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A new electronically steerable 1056 dipole array at 327 MHz for the Ooty radio telescope

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Abstract. An electronically steerable 1056-element linear dipole array having a loss of $\lesssim$1.5dB at 327 MHz installed on the 530m long and 30m wide parabolic cylindrical Ooty radio telescope is described. The antenna beam is steered in the north-south direction by suitably phasing the array. The colinear dipole array is divided into 22 subarrays, called modules, of 48 dipole each. A 4-bit low loss ($\sim$0.6dB) PIN diode phase-shifter follows each dipole for phasing the array. The phased signals from all the dipoles are combined using 'Christmas-tree' type branching system with 3dB power combiners and 4.77 dB directional couplers. The phase-shifters are controlled by a computer and transistor drivers. To protect the PIN diodes against electrical transients, Zener diode surge suppressors are put in the diode driver circuitry placed near the phase-shifters. The rms variation in phase and amplitude at any dipole in the array is about 6° and 0.3dB respectively. The maximum VSWR at the input of any dipole is $\lesssim$ 2 over a scan range of $\pm$40° and the frequency range of $\pm$6 MHz. The completed array has been in satisfactory operation for the past five years.

Key words : radio telescope—phased array

1. Introduction

The Ooty radio telescope (ORT) consists of an asymmetric parabolic cylindrical antenna, 530m long in north south (NS) and 30m wide in east west (EW) (Swarup *et al.* 1971). The antenna surface is formed by stainless steel wires running parallel to the axis of the cylinder. It is illuminated by an array of dipoles located along the focal line. The telescope beam is steered in EW (hour-angle) direction by mechanical rotation of the antenna around its long axis. For steering the beam in NS (declination) direction the whole array is divided into 22 sub-arrays, called modules. The module beam is scanned by phasing each dipole at RF and the telescope beam is scanned by introducing appropriate time delays at IF before combining the signals of all the modules.

The telescope was primarily designed for observations of radio sources at 327 MHz using the moon as an occulting disc (lunar occultation technique) and therefore the declination scan range of $\pm 36°$ achieved earlier was adequate. In the original telescope feed, which was developed in the mid sixties (Kapahi *et al.* 1975), the dipoles in each module were connected to a main line through directional couplers with trombone type mechanical phase-shifters inserted between each coupler for scanning the telescope beam in declination. These mechanical phase-shifters suffered from back lash errors, poor reliability and slow scanning rate. With the availability of low loss PIN diodes and need for a wider declination coverage required for the Ooty synthesis radio telescope (Sukumar *et al.* 1988), it become imperative that a new feed system be designed which will have wider declination scan range, larger bandwidth and capability of rapidly switching the beam in declination.

In this paper we describe a new electronically steerable dipole array installed on the Ooty radio telescope. The array consists of 1056 dipoles and employs a low loss ($\sim$ 0.6dB) 4 bit PIN diode phase-shifter, after each dipole, for steering the telescope beam in declination.

2. The corner reflector

The parabolic surface of the telescope subtends an angle of about 85° at the focal line. This requires that the H plane pattern of the feed element fall down to $\gtrsim$10dB at about $\pm 42°$ so that spillover is minimized. Also the E plane pattern should be broad enough for achieving adequate declination scan range. Therefore, based on the experimental work of Cottony & Wilson (1960) and Kapahi *et al.* (1975), dipole in a 90° corner reflector with one side of the corner reflector $\doteq 1.17\lambda$ (λ = RF wavelength) was chosen as the feed element for the antenna. In the earlier feed design (Kapahi *et al.* 1975) the E plane scan range was limited to $\pm 36°$. As we desired a wider E plane half power beam width (HPBW), fresh set of measurements were made to optimize the corner reflector parameters. A $\pm 40°$ E plane HPBW was achieved with a dipole to corner spacing of $S = 0.608\lambda$ at a cost of a slight dip in the E plane pattern as shown in figure 1. It is known that wider half power E plane beam width can be achieved by increasing the value of S but two constraints, namely that (i) the dip in the E plane pattern be $<$ 1dB, and (ii) the 10dB H plane pattern be approximately $\pm 40°$ restricted S to 0.608λ. The E and H plane patterns achieved for $S = 0.608$ are shown in figure 1. The resulting antenna aperture illumination in EW direction is shown in figure 2.

It is essential to know the phase centre of a dipole in a corner reflector for correctly mounting the array along the focal line of ORT. The phase centre was experimentally determined by means of a set up consisting of a corner reflector with a dipole mounted on a rotatable wooden structure and a receiver placed on a wooden platform held between two long wooden poles vertically above the corner

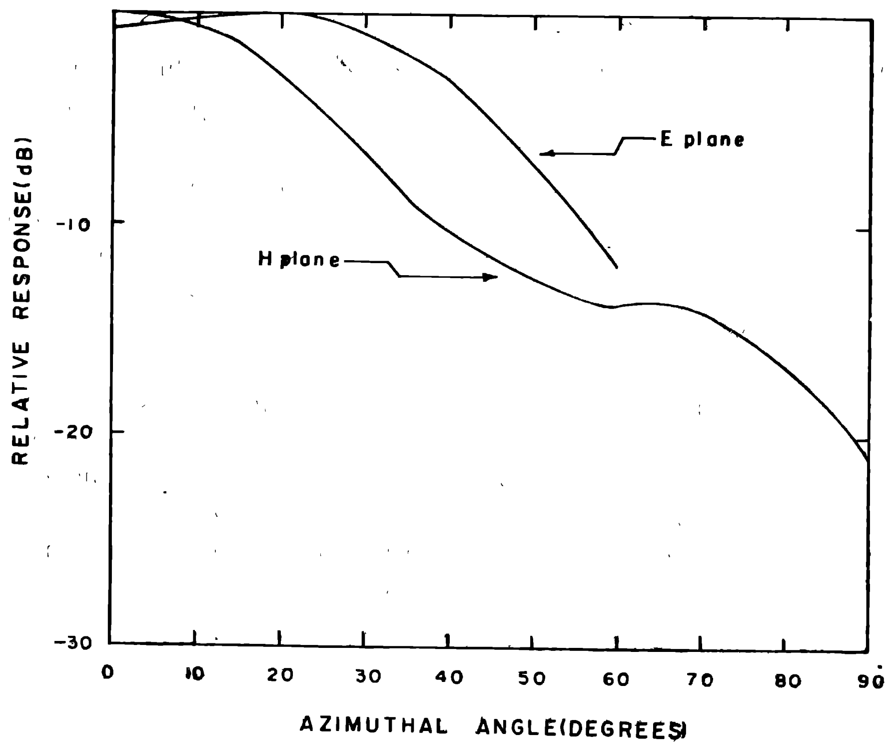

Figure 1. The E and H plane patterns of the corner reflector for a dipole to corner spacing of 0.608λ.

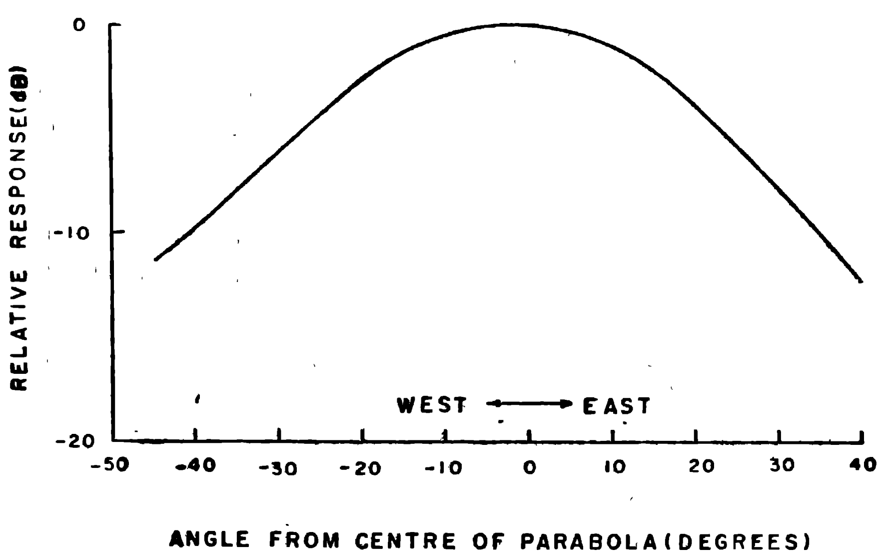

Figure 2. The EW aperture illumination pattern for ORT.

reflector at a height of about 10m. The dipole was made to radiate at 327 MHz and the phase changes in the received signal with the corner reflector at different angular positions were measured. Knowing the geometry of the set up and the phase changes, the position of the phase centre from the apex of the corner reflector was calculated to be $0.36 \pm 0.02\lambda$.

3. The combining network

In the earlier feed system of ORT (Kapahi *et al.* 1975) the modules were subdivided with two submodules each containing 22 dipoles. These dipoles were connected to a main line of 108ohm characteristic impedance through 22 directional couplers of appropriate coupling. The mechanical phase-shifters were inserted in the main line between each pair of dipoles. In this series type of combining network the losses due to the 22 phase-shifters all add up and hence it cannot be used for the present array as the diode phase-shifters have a loss of $\sim$ 0.6dB each. Also in the earlier array malfunctioning of any one phase-shifters affected all those preceding it. Therefore, we chose a parallel type of combining network in which the diode phase-shifters are placed between the dipoles and the main combining network as shown in figure 3. A Christmas tree branching type of network was chosen for combining the signals to increase the achievable bandwidth. The power combiners are of split-tee type consisting of two quarterwave lines of characteristic impedance of $Z_0\sqrt{2}$ and a resistor of $2Z_0$ ohms as shown schematically in figure 3. The measured isolation between the two ports is nearly 23dB. One group of 16 dipoles and other of 8 dipoles of a submodule are then finally combined using a 4.77dB directional coupler.

In the present array the number of dipoles per submodule has been increased to 24 from 22 ; Inter-element spacing of 0.522λ was chosen considering that (i) the length of the dipole is greater than 0.5λ, (ii) the length of the submodule is 11.5m and requiring that (iii) no grating lobe be present in the scan range of $\pm$ 60°. As shown in figure 4 the combining network is made of a six compartment aluminium extruded channel in which the larger compartment is used for mounting the diode phase-shifter blocks (see below) and the remaining for the coaxial lines of $Z_0 = 68.5\ \Omega$ and the power combiners. Copper tubes supported by

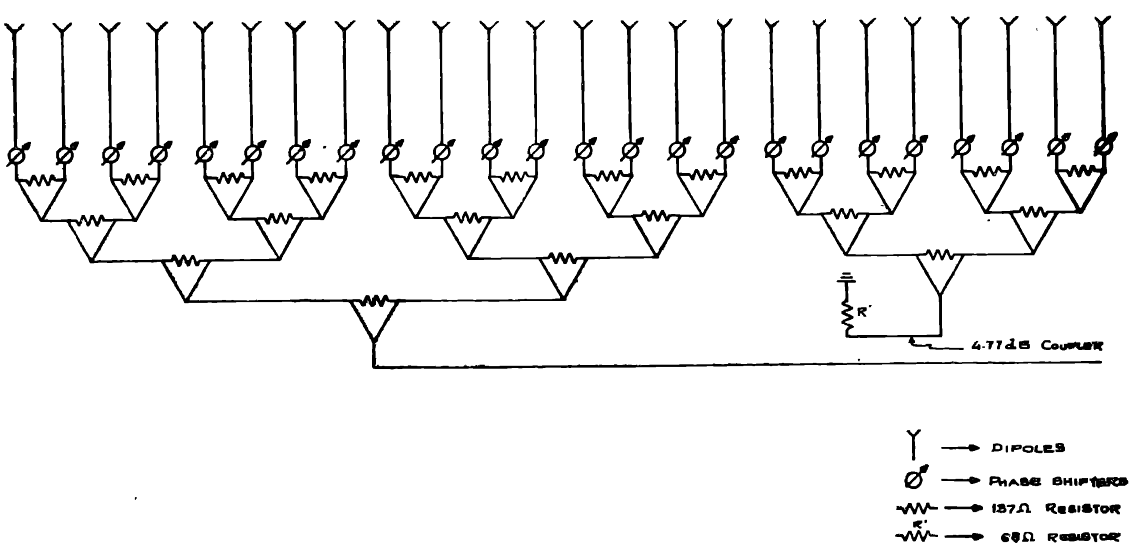

Figure 3. Schematic showing the Christmas-tree branching type network used for combining the signals from the 24 dipoles of a submodule.

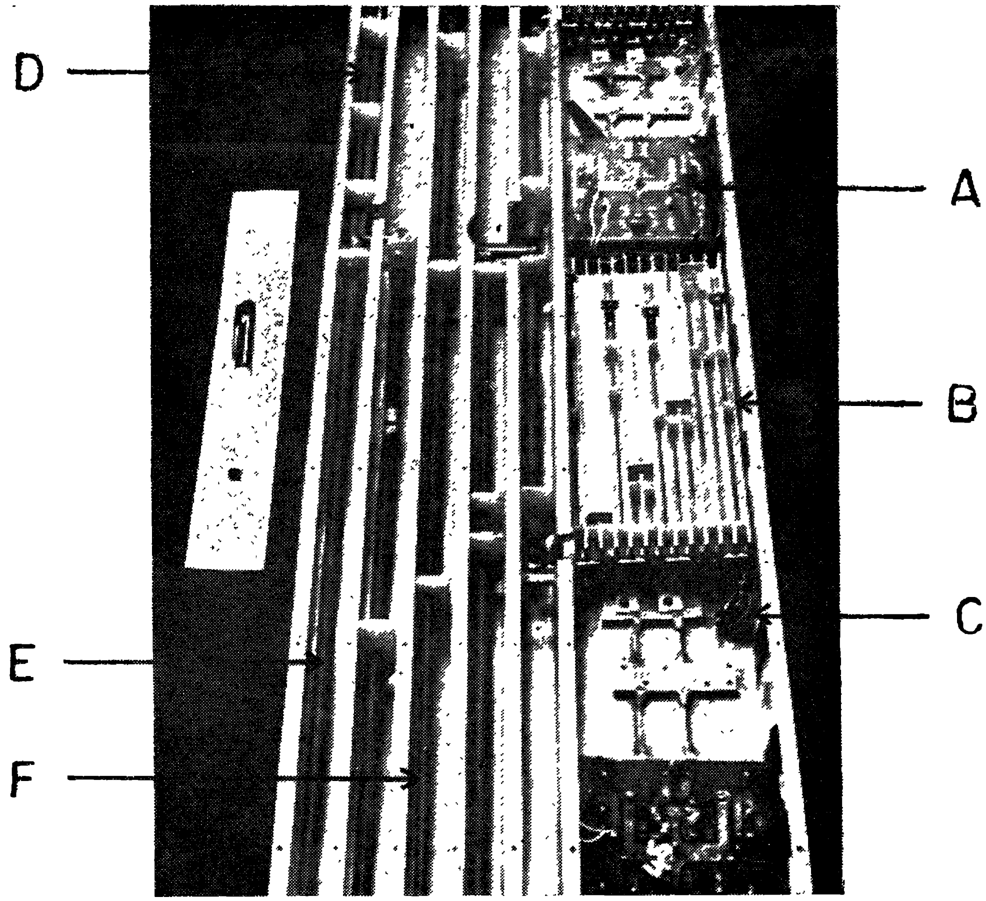

Figure 4. Photograph of a section of the aluminium extruded channel showing A — diode driver circuitry, B — diode phase-shifter block, C — output from dipole, D — the 4.77 dB directional coupler, E — 68 ohm coaxial line and F — split-tee type power combiner.

low-loss polypropeline spacers form the central conductors. Limitation of size forced us to use a structure that contained a number of bends from which reflections could occur but they were matched out by means of sleeves placed at appropriate places in the coaxial lines.

4. Mutual coupling

An important parameter of an array is the mutual coupling between its elements because of which the active impedance of the array varies as a function of scan angle. It is therefore essential to measure it and correct it by, for example, suitably modifying the matching transformers of the elements. Mutual coupling can be modelled as a n-port network described by $n \times n$ $[S]$ matrix where n is the number of elements in the array. The $[S]$ matrix is chosen primarily because the equivalent driving sources exciting the elements are of constant power type, i.e., for an infinite array, the power fed to each element is equal, although voltages and currents need not be. In order to measure the individual terms of the matrix, it is essential that the elements be of minimum scattering type, i.e., they must become 'invisible' when terminated by the characteristic impedance. Since this is true at metre wavelengths the active impedance of the elements as a function of scan angle can be determined by knowing the [*S*] matrix. Since mutual coupling falls off rapidly with distance, it is sufficient to consider only about five elements on either side of the element whose impedance is to be determined. An alternative approach for determining the mutual coupling it to directly measure the dipole impedance, *in situ*, as a function of scan angle, which was also done by us and the results are shown in figure 5.

Once the variation in the active impedance of the elements is known it is necessary to determine the optimum value of the impedance to which the elements are to be matched so as to minimize gain variation over the scan angle. This implies that certain amount of mismatch occurs at some or all the scan angles, which leads to multiple ghost responses at different scan angles. In our case, however, they are considerably reduced because of the time delay introduced in the IF channels of each module. The relation between the response of an isolated

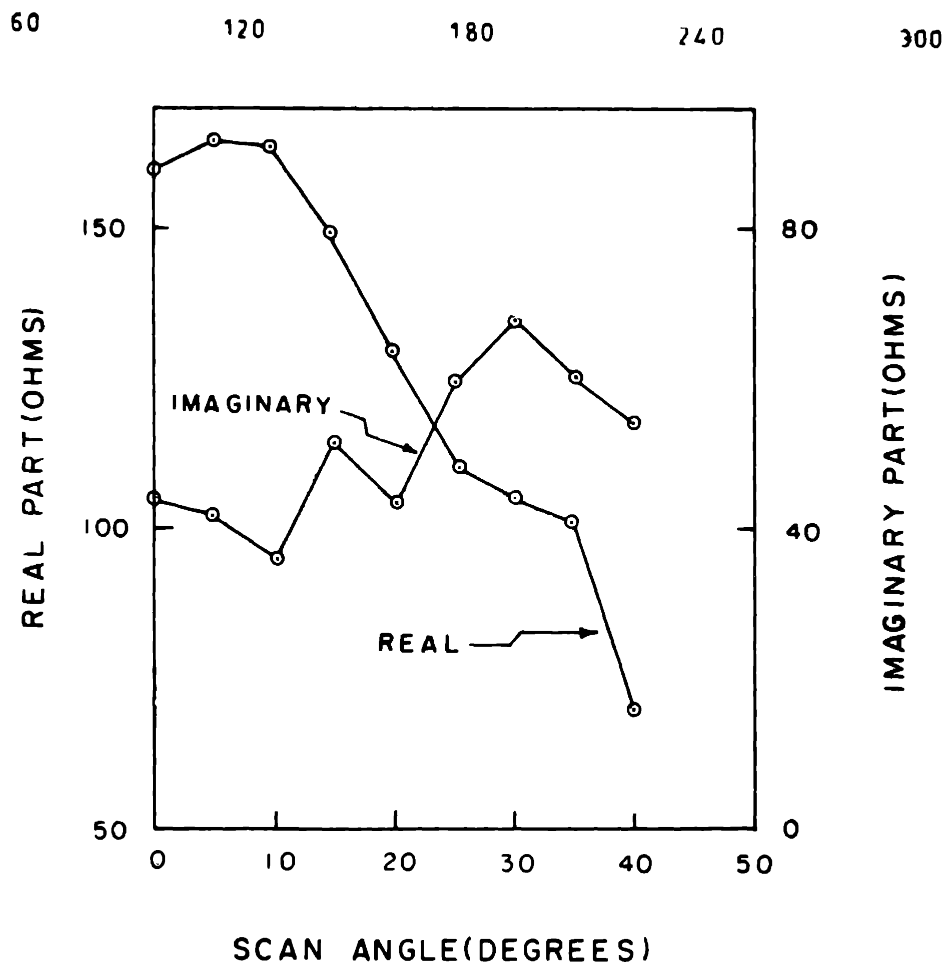

Figure 5. Dipole impedance as a function of scan angle used for calculating mutual coupling.

element and that of an element in an array including the effect of reflections due to mutual coupling is given by Allen (1964), which is

$$g_r(\delta) = g_i(\delta) \cdot \frac{R_i}{R(\delta_0)} (1 - |\Gamma(\delta)|^2)$$

where $g_r(\delta)$ and $g_i(\delta)$ are the responses due to array and isolated element respectively, as a function of the scan angle δ (which is declination in our case), R_i is the real part of the isolated dipole impedance, $R(\delta_0)$ is the real part of the active impedance at δ_0 at which the array is matched and $\Gamma(\delta)$ is the reflection coefficient as a function of scan angle. It is seen that $g_r(\delta) = g_i(\delta)$ only where $\Gamma(\delta)$ is zero for all scan angles. Otherwise, $g_r(\delta) < g_i(\delta)$.

Consider an array matched at $\delta_0 > 0$. Since the response of the isolated element tends to fall off as δ increases, it is clear that in the range $0 < \delta < \delta_0$, the element response gets flattened though at the expense of setting up large reflections, resulting in gain reduction. For us it was important to reduce the reflections near the broadside ($\delta = 0°$). Hence, we partially matched the measured active impedance at $\delta = +20°$ and $-20°$ subject to the constraint that at $\delta = 0$, the VSWR did not exceed 1.4. The results are shown in figure 6.

5. The diode phase-shifters

A reliable low loss phase-shifter is one of the basic requirements of a phased array. For a N-bit phase-shifter the loss in sensitivity due to quantization is given by

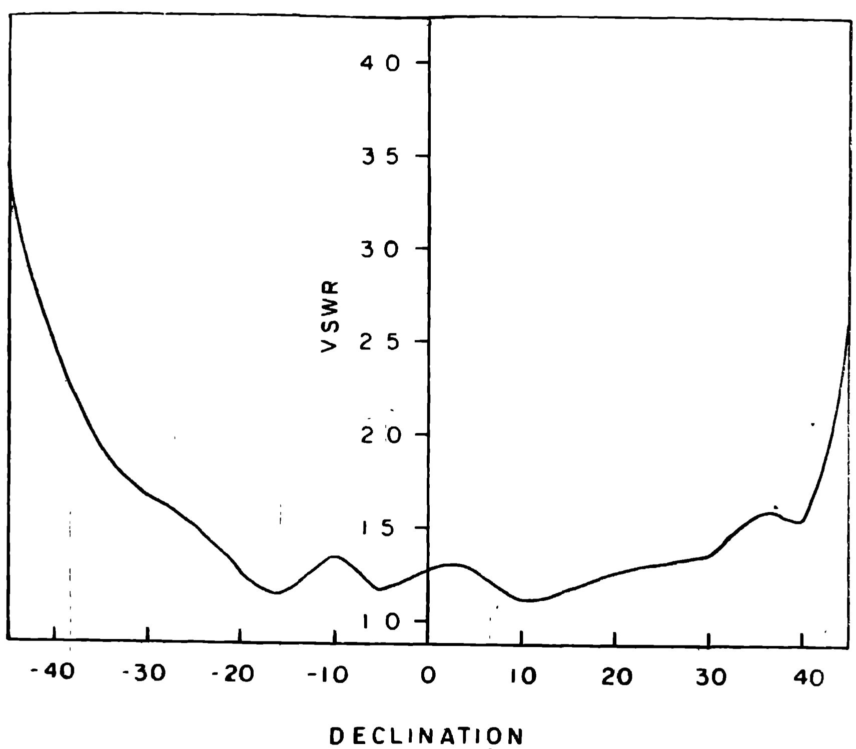

Figure 6. VSWR of a dipole as a function of scan angle after partially matching it to take care of mutual coupling.

$1/(1-\epsilon^2)$ where ϵ is the rms phase error across the array. The error in beam position is given by (HPBW of the array)/2^N. Therefor, it is easily seen that a 4-bit phase-shifter consisting of 180, 90, 45 and 22.5 degrees as the four suitable phase-shifters is adequate for our purpose. The configuration of the phase-shifters developed for this purpose is that of a switched line type, a schematic sketch of which is shown in figure. 7. Its chasis is made up of an aluminium block cut from an aluminium extruded channel (figure 4). There are eleven channels on one face and twelve channels on the other, each of 6mm width and 14mm height to accommodate the four phase lines each taking two channels, eight quarterwave chokes for feeding the voltages for switching the eight pairs of diodes and five quarterwave grounded chokes. The lengths of the eight chokes feeding the voltages were determined by experimentation to get good VSWR. The lengths of the grounded chokes are adjustable and were used for final tuning of the phase-shifters to give VSWR $\lesssim$ 1.5 for any of the 16 possible positions of the switches. The total insertion loss of a phase shifter is equal to the sum of the losses due to the coaxial lines and those due to dissipation in the diodes. The line losses were minimized by using air as dielectric. Losses due to diodes are given by $R_d/(2Z_0)$ where R_d is the dynamic resistance of the diode at 327 MHz and Z_0 is the characteristic impedance of the lines. Z_0 cannot be increased beyond a certain limit without making the lines mechanically unstable and without increasing the line losses. A

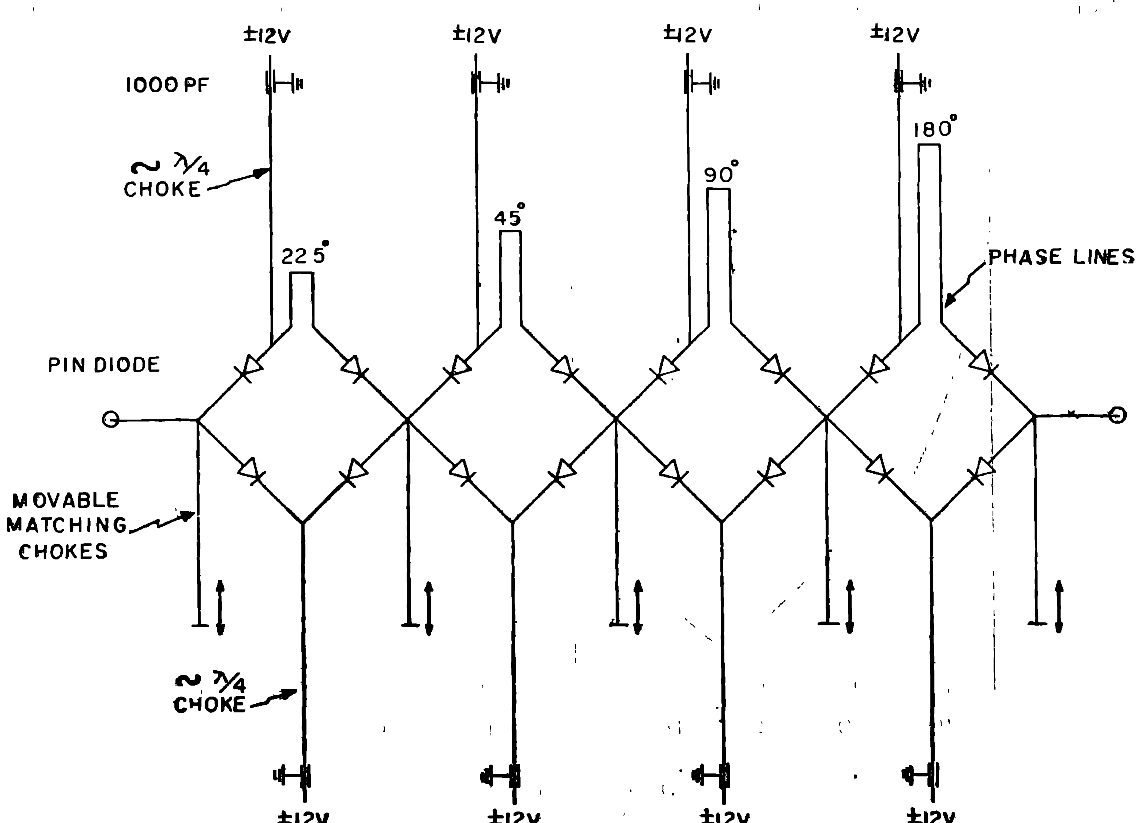

Figure 7. Schematic sketch of the 4-bit diode phase-shifter.

Z_0 of 68.5 ohm was chosen from practical considerations. Hewlett Packard PIN diode 5082–3188 was chosen because of its low R_D of $\sim$ 0.35 ohm and its low cost. A total of 16,896 of these diodes are used in the array. The total insertion loss of a phase-shifter varies between 0.5dB and 0.8dB as the diodes are switched. Figure 8 shows the variations in loss and VSWR of a typical phase-shifter unit with the 16 possible positions of the 4 bits.

6. Phase-shifter control

Since the phasing of each module is identical, phase-shifters can be set by a set of 192 voltages. For the purpose of phase-shifter setting, the telescope is divided into two halves. Each bit of a corresponding dipole in each module is paralleled and is driven by transistor drivers located at the control huts situated at the centres of north half and south half of the telescope. The transistor drivers can be controlled from the telescope control room either manually or by computer. In computer mode, for a given declination, the computer calculates the required phase-shift with quantization and encodes these in strings of ten bits. Phase setting information in strings of ten bits data along with five bit address and a parity bit are then transmitted using multiple core cables to the two control huts, one for the north half and the other for the south half of the telescope. These TTL level signals, received at the two control huts, are converted to $\pm$ 12 volt level signals using transistor driver circuits and are used to control the phase-shifters through diode biasing circuitry located near each phase-shifter. The transistor driver output voltage of $\pm$12V was determined by taking into account wire losses and the minimum voltage of $\pm$10 volt

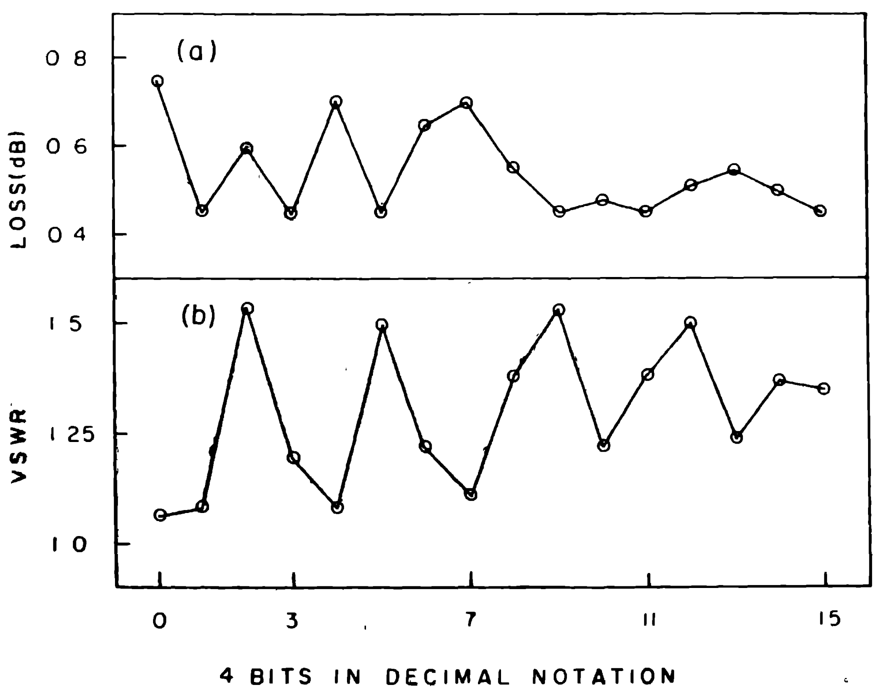

Figure 8. Variations in (a) loss and (b) VSWR of a typical diode phase-shifters for the 16 possible positions of the 4 bits.

required to control the phase-shifter diodes. Since the whole telescope and the phase-shifter structure are made up of steel and aluminum, any large surges due to welding or lightning etc. can induce large surge voltages across the phase shifter diodes and damage them. In order to protect the diodes, two back-to-back Zeners, with response time $< \pm 1$ μs, have been used along with the diode biasing circuitry to protect the phase-shifter diodes from large surges. Also the 192 wires carrying the voltages to the diodes are run along the focal line inside the aluminium channels containing the phase-shifters and the combining network, thus shielding them from lightning.

7. Performance

The new array has been in operation since 1982. It has improved the sensitivity of the telescope. Soon after the installation of the new array, the signal to noise ratio was about 10 per Jy for a phase switched system having system bandwidth of 4 MHz and a time constant of 1 s. Measurements on the 44 submodules showed that VSWR at their outputs was generally $\lesssim 1.4$ within the declination range of $\pm 45°$ of the array except over a small range near $\delta = 0°$ where it built up to about 1.5 as shown in figure 9 for one of the submodules. In figure 10 are shown the relative phases and absolute losses (measured loss $-10 \log (1/24)$) at the 24 output ports of a typical submodule. About two years after the installation of the new array, relative amplitude and phase at each of the 1056 dipoles were measured *in situ* using a dipole probe and vector-voltmeter. The scatter in phase was found to be about 6° rms and that in amplitude within 1dB of the expected value. About 100 phase-shifters had to be replaced as they were found to have developed high losses. As of now, the sensitivity of ORT as defined above has slightly gone down to a signal to noise ratio of about 8 per Jy. Also we find that once in a month or so, one of the diodes suddendly goes to Zener mode giving high noise and hence a high second detector current of the module concerned which is

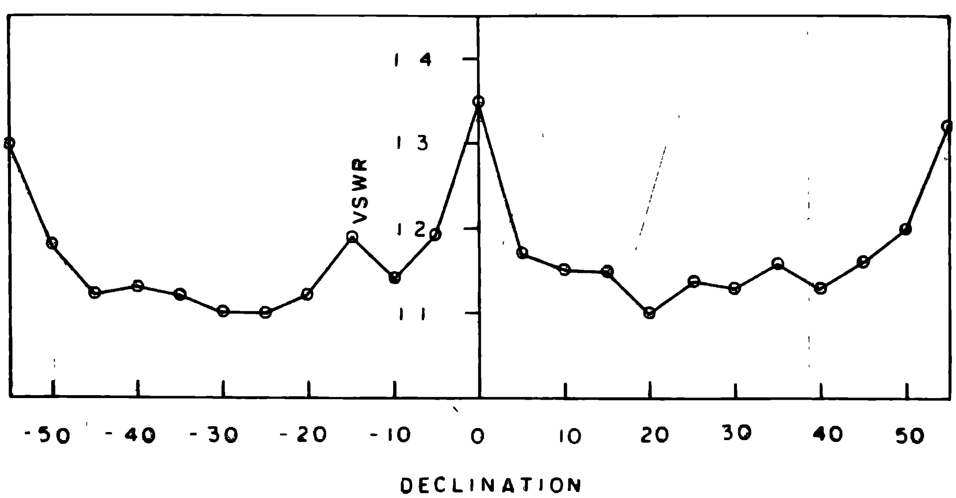

Figure 9. VSWR at the output of a typical submodule as a function of declination.

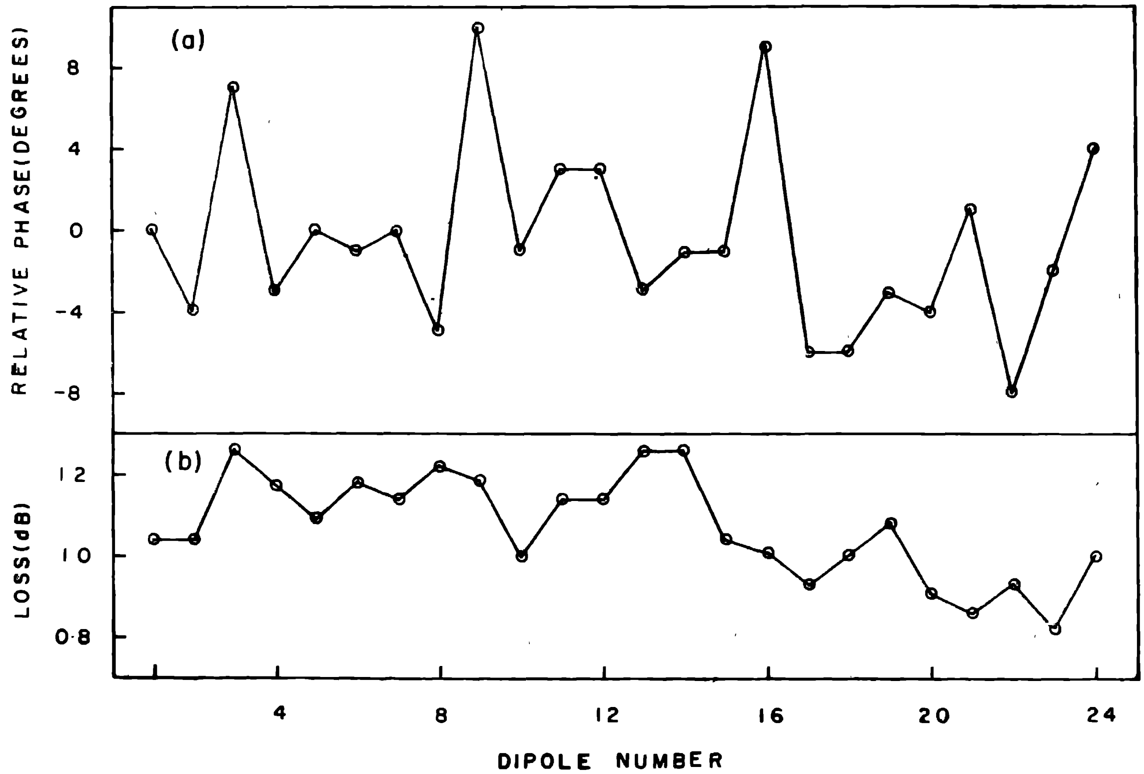

Figure 10. (a) Relative phases and (b) absolute losses (measured loss —10 log (1/24)) at the 24 outputs of a typical submodule for 0 degree declination.

routinely monitored. The defective phase-shifter can be easily traced within about 10 minutes by manipulating the switches meant for manual control of declination.

Acknowledgements

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