the quadrature samples of an RF band, obtained by two different techniques was above 95% after applying the Van Vleck correction. 5% decorrelation is expected due to the noise from the mixers, dissimilarity in the zero cross detectors and samplers, difficulty in DC coupling the output of the video filter and the finite size of the delay steps obtainable.

2.6 CROSS CORRELATION OF TWO UNEQUAL BANDWIDTH SIGNALS IN A ONE-BIT CORRELATOR

Two major obstacles to low 'frequency radio astronomy observations are interferometer phase fluctuations caused by the ionosphere, and the high level of man made radio interference. Two major design goals Of the present receiver system are:

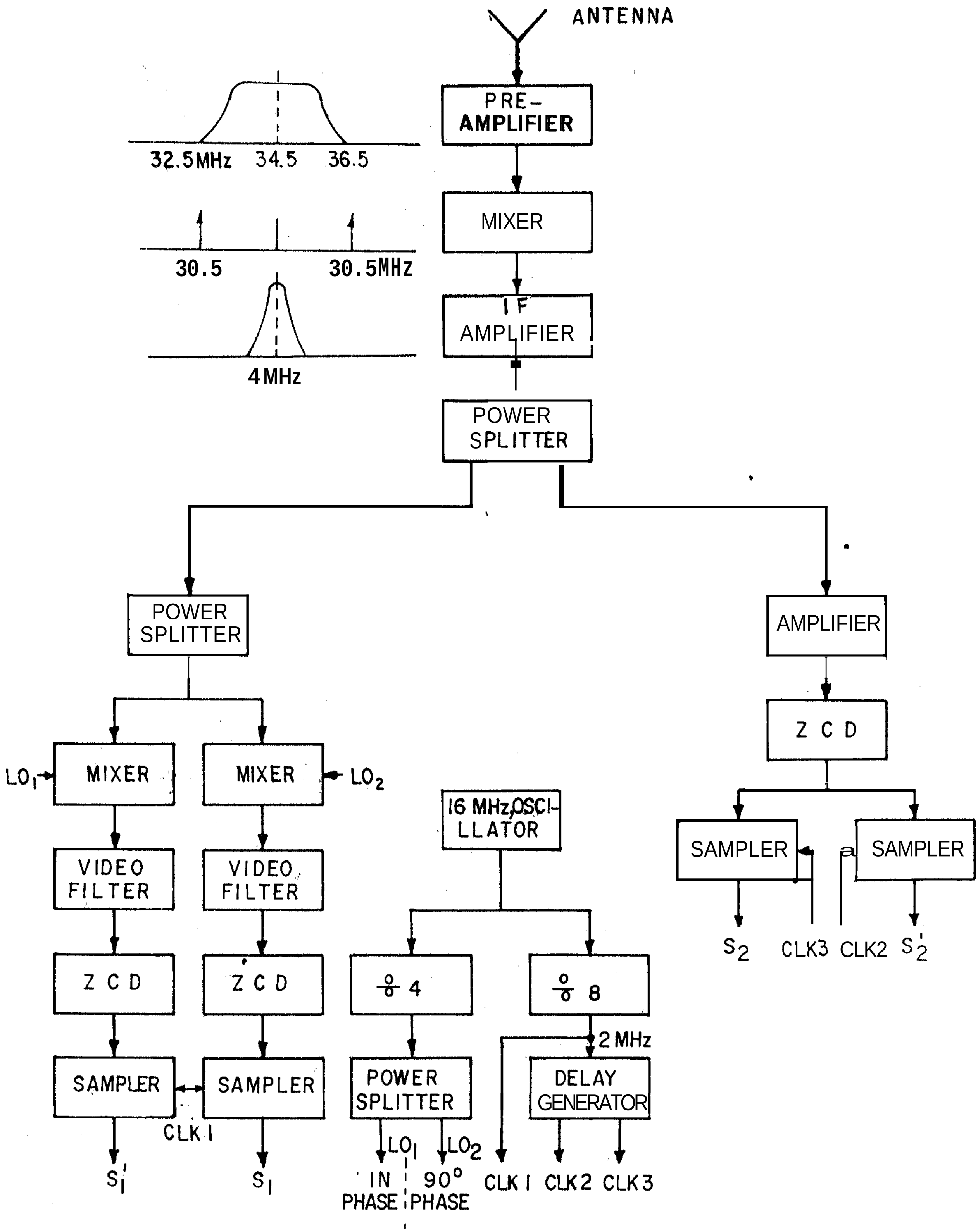


FIG. 2.14 EXPERIMENT TO SHOW THE EQUIVALANCE OF HYBRID AND CONVENTIONAL TECHNIQUES OF GENERATING QUADRATURE SAMPLES.

1. To obtain observation of a large patch of sky in a short time. This is to avoid the time variability of the ionospheric effects, which can cause an apparent shift in the position and intensity of the radio sources. Multibeam forming capability of the receiver system achieves this goal.

2. To obtain long periods of interference free observation.

Steps taken to achieve this objective are (a) The Radio Frequency Interference (RFI) spectrum was monitored to decide the bandwidth and the centre frequency of operation. (b) Ability to alter the centre frequency and the bandwidth of the observation were built into the receiver system.

An "Interference Monitor" has been incorporated into the system to help the observer to choose the centre frequency and bandwidth of operation.

2.6.1 RFI Monitoring

Light pollution is a well known problem to optical astronomers. It has become increasingly difficult to find a dark site for an observatory. Radio Astronomy faces a similar situation. The consequences of interference can lead to at one extreme to total disruption of observations, or in the other, to very subtle distortions of the signal.

1. When the principal frequency of the interference lies in the band of interest and is very strong, it can ruin the entire observation. However, such high levels of interference are normally intermittent and only occassionally continuous. Software techniques can be employed to excise intermittent strong interference.

2. Even though the radio-telescope is not pointing in the direction of the interference, it can still pick up the interfering signal through one of Its sidelobes. Harmonics of the principal Interference frequency may get picked by the receiver. These are normally low level interference and are generally continuous in time. They may go unnoticed while observing strong sources. But they affect the sensitivity of the telescope and render weak sources and spectral line observations impossible.

International Radio Regulations contain the table of frequency allocations in which servicks are alloted bands from 9 KHz to 275 GHz. Radio astronomy was officially recognised as a service at the 1959 World Administrative Radio Conference (WARC) (Vernon Pankonin, 1981). For decameter radio observations, there is a band from 37.5 to 38.25 MHz specially allocated to radio astronomy users. The well known WKB maps of the sky were made at 38MHz (Williams et.al. 1966). The Gauribidanur telescope started operation initially at 30 MHz in 1980, but good quality observations were rendered impossible by interference. RFI monitoring indicated that the band around 34.5 MHz was relatively free.

It was found more suitable than the official band around 38 MHz. Initial observations were made at 34.5 MHz using a bandwidth of 50/200 KHz. Then, to choose an appropriate bandwidth the RFI was monitered using the set up shown in Fig. 2.15. The antenna shown consists of 16 dipoles (shown in Fig.2.4), 4 rows arranged in the E-W and 4 rows in the N-S direction. Since the LO frequency was chosen to be in the middle of RF band, the effective RF bandwidth was twice the video bandwidth. The output of the 4 detectors were recorded on a chart recorder. Some important observational results are :

1. Nights are relatively interference free. Continuous interference starts around 10 a.m. on most of the days and dies down only after 6.00 pm.

2. Interference free time was almost the same for 50 KHz, 100KHz and 200 KHz IF bandwidth receivers. The 500 KHz receiver showed less interference-free time. Man made interference is generally very strong compared to cosmic noise. Thus one requires filters with very high stop band attenuation.

3. Variable bandwidth receivers are useful in tackling only low level interference.

On the basis of the above observations, the following conclusians were drawn.

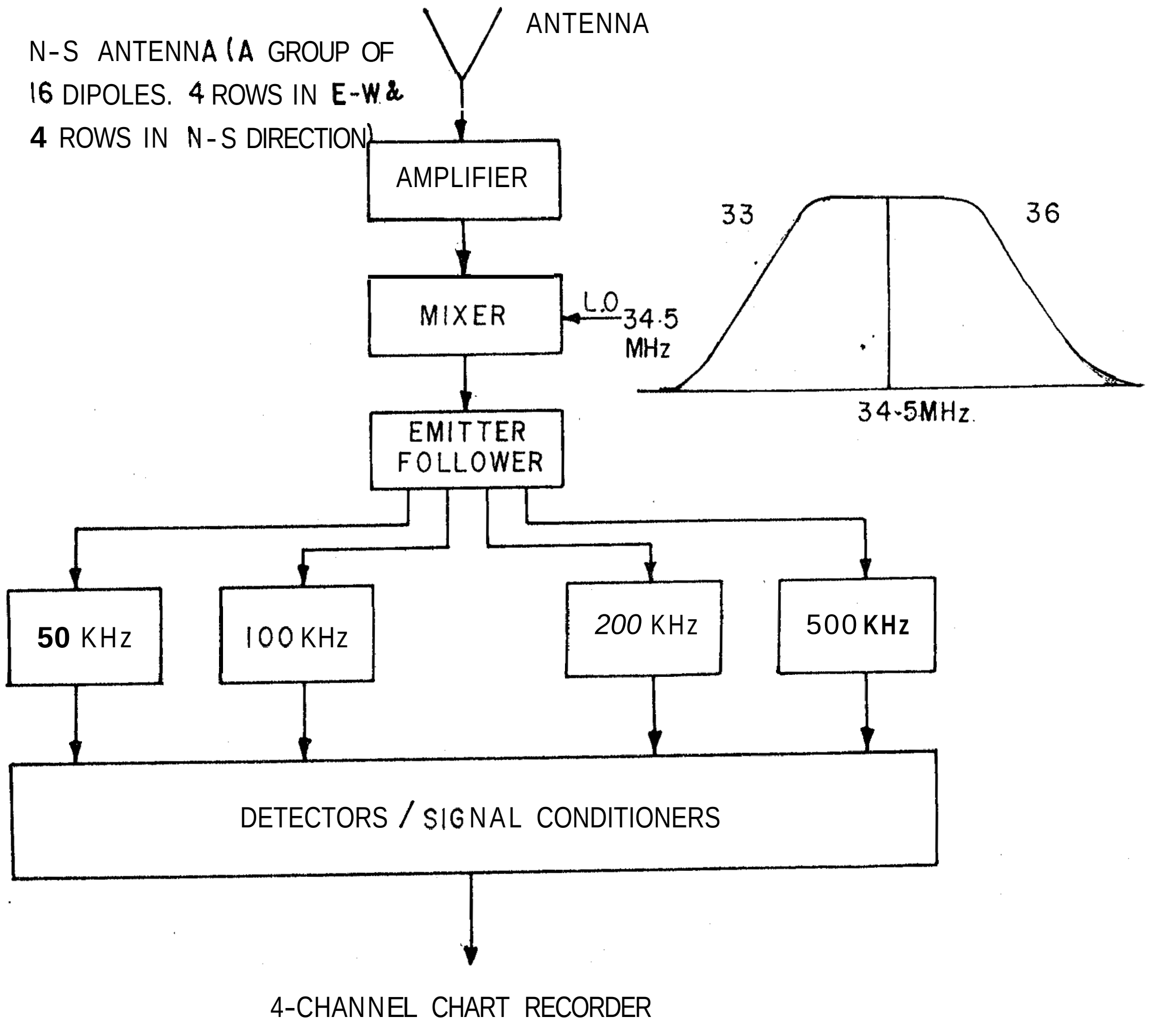


FIG. 2·15 RFI MONITORING SCHEME.

1. A bandwidth of around 400 KHz is optimum for observations.

2. It should be possible to change the operating frequency at least over a 2 MHz band (say from 33.5 to 35.5 MHz).

3. Interference free observation can be increased by having variable bandwidth filters for all the interferometers. But the experiment showed that the complexity of the hardware and its cost is not commensurate with the increase in observation time obtainable. To tackle low level interference, it is enough to cut down the bandwidth of the E-W signal only. Thus a variable Bandwidth receiver was incorporated in the E-W arm alone.

Attempts were made to vary the bandwidth of the receiver after one-bit quantization. The appendix A describes the one-bit filters. They were found suitable for cutting down the bandwidth. But they run into serious problems when strong narrow-band interference is present.

Let us examine the aspects 2 and 3 in greater detail.

The outputs of the E-W array and the 23 N-S groups have to be brought to the laboratory. The length of the E-W array is 1367 m and its centre is 400 m from the laboratory. Thus the signal from each dipole of the E-W array travels a cable length of 1767 m before reaching the lab. Each N-S group is only 20 mts long. If each of the N-S group outputs is brought to the lab by the shortest possible lengths of cables, the N-S signals travel distances varying from 300 to 600 m before reaching the lab. This large difference in the cable lengths at the input of the correlator causes (a) Bandwidth decorrelation. (b) An RF phase-calibration change with centre frequency. For a difference in cable length of around 1 km (delay of 3 μ secs), the phase changes by 360°, for a shift in the centre frequency of 333 KHz. The RF instrumental phase changes by different amounts for different N-S groups. To overcome the above effects, a large delay in the IF stage is introduced in each N-S group. A variable delay in the path of the E-W signal is adjusted to maximise the correlation. The lengths of cables bringing the various N-S group outputs to the lab are equalised. This makes the RF phase to change by the same amount in all the 23 correlators. Now the RF instrumental phase change becomes a function of the centre frequency only. The variation of the instrumental phase was measured and stored in a look-up table. This look-up table is made to control a phase shifter in the path of the E-W LO. The lookup table is addressed by the synthesizer used as the LO which will also determine the centre frequency of operation. The phase shifter maintains the RF phase constant throughout the band of interest. The scheme used is shown in Fig. 2.16.

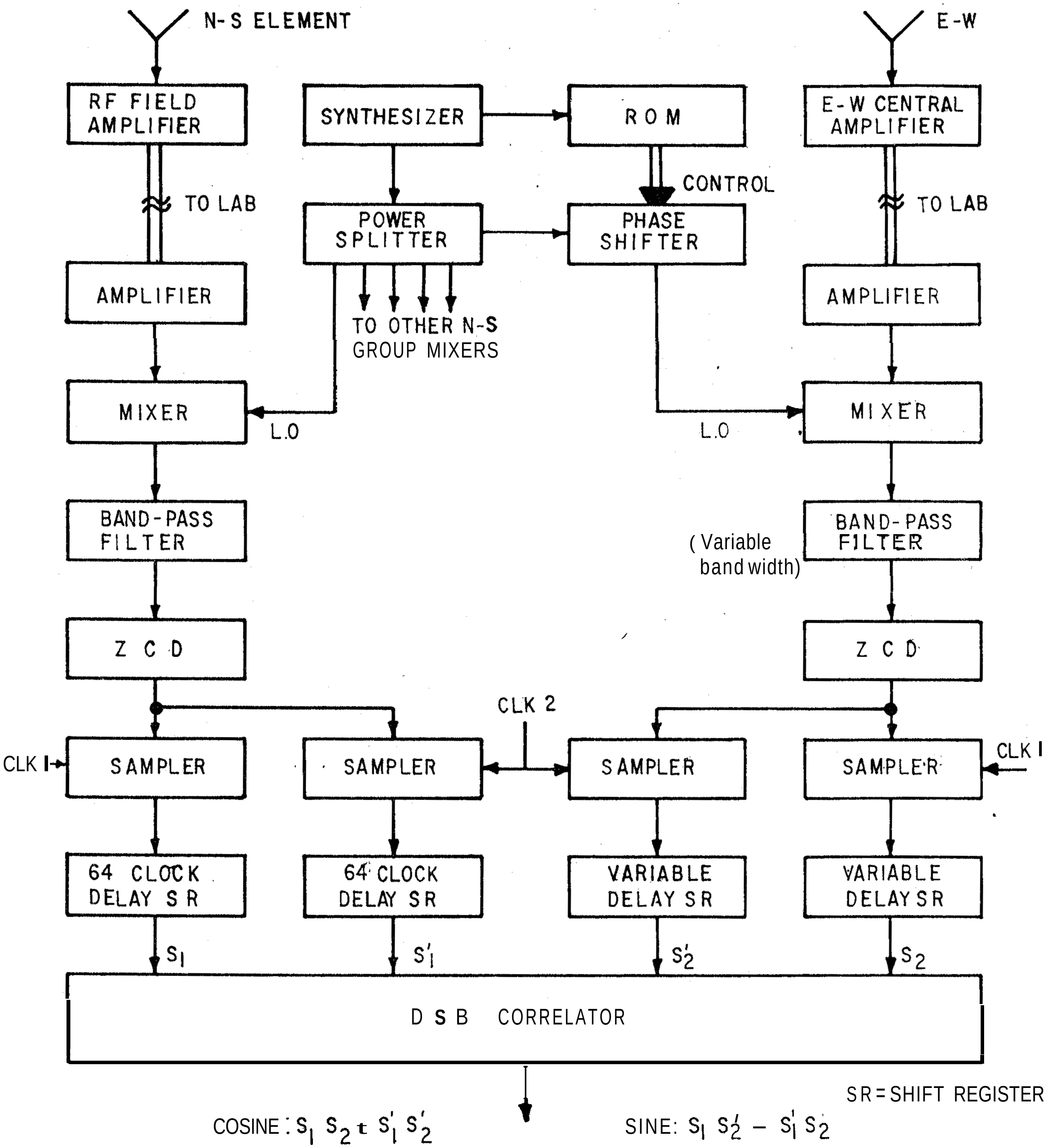


FIG. 2·16 BLOCK DIAGRAM OF THE SCHEME USED TO OBTAIN ONE OF THE COMPLEX VISIBILITIES.

2.6.2 Effect Of E-W Variable Bandwidth On The Measured Correlation Coefficient And Its SNR.

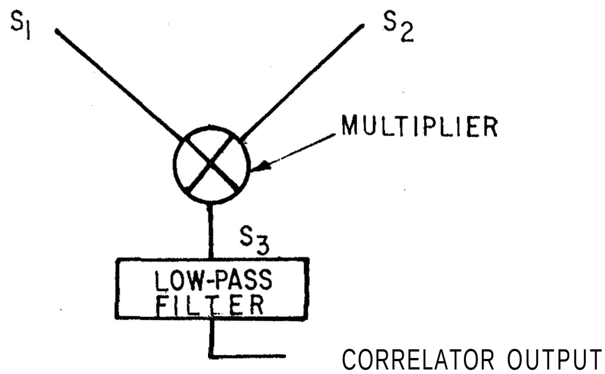
A one-bit correlator measures the normalised correlation coefficient. If ρ_A is the output of an analog correlator, a one-bit correlator with the same inputs will measure a correlation coefficient of

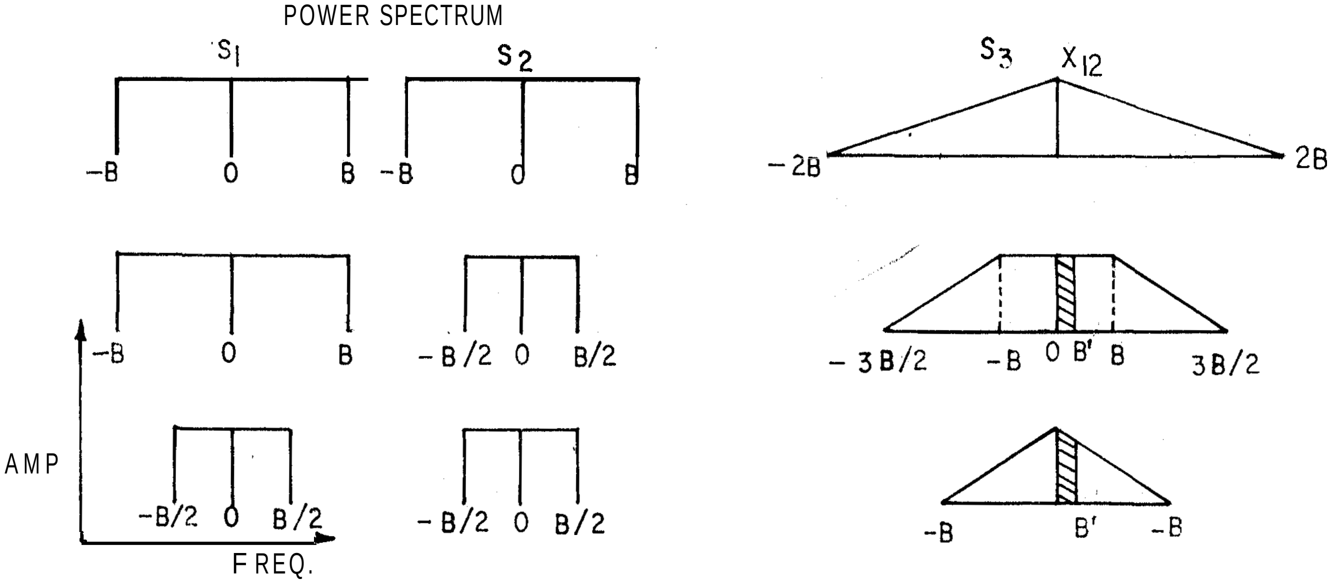
$$\rho_c = \rho_A / \sigma_1 \sigma_2 \qquad (2.8)$$

where σ_1, σ_2 are the R.M.S. noise fluctuations of the input, signals. The bandwidth B of a signal and its R.M.S. fluctuation σ are related by, $\sigma^2 = B$, if one assumes a rectangular pass band signal of unit height. Fig. 2.17 sketches the power spectrum of the E-W and N-S signals and the correlator output for various values of their bandwidths. The Table in Fig.2.17 lists the output of analog and one-bit correlators, for various input bandwidths. It is clear from the table that a one-bit correlator measures a reduced correlation coefficient for unequal bandwidth signals. Since the reduction is linearly proportional to the bandwidth ratios, it does not affect the measured brightness distribution.

2.6.3 SNR Of An Analog Correlator With Unequal Bandwidth Inputs.

The D.C. power at the output of the correlator represents signal power. The A.C. power (fluctuating power) which lies within the pass band of the integrator





INPUT BANDWIDTH		O/P OF AN ANALOG CORRELATOR	O/P OF A ONE-BIT CORRELATOR
S_1	S_2		
B	B	X_{12}	X_{12}/B
B/2	B/2	$X_{12}/2$	X_{12}/B
B	B/2	$X_{12}/2$	$X_{12}/\sqrt{2}B$
B	B/K	X_{12}/K	$X_{12}/\sqrt{K}B$

FIG. 2·17 POWER SPECTRUM OF SIGNALS AT THE INPUT AND OUTPUT OF AN ANALOG CORRELATOR.

represents the noise power. If 'B' is the bandwidth of the Low Pass Filter (L.P.F.) then one can write the **SNR** of correlators with input signal bandwidth; B, B/2 and B/2, B/2 as

$$\mathrm{SNR} = \frac{\text{DC power at the O/P of the correlator}}{\text{AC power at the O/P of the correlator}} \qquad (2.9)$$

$$= \frac{\text{Height of the power spectrum } \mathbf{S3} \text{ at DC}}{\sqrt{\text{Area of the shaded patch}}} \qquad (2.10)$$

$$\mathrm{SNR}_{B,\,B/2} = \frac{X_{12}/2}{\sqrt{(X_{12}/2)\cdot B'}} \qquad (2.11)$$

$$\mathrm{SNR}_{B/2,\,B/2} = \frac{X_{12}/2}{\sqrt{\frac{B'}{2}\left\{\frac{X_{12}}{2} + (X_{12}/2) - (X_{12}/2B)\cdot B'\right\}}} \qquad (2.12)$$

$$\frac{\mathrm{SNR}_{B,\,B/2}}{\mathrm{SNR}_{B/2,\,B/2}} = \sqrt{1 - \frac{B'}{2B}} \qquad (2.13)$$

$$= \sqrt{1 - \frac{1}{2K}} \qquad K = \frac{B}{B'} \qquad (2.14)$$

For large values of K this is very close to unity. Thus the example chosen shows that for unequal bandwidths, **SNR** does not deteriorate in an analog correlator.

2.6.4 SNR Of One-Bit Correlator With Unequal Bandwidth Inputs

A filter present in the E-W arm introduces both delay and a phase which is a function of frequency. It is therefore necessary to delay the N-S signal and also pass it through an all pass phase compensating network to retain the correlation value. This defeats the very purpose of building variable bandwidth filters only for the E-W signal. Hence we should consider SNR only when the delay is compensated and the phase is left uncompensated 'since delay can be easily obtained by shift registers in one-bit correlators. The hardware configuration is as shown in the Fig. 2.16. Delay compensation will be discrete and in steps of sampling time. A gross delay is introduced in the N-S signal path. A single chip of 64 stage shift register in each group will be sufficient. A single variable delay is incorporated in the E-W arm. This can be varied to minimise the correlation loss. Appendix B describes detailed calculations of SNR for one-bit digital correlators. It clearly proves that, limiting the bandwidth of the E-W channel only does not deteriorate the SNR. The delay required to maximise the correlation was experimentally determined. The main emphasis in the design of these filters was to obtain linear phase characteristics. This ensures that the Shift Register delay compensates almost completely the phase introduced by the filter in the E-W channel. Normally, the linear phase filters do not have a sharp cut-off which is essential for interference rejection. Coupled resonators which give the best of both were chosen for the variable bandwidth filters. To help an observer to choose the bandwidth and centre frequency of. observation, and also to display the interference, an "interference monitor/indicator" was built into the system (Fig 2.18). The Interference indicator measures the autocorrelation function of the E-W signal at three randomly chosen large delays. This function is very close to zero for a pure noise band. In the presence of narrow band interference, these values will not be near zero. The autocorrelation function exceeding or becoming less than a threshold. in any one of the delays, indicates interference. The threshold depends on the bandwidth, delay at which the measurement is made, and the integration time used for the measurement. To help choose the centre frequency of observation, a system resembling a spectrum analyzer has been built. This has a fixed resolution of 50 KHz. Two sets of thumb wheels permit the selection of the total bandwidth. A commercially available syntest synthesizer is used as LO. To obtain a good SNR on the measurement, five seconds are required to measure the total power in each 50 KHz band. The display on the osciloscope will not be continuous if such large times are spent in making a measurement at a point. To obtain a continuous trace, a RAM is used to store the total power in the band which is then displayed at a faster rate. The memory is refreshed as and when new data points arrive. To increase the dynamic range

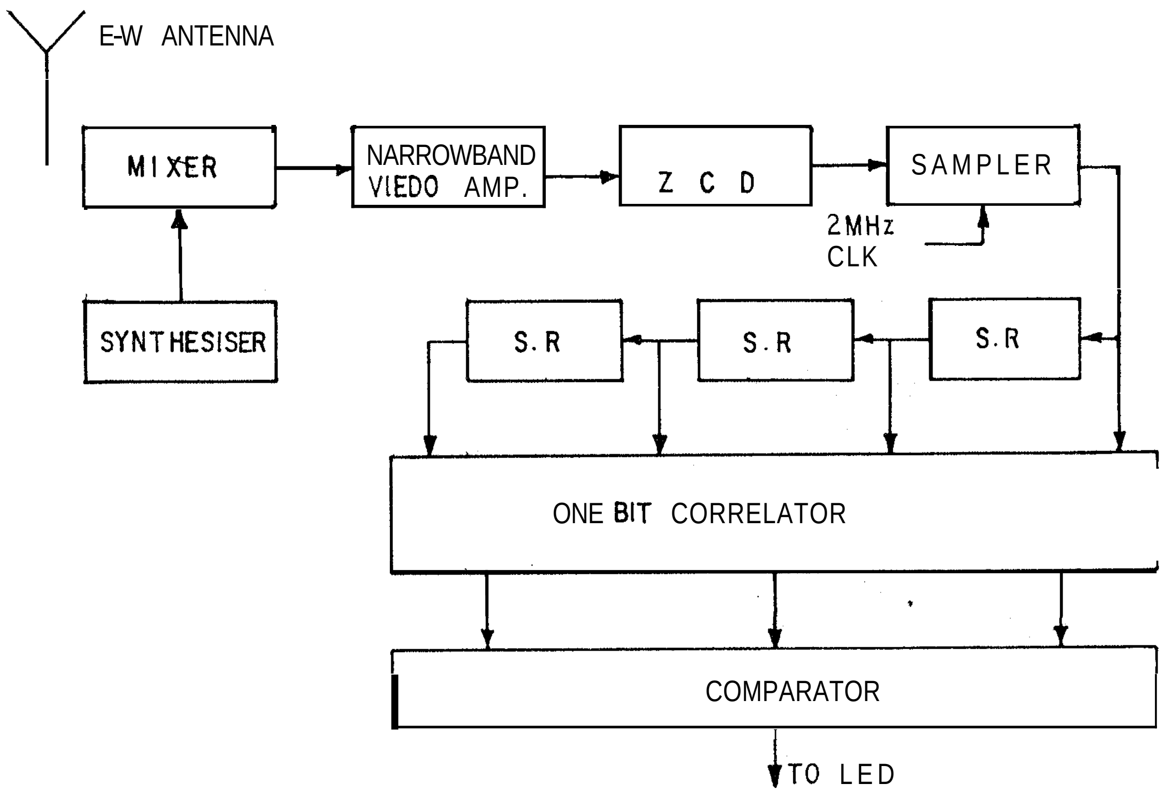


FIG. 2·18A INTERFERENCE INDICATOR.

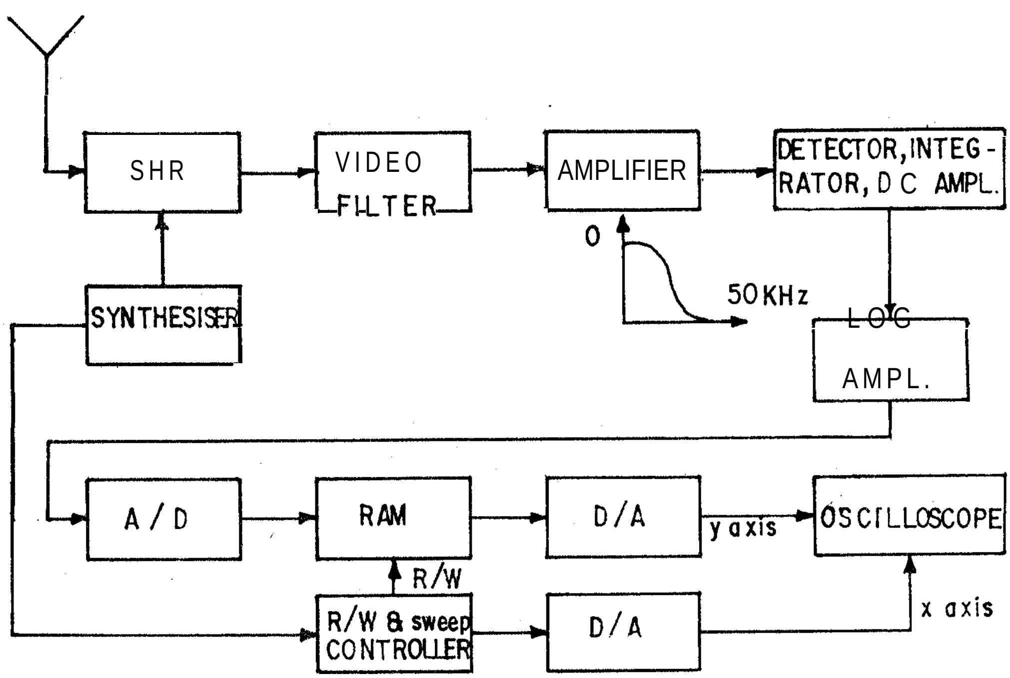


FIG. 2·18B INTERFERENCE MONITOR.

of the display, a logarithmic amplifier is used at the output of the detector.

2.7 AMPLITUDE INFORMATION USING A ONE-BIT CORRELATOR

In a one-bit correlator the amplitude information of the signal is lost. A one-bit correlator measures the ratio of the correlated to the sum of correlated and uncorrelated powers. An analog correlator measures correlated power directly.

Let us discuss the nature of the outputs of a one-bit carrelator and an analog correlator. (Ref fig 2.19)

For an analog correlator

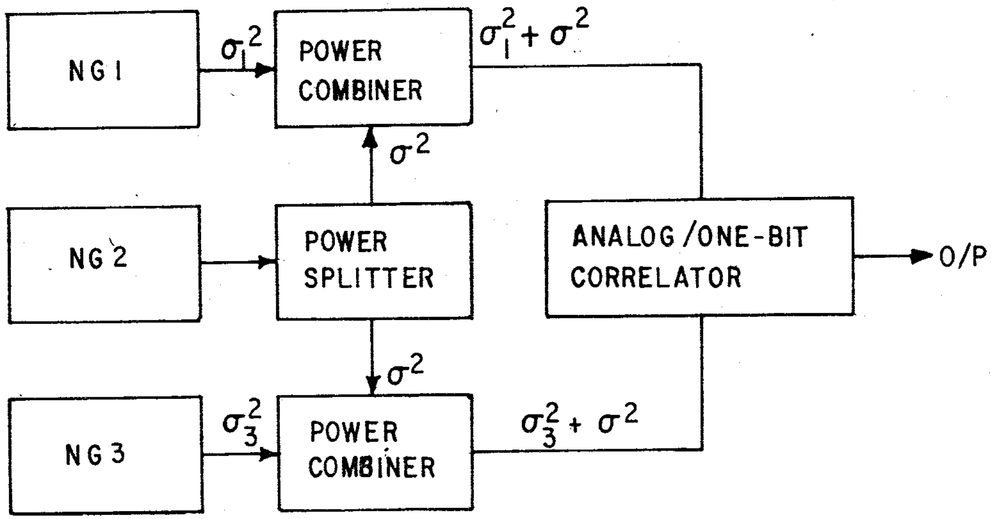
1. When $\sigma_1 = \sigma_3 = 0$, output deflection is $\propto \sigma^2$. **SNR** is finite and is a function of the integration time and bandwidth.

2. When $\sigma_1 = \sigma_3$, and is non zero(considered equal for simplicity) output is still $\propto$ to σ^2. SNR drops from its old value and is proportional to

$$(SNR)_A = \sigma^2/(\sigma^2 + \sigma_1^2)(\sqrt{B\tau}) \qquad (2.15)$$

For a one-bit correlator

1. When $\sigma_1 = \sigma_3 = 0$, independent of σ for an ideal Z.C.D. and samplers, the output is always 1 and SNR is infinite.



NG1 NG2 NG3 ARE INDEPENDENT NOISE GENERATORS

NG1 σ_1	NG2 σ_2	NG3 σ_3	ANALOG CORRELATOR OUTPUT	ANALOG CORRELATOR SNR	ONE BIT CORRELATOR OUTPUT ρ_c	ONE BIT CORRELATOR SNR
0	σ	0	σ^2	$\propto 1/\sqrt{B\tau}$	1	INFINITE
σ_1	σ	σ_1	σ^2	$\propto \frac{\sigma^2}{\sigma^2+\sigma_1^2}\frac{1}{\sqrt{B\tau}}$ $(SNR)_A$	$\propto \frac{\sigma^2}{\sigma_1^2+\sigma^2}$	$\propto \frac{2}{\pi}(SNR)_A$

FIG. 2·19 COMPARISON OF THE OUTPUTS OF ONE-BIT AND ANALOG CORRELATORS FOR IDENTICAL INPUTS.

2. When $\sigma_1 = \sigma_3 = \sigma$,and is non zero the output drops down from unity to

$$\rho_c = \sigma^2 / (\sigma^2 + \sigma_1^2) \tag{2.16}$$

$$SNR \propto \frac{2}{\pi} (SNR)_A \tag{2.17}$$

The effect of this on astronomical observations are

1. In an analog correlator the deflection due to a source is proportional to its strength. The SNR depends on the sky background in the region of the source. The region is defined by the primary beams of the interferometer elements.

2. In a one-bit correlator, the deflection itself depends on the source strength and the background in which it is present. The SNR will still be a function of the background

3. ρ_c and ρ_A the outputs of a digital and analog correlator are related by

$$\rho_c = \rho_A / (\sqrt{\sigma_1^2 + \sigma^2})(\sqrt{\sigma_3^2 + \sigma^2}) \tag{2.18}$$

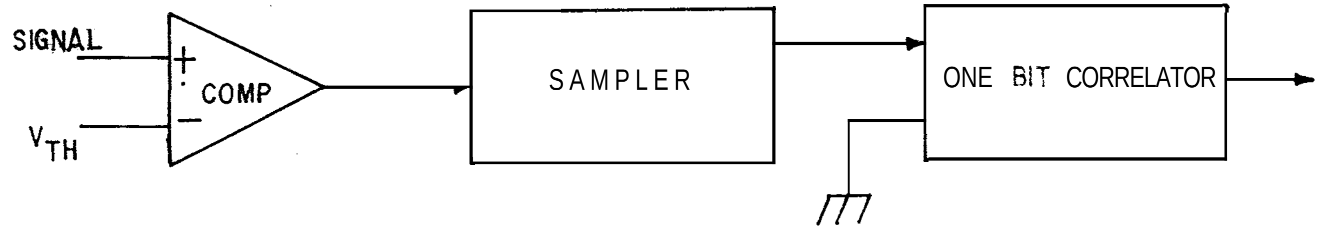
$$= \rho_A / \text{Product of input R.M.S fluctuations.} \tag{2.19}$$

Thus in observations made using one-bit correlators to obtain the absolute brightness of the sources, it is necessary to measure the total power of each interferometer output being correlated. In the present case, all the N-S interferometers have identical beams and since the output of a one-bit correlator does not depend on the gain of the amplifiers, it is enough to monitor the total power of any one of the N-S interferometers, and that of the entire E-W array.

A conventional total power receiver requires diode detectors, D.C. amplifiers and A/D converters. The present receiver system employs a novel scheme using a threshold detector and a one-bit correlator to measure the total power. If the noise statistics is known, then by determining the probability that the signal amplitude is above (or below) a certain threshold (V_{TH}), it is possible to determine the σ of the noise. In the present case, the signal is assumed to be gaussian. The best value of V_{TH} is around σ, since the slope of a yaussian is maximum near σ. Since σ will vary depending on the sky brightness it is preferable to have more than one threshold detector and correlator for each channel. The thresholds of the comparators can be calibrated by feeding a sine wave of known amplitude. The method used is shown in Fig 2.20

The advantages of the above set up compared to conventional scheme are

1. D.C. amplifiers and A/D converters are not required.



IF ρ_c IS THE MEASUAD CORRELATION COEFFICIENT, PROBABILITY THAT THE SIGNAL AMPLITUDE $< V_{TH}$

$$= \frac{1+\rho_C}{2}$$

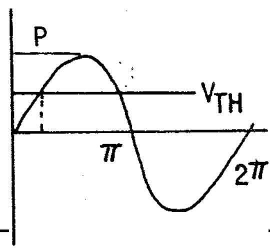
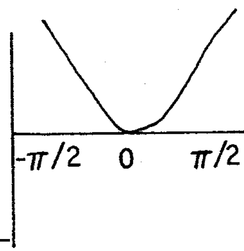
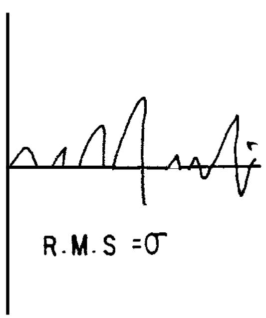
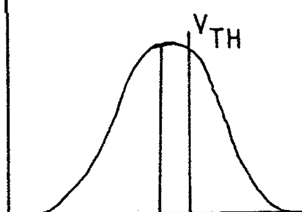
NATURE OF SIGNAL	TIME WAVEFORM	STATISTICS	PROBABILITY THAT SIGNAL AMPLITUDE IS $< V_{TH}$	REMARKS
SINUSOIDAL		PROBABILITY 	$\frac{1+\rho_C}{2} = \dfrac{\frac{\pi}{2\text{-K}} + 2\sin^{-1}\frac{V_{TH}}{PEAK}}{2\pi}$	USING THE RELATION : $V_{TH} = P\,\text{SIN}\left[\frac{\pi}{2}\rho_C\right]$ KNOWING P AND BY MEASURING ρ_C COMPARATOR CAN BE CALIBRATED FOR V_{TH}
NOISE	 R.M.S $=\sigma$	 $P(t) = \left[\frac{1}{\sigma\sqrt{2\pi}}\,Exp(-t/2\sigma^2)\right]$	$\frac{1}{2} + \frac{1}{\sqrt{\pi}}\int_0^{V_{TH}/\sqrt{2}\sigma} Exp(-\tau^2).d\tau$	USING THE RELATION $\rho_c = \phi\left[\frac{V_{TH}}{\sqrt{2}\sigma}\right]$ WHERE ϕ (Z) IS DIFINED AS $\phi(Z) = \frac{2}{\sqrt{\pi}}\int_0^Z Exp.(-\tau^2).d\tau$ ϕ(Z) IS KNOWN AS ERROR FUNCTION. ONE CAN CALCULATE σ KNOWING V_{TH}, READING Z FROM ERROR FUNCTION TABLE, OR USING AN ANALYTICAL EXPANSION OF ϕ (Z) WHEN ρ_C IS MEASURED

FIG. 2·20 METHOD OF OBTAINING THE AMPLITUDE INFORMATION OF A SIGNAL USING A ONE-BIT CORRELATOR.

2. It makes use of the existing correlators and makes recording on magnetic tape easy without the need for additional multiplexers.

3. By using more than one threshold detector, the assumption that the signal is gaussian can be checked. This assumption is the basis for using a one-bit correlator and hence is a good check.

In the present calibration system, the 'total power of the E-W antenna and any one of the N-S interferometers is recorded. For comparing observations made over a long stretch of time in any total power system, it is essential to monitor the gains of the amplifiers in the path of the signal.

In the E-W antenna there are 160 preamplifiers, 10 group amplifiers and 4 central amplifiers (2 for each arm). It is not possible to monitor the gains of all these amplifiers. If the gain changes of the amplifiers are random, the effect of 160 preamplifiers and 10 group amplifiers on the signal strength will be much less than that due to the 4 central amplifiers. Thus the gains of only these amplifiers are monitored. (Ref fig 2.21) .The N-S interferometer used for total power measurement has 4 FET amplifiers and one common amplifier. The net effect of the gain change due to all these amplifiers are monitored by injecting independent noise to the FET inputs (Ref fig 2.22). This makes gain calibration independent of the phase

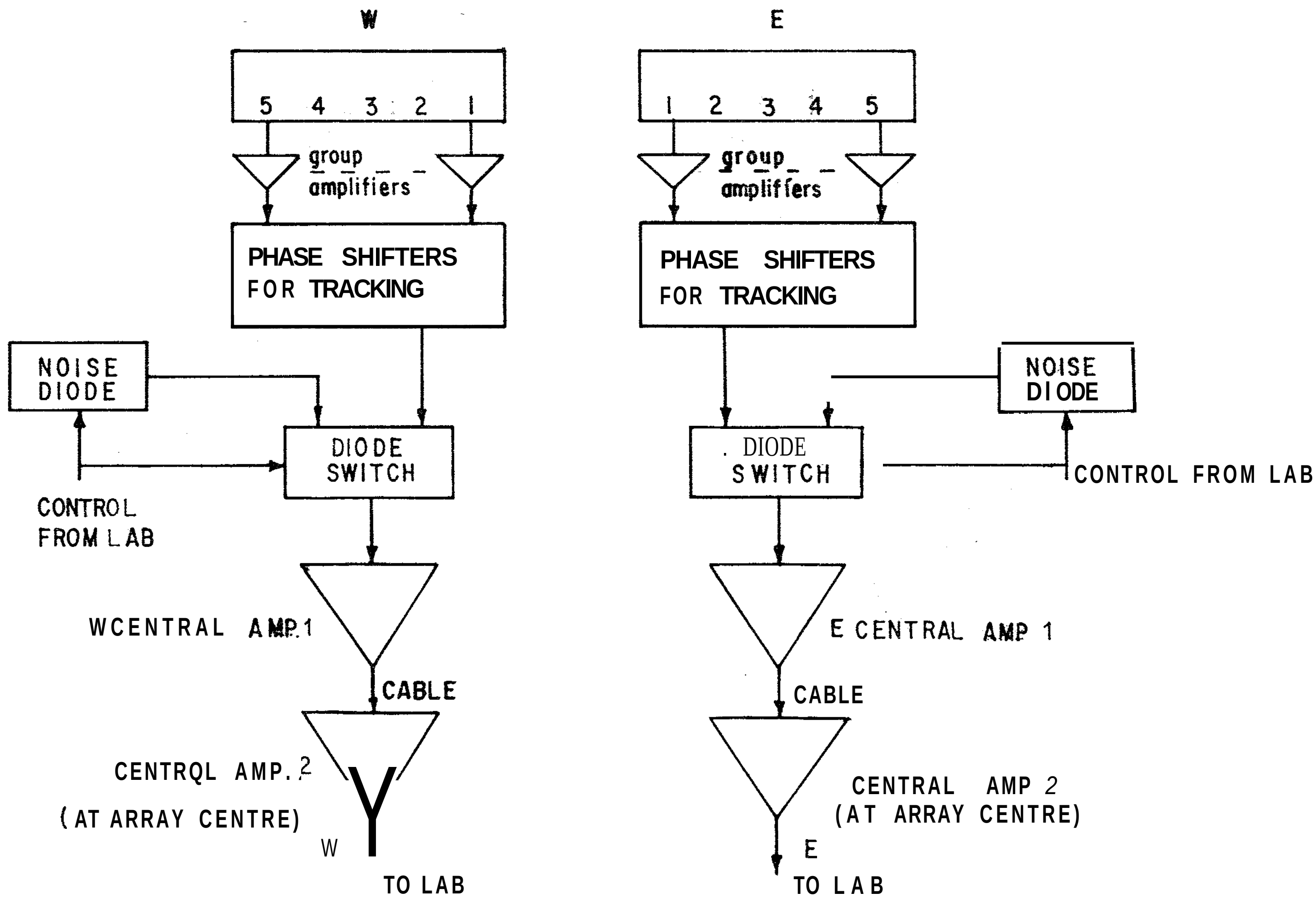


FIG. 2·21 **SCHEME USED TO MONITOR THE GAINS OF CENTRAL AMPLIFIERS IN THE E-W ARRAY.**

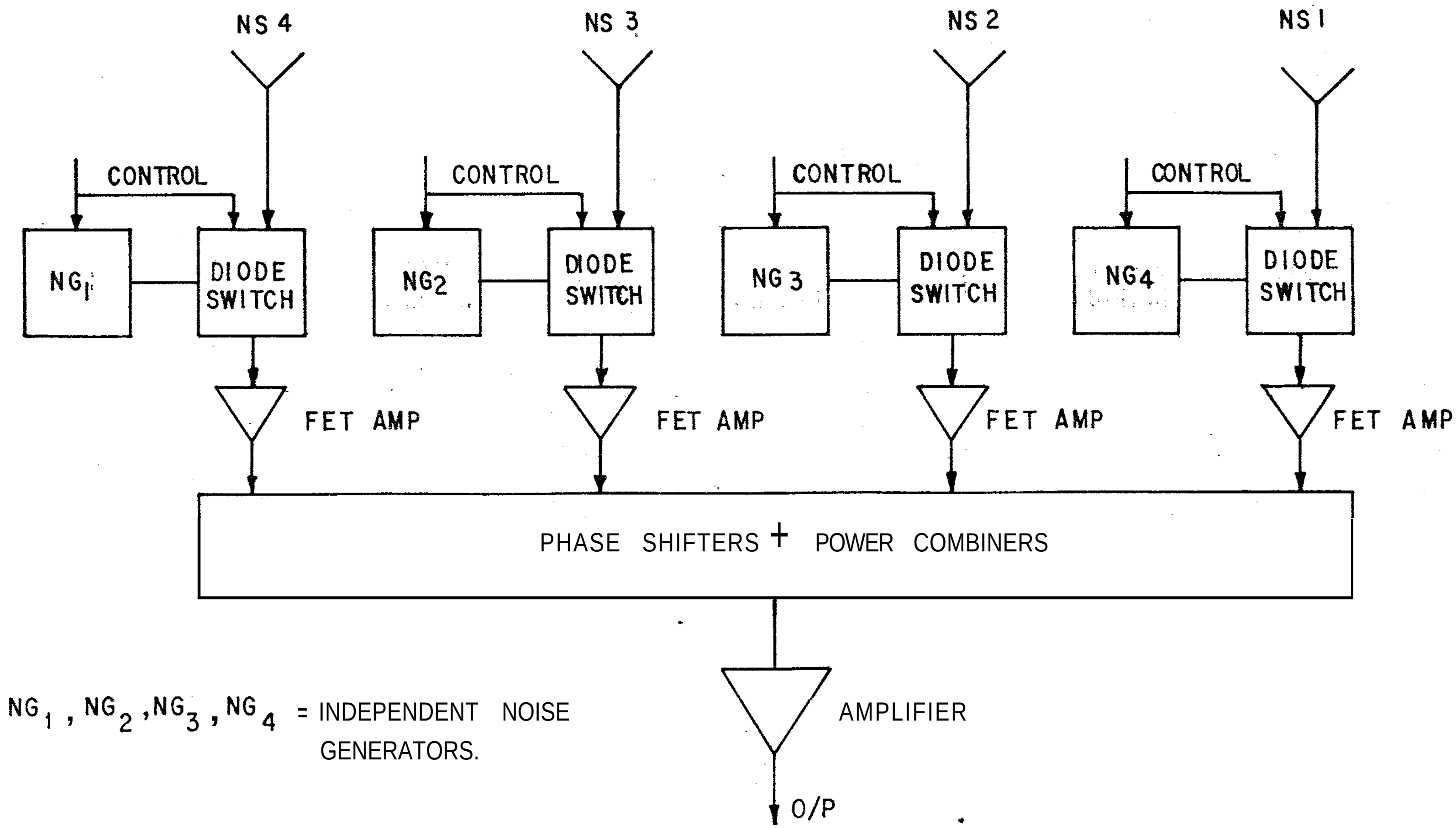


FIG. 2·22 SCHEME USED TO MONITOR THE GAINS OF FET AMPLIFIERS. AND A COMMON AMPLIFIER USED IN THE **N-S** ELEMENT OF THE INTERFEROMETER.

shifter setting. East and West arms also have independent Noise Generators since Noise diodes are less expensive than 1 km cable.(distance of the E or the W centre from the lab) The gains are monitored once an hour. This time was arrived at by leaving the noise generators on for 24 hours and recording the total power output on a chart recorder. The gain variations within an hour were found to be less than 2%.

2.8 THE ZERO-SPACING INTERFEROMETER

The E-W array in Gauribidanur is not continuous. It is broken at the centre, and the Southern arm pases through the gap. This results in an E-W beam which is surrounded by an extensive negative lobe. This means that the transfer function of the telescope which is the fourier transform of the ,beam is zero for the zero spatial frequency. Thus such a telescope does not respond to the D.C. component of the sky brightness. A map or a survey made using such an instrument will not contain the absolute temperatures of the sky, and limits the usefulness of the map for astronomical interpretation. One way to overcome this limitation is to add to the receiver output, the total power output of the missing component measured separately. Even this will not solve the problem completely, because such a system will still not be able to respond to all the frequency components of the sky in the N-S direction. The most accurate method of solving the' zero spacing problem is by treating the elements in the centre as part of both the E-W and N-S arrays.The power output of these elements can be split in a hybrid and fed to both the arrays. In a cross, the output of the central dipole should be added with equal weights to both the arms. Thus the output of a cross would be

$$(E + W + C)\ (N + S + C)$$

$$(EW + C)\ (N + S + C)$$

$$(EW + C)\ (N + C/2 + C/2 + S) \tag{2.20}$$

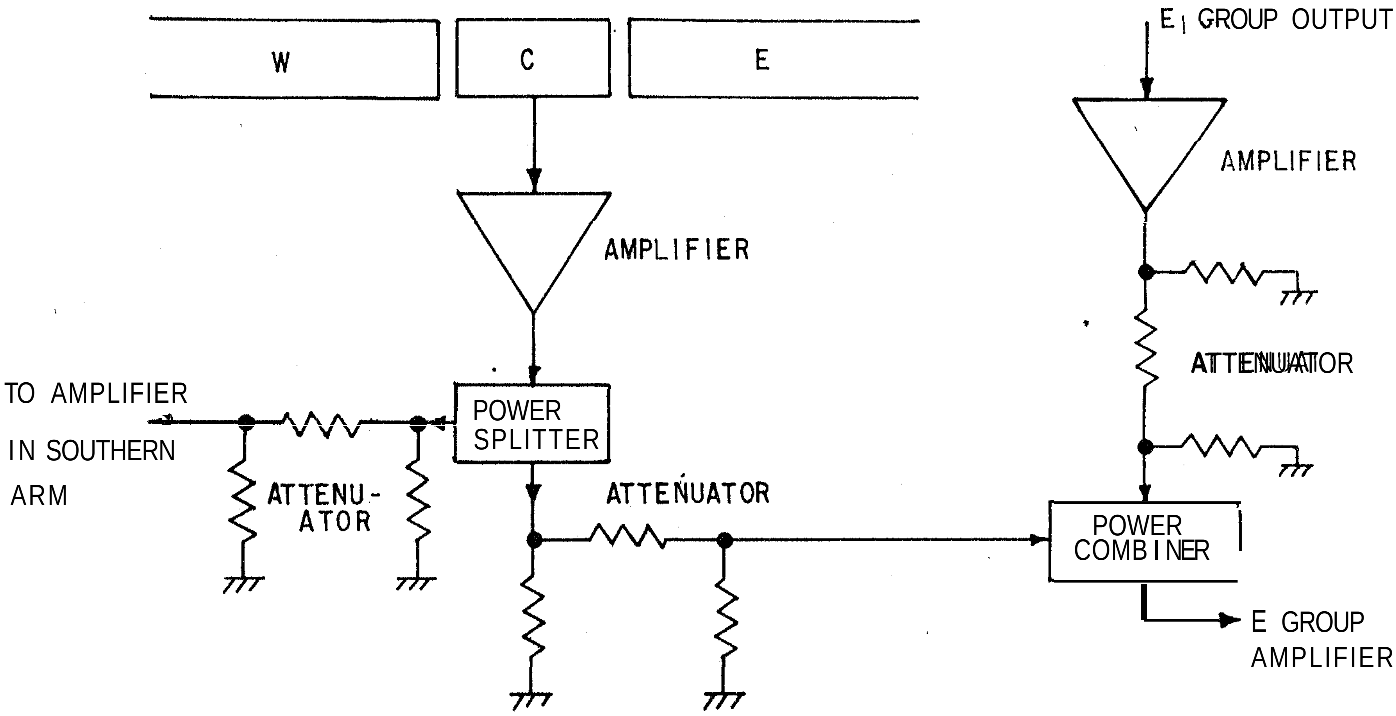
Since a T has only 1/2 the **N-S** array of a cross, its South arm should be S + C/2. Thus the output of a T antenna will be

$$(E + W + C)\ (S + C/2) \tag{2.21}$$

Thus the central part should be added with a weight of 1/2 to the South arm and 1 to the E-W arm. The schematic of the set up used is shown in figure 2.23.

2.9 THE LINE RECEIVER

The possibility that recombination lines may be seen in absorption at very low frequencies was first discussed by Shaver (1975). Konovalenko and Sodin were the first to observe an absorption line at 26 MHz towards Cas A (1980). Blake, Crutcher and Watson (1980) interpreted this as due to carbon or heavier elements. Konovalenko and Sodin have also reported attempts to detect low frequency recombination



1. C IS ADDED TO E-W ARRAY AT THE OUTPUT OF THE 1ST E-GROUP OF DIPOLES. GAIN, DELAY AND PHASE OF THE SIGNAL FROM ANY DIPOLE BELONGING TO C MUST BE SAME AS THE REMAINING DIPOLES IN THE E-W ARRAY

2. THE GAIN OF C IN THE SOUTH ARM IS HALF THE GAIN OF OTHER DIPOLES IN THAT ARM

FIG. 2·23 ZERO SPACING INTERFEROMETER TO OBTAIN THE ZEROTH ORDER FOURIER COMPONENT OF THE SKY BRIGHTNESS DISTRIBUTION.

Register circuitry and the function performed by the correlators are shown in figure 2.24. The Front end is a conventional DSB receiver. The bandpass sampling technique to produce a video band is not used here. In this technique, the bandwidth cannot be changed after conversion to video frequencies. Thus, one has to use the entire IF bandwidth for spectral observations. This produces curved baselines since it is impossible to obtain a symmetrical band of the desired accuracy. In the conventional technique, the bandwidth can be restricted after folding, to a small fraction of the RF bandwidth. Here, symmetry leads to flat baselines, or in the worst case results in a linear baseline. The correlations are recorded on magnetic tape and later Fourier transformed using a computer. The cosine transform of the cos correlation function gives the sum of the USB and LSB whereas the Sine transform of the sine correlation function gives the difference between the USR and LSB. (Ref equations in Fig 2.24). If a spectral line is present in only one of the sidebands, or if the spectral lines are asymmetrically spaced about the L O in the two sidebands, the difference between the USB and LSB automatically measures the difference between the response of the system with line present (on line) and the response of the system with the line absent (off line). This measurement is essential in all line observations because recombination lines will appear as a very small perturbation in frequency space over the underlying continuum plus receiver noise. For low frequency recombination lines of

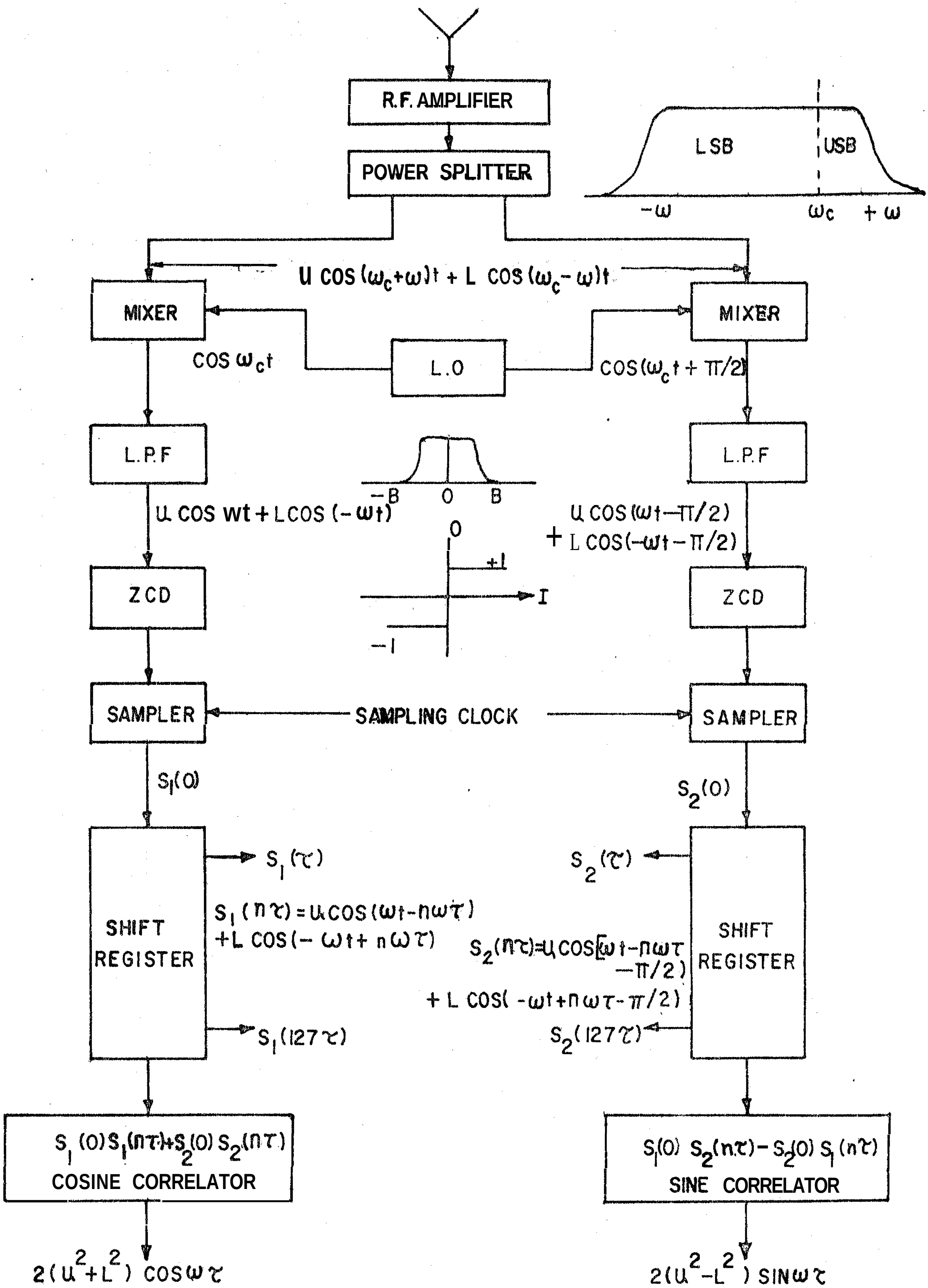


FIG. 2·24 BLOCK DIAGRAM OF THE LINE RECEIVER.

carbon in the direction of Cas A, the T_L / T_c is of the order of ,001. Thus it is necessary to determine the frequency response of the system to accuracies better than 1 part in 10,000 (for 10σ detection). A one-bit correlator is insensitive to absolute gain variations, but drifts in the bandpass shape anywhere in the system are important. The time scale for such drifts range from msecs to hours. They depend on the details of the design and on the fraction of the front end bandwidth used for observation. In order to keep track of the bandpass variations, one has to measure the difference between the on line and the off line spectrum. A DSB system measures this without resorting to switching schemes such as load switching, frequency switching or beam switching. Some of the disadvantages of these schemes when compared to the DSB system are:

1. Observing time has to be equally split between on line and off line measurements.

2. Load switching can cause changes in the band pass characteristics and produce curved baselines due to

a) change in the receiver level produced due to the temperatures of the load and source being unequal, and

b) reflections due to an impedance mismatch.

3. Frequency switching requires a fast settling local oscillator system which is expensive.

4. Beam switching is limited to sources of small angular extent and excludes the study of extended sources.

5. The RF hardware for SSB generation is more complex.

The main disadvantages of the DSB system are

1. In load and beam switching, lab generated "spectral-lines" do not appear in the final spectrum as they will be present at the same place both during on and off line measurements.

2. A DSB system requires four correlators while SSB system needs only two for the same SNR. This is not a serious disadvantage since the receiver uses one-bit correlators, which are inexpensive.

For a Bandwidth B with a 128 delay Shift Register, a DSB system measures 512 correlations and gives 256 points in the entire band (128 corresponding to U + L and 128 corresponding U - L). An SSB system needs only 256 correlators to produce 256 points in a bandwidth B. The above statements are true only for single dish observations. When one needs to take a cross spectra (say between E-W and N-S), even an SSB system will need 512 correlators (256 for cosine correlation and 256 for sine correlation).

For spectral line observations, U-L is more important. U+L is required only to equalise the gains of each frequency channel. As mentioned earlier, U-L where the spectral line appears is of the order of 0.001 of U+L. It may be enough to spend only a fraction of total observing time to measure U+L. The time to be spent and the switching frequency depends on the time scales over which the bandpass changes appreciably. This enables one to use all the correlators to measure U-L and hence get more resolution. This reduces the number of correlators required for a DSB system to the same as that of an SSB system.

This thesis describes line observations made in the direction of Cas A, the Galactic centre and Cygnus. The C574α and C575α lines were detected in the direction of CAs A.