Modification of Large Reflector Antennas for Low Frequency Operation

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> Doctor of Philosophy in Electrical Engineering

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Abstract

Modifications of large reflector antennas, such that their observing capabilities are enhanced in the range of about 10–500 MHz without affecting operation of the pre–existing higher–frequency systems, are addressed in this dissertation. The major contributions of this dissertation can be divided into two parts: 1) designing new low frequency feeds, and 2) developing new analysis methodologies which, as opposed to traditional techniques, are suitable for analyzing low frequency systems.

First, we consider the performance of existing schemes that provide low frequency capability. Then, a new class of dipole–based low frequency feeds – namely, the "distributed feed array" – is designed to cover the frequency range of interest without affecting operation at higher frequencies. As an example, distributed feed arrays are designed for the Expanded Very Large Array (EVLA) to cover the range of 50–250 MHz. A method of moments (MoM) model of an EVLA antenna is developed for this purpose. The new design shows performance comparable to the existing 4 m system on the EVLA in the range of 50–88 MHz, and introduces observing capabilities in the range of 110–250 MHz (currently not covered by the EVLA). Moreover, the blockage presented to the existing EVLA L-band system is reduced significantly when the existing 4 m system is replaced by the proposed system.

At low frequencies, external noise can be a significant or dominant contribution to the total noise of the system. This, combined with mutual coupling between the array elements of the distributed feed array, makes it difficult to predict the sensitivity of these systems. This dissertation describes a system model and procedure for estimating the system equivalent flux density (SEFD) – a useful and meaningful metric of the sensitivity of a radio telescope – that accounts for these issues.

We consider the efficiency of methods other than MoM – in particular, Physical Optics (PO), Uniform Geometrical Theory of Diffraction (UTD), and hybrid methods – for accelerated computation at low frequencies. A method for estimating the blockage presented by low frequency systems to the pre–existing higher–frequency systems is also described. To Anindya S. Harun, the apple of my eye

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Contents

A	bstra	nct	1		
A	cknov	wledgements	iv		
1	Introduction				
	1.1	State of the Art	5		
	1.2	Current Trends in Low Frequency Feed Development	11		
	1.3	Contributions	16		
	1.4	Organization	18		
2	Elec	ctromagnetic Analysis of Large Reflector Antennas at Low Frequencies	19		
	2.1	Physical Optics	20		
	2.2	Uniform Geometrical Theory of Diffraction	22		
	2.3	Method of Moments	27		
	2.4	Comparison of MoM, PO and UTD at Low Frequencies	33		
	2.5	MoM Model of an EVLA Reflector Antenna	35		
	2.6	Summary	43		
3	An	alysis of Sensitivity of Large Reflector Antennas at Low Frequencies	50		
	3.1	System Model and Sensitivity Analysis of a Reflector Antenna System Con-			
		sisting of a Single Feed Element	51		
	3.2	Example: Analysis of Sensitivity of a Reflector Antenna Fed with a Single			
		Feed Element	56		
	3.3	Prospects for Broad Low Frequency Coverage Using Single Dipole Feeds	57		
	3.4	Analysis of Sensitivity of a Reflector Antenna Fed with an Array Feed	63		
	3.5	Comparison of the Methods of Sections 3.1 and 3.4 for Analysis of a Single			
		Dipole Feed System	68		

	3.6	Example: Analysis of Sensitivity of a Reflector Antenna Fed with an Array	70
	3.7	Summary	70 72
4	Ana	alysis of Existing Low Frequency Systems on the EVLA	74
	4.1	Existing 4 m System on the EVLA	74
	4.2	Existing 90 cm System on the EVLA	78
		4.2.1 Pattern and Effective Aperture	79
		4.2.2 Sensitivity and Bandwidth	82
	4.3	Summary	88
5	Foc	us– or Near Focus–Fed Systems	90
	5.1	Dipole Feeds Located at the Focus of a Simple Reflector Antenna $\ . \ . \ .$.	91
	5.2	Application Example: Dipole Feeds Located near the Focus of an EVLA	
		Antenna	92
	5.3	Dual Length Dipoles for the EVLA	94
	5.4	Summary	98
6	Dis	tributed Feed Array	100
	6.1	High Frequency Safe Areas in Front of a Main Reflector	102
	6.2	Distributed Feed Arrays on a Simple Reflector Antenna	102
		6.2.1 Dipole Arrays in the Focal Plane of a Simple Reflector Antenna \ldots	104
		6.2.2 Dipole Arrays at the Border of Regions 1 and 2 of a Simple Reflector	
		Antenna	107
		6.2.3 Summary	114
	6.3	Application Example: Distributed Feed Arrays on an EVLA Antenna	116
		6.3.1 Strut–Straddling Dipoles in the Plane Containing the EVLA Prime	
		Focus	117
		6.3.2 Strut-Straddling Arrays in Planes Parallel to the $z = 0$ m Plane on	
		the EVLA	119
		6.3.3 Double Ring of Dual Length Dipoles for the EVLA	124
	6.4	Summary	124
7	\mathbf{Ass}	essment of Blockage due to Low Frequency Feeds on the EVLA	128
	7.1	Assessment of Blockage due to the Existing EVLA 4 m System	129
		7.1.1 Assessment of Blockage at 1.4 GHz due to the EVLA 4 m System	129

	7.1.2 Effects of the EVLA 4 m System on the Performance of the Existing					
			90 cm System	133		
7.2 Assessment of Blockage due to the Proposed Strut–Straddling System			134			
		7.2.1	Assessment of Blockage at 1.4 GHz due to the Proposed Strut–Straddlin	g		
			System	135		
		7.2.2	Effects of the Proposed Strut–Straddling System on the Performance			
			of the Existing 90 cm System	136		
	7.3	Summ	ary	137		
8	Con	clusio	ns	138		
	8.1	Findin	gs	138		
	8.2	Future	e Work	141		
A	Con	nputat	ion of the UTD Transition Function	143		
в	3 Determination of Wire Grid Model Parameters for Method of Moments					
	Moo	deling	of a Large Reflector Antenna	146		
С	EVI	LA Sul	breflector Model	150		
D	Calo	culatio	n of Antenna Temperatures for Different Pointing Directions	153		
\mathbf{E}	Copyright Permission 157					
Re	efere	nces		161		

List of Figures

1.1	An EVLA reflector antenna. Image source: NRAO/AUI and Patrick A. Lofy,	
	Photographer (used with permission, see Appendix E)	5
1.2	Close up of feeds and struts on an EVLA reflector antenna. Courtesy of	
	Lincoln Greenhill, Harvard/SAO (used with permission, see Appendix E)	6
1.3	Low frequency feeds on an EVLA reflector antenna. Only the 90 cm feed is	
	permanent. The 153 cm feed was experimental and never adopted (see text).	
	The 4 m feed is only intermittently used. Courtesy of Lincoln Greenhill,	
	Harvard/SAO (used with permission, see Appendix E)	7
1.4	A GMRT reflector antenna. Courtesy: Suman Majumdar, IIT Kharagpur,	
	India (used with permission, see Appendix E)	9
1.5	The GMRT 50 MHz feed [19]. The left image shows the folded V–dipoles in a	
	boxing ring configuration. The right image shows its mounting on the turret	
	of a GMRT antenna (used with permission, see Appendix E)	10
1.6	The 150 MHz feed on a WSRT reflector antenna [21]. The left image shows	
	the position of the feed when it is in operation and the right image shows the	
	stowed position of the feed (used with permission, see Appendix E)	11
1.7	The optimized Eleven antenna for the GMRT $[25]$ (used with permission, see	
	Appendix E). \ldots	13
1.8	The concept of a reflector IRA.	14
1.9	Distribution of power density in the focal plane of a reflecting paraboloid ($D =$	
	25 m, $f/D = 0.36$) relative to the power density at the focus in response to	
	plane wave illumination at (a) 500 MHz, (b) 50 MHz	16
2.1	Geometry for UTD analysis of an axisymmetric paraboloidal reflector [33].	23
2.2	Illustrating the plane passing through the edge of the paraboloid [33].	24
2.3	Geometry of incident and diffracted ray at Q_1 [33]	25
2.4	Geometry of incident and diffracted ray at Q_2 [33]	26
2.5	Wire grid model of a 26 m-diameter reflector surface.	33

2.6	E-plane pattern at 88 MHz using MoM, PO and UTD	35
2.7	H-plane pattern at 88 MHz using MoM, PO and UTD	35
2.8	E-plane pattern at 74 MHz using MoM, PO and UTD	36
2.9	H-plane pattern at 74 MHz using MoM, PO and UTD	36
2.10	E-plane pattern at 38 MHz using MoM, PO and UTD	37
2.11	H-plane pattern at 38 MHz using MoM, PO and UTD	37
2.12	E-plane pattern at 20 MHz using MoM, PO and UTD	37
2.13	H-plane pattern at 20 MHz using MoM, PO and UTD	37
2.14	E-plane pattern at 10 MHz using MoM, PO and UTD	38
2.15	H-plane pattern at 10 MHz using MoM, PO and UTD	38
2.16	Directivity of a 26 m-diameter reflector antenna, fed from the focus with a	
	half–wave dipole, computed using PO and MoM.	39
2.17	Detail of the EVLA reflector geometry, showing the subreflector, the subreflec-	
	tor mount, the struts, the support wires, and the feed assembly. The drawing	
	is provided by Jim Ruff of NRAO	40
2.18	Wire grid model of an EVLA subreflector.	41
2.19	Wire grid model of the subreflector mount	42
2.20	Wire grid model of an EVLA antenna including main reflector, subreflector,	
	struts, support wires, mount, and feed assembly	43
2.21	E-plane co–pol pattern at 29 MHz for different configurations	45
2.22	E-plane co–pol pattern at 38 MHz for different configurations	46
2.23	E-plane co–pol pattern at 74 MHz for different configurations	47
2.24	E-plane co–pol pattern at 88 MHz for different configurations	48
3.1	System model of the signal path from the incident signal through the receiver.	52
3.2	Results of sensitivity analysis for a 25 m–diameter reflector antenna pointing	
	toward Sagittarius. Note: HWDP stands for half–wave dipole	58
3.3	Results of sensitivity analysis for a 25 m–diameter reflector antenna pointing	
	toward Virgo. Note: HWDP stands for half–wave dipole	59
3.4	Comparing the power density at the dipole terminals due to external noise	
	and internal noise.	61
3.5	SEFD for a 25 m-diameter reflector antenna system fed with a 74 MHz reso-	
	nant dipole from focus.	69
3.6	SEFD for a 25 m–diameter reflector antenna system, fed with an array of	
	74 MHz resonant dipoles employed around the focus in the focal plane	71

4.1	Feed dipole located at the image of the prime focus with respect to the sub-	
	reflector. Tilt of the subreflector with respect to the main reflector axis is	
	2.4° .	76
4.2	Orientation of the support wires in the EVLA antenna model with respect to	
	a single (a) co-aligned, and (b) cross-aligned dipole feed.	77
4.3	Calculated E–plane co–pol directivity of the EVLA 4 m system for (a) co-	
	aligned support wires, and (b) cross-aligned support wires	78
4.4	Estimated SEFD for the existing EVLA 4 m system. Also, shown in inset is	
	the degree to which external noise dominates the internal noise	79
4.5	SEFD for different noise conditions, and reduced effective aperture for the	
	EVLA 4 m system in the range 50–170 MHz.	80
4.6	Comparison of directivity calculated using the MoM method and the hybrid	
	method as a function of frequency	82
4.7	Calculated E-plane co-pol directivity of the EVLA 90 cm system for (a) co-	-
	aligned support wires and (b) cross-aligned support wires	83
4.8	Error in calculation of effective length's phase when the main reflector is not	00
1.0	considered	85
4.9	Error in calculation of IME when the main reflector is excluded from the MoM	00
1.0	analysis	86
4 10	Estimated SEFD for the existing EVLA 90 cm system. Also, shown in inset	00
1.10	is the degree to which external noise dominates the internal noise	89
	is the degree to which external holde dominated the internal holde	00
5.1	Estimated SEFD for a simple 25 m -diameter reflector antenna fed with 74 MHz -	
	, 150 MHz– and 327 MHz–resonant crossed dipole feeds located at the focus.	
	Note that the frequency range covered by each set of dipoles is identified. $.$	93
5.2	Estimated SEFD for the 74 MHz–, 150 MHz–, and 327 MHz–resonant dipoles $$	
	on the EVLA	95
5.3	The concept of dual length dipoles $\ldots \ldots \ldots$	96
5.4	SEFD for the dual length dipoles located at the image of the prime focus of	
	the EVLA. Also shown is the performance of 150 MHz–resonant dipoles of	
	radius 3.37 mm when they are located 0.50 m ($\lambda/4$ at 150 MHz) in front of	
	the subreflector, and when they are located at the image of the focus	97
5.5	Estimated SEFD for the dual length dipoles (74 MHz– and 150 MHz–resonant	
	dipoles), and the 327 MHz–resonant dipoles on the EVLA. (Compare to Fig-	
	ure 5.2)	98
0 1		101
0.1	An example of a distributed feed array	101

6.2	Regions of different power densities in the volume in front of a Cassegrain reflector antenna.	103
6.3	Feeds around the prime focus of a simple reflector in a boxing ring configura-	1.01
6.4	Power density in the focal plane of a simple 25 m-diameter reflector in response to a plane wave illumination at 74 MHz. PO and MoM are separately used	106
6.5	to do the analysis	107
6.6	refer to our choices for prospective locations (see text)	108
67	f/D = 0.36	110
6.8	Power density versus the distance from the focus along the border of Regions 1 and 2 in a boxing ring configuration Power density versus the distance from the focus along the border of Regions 1 and 2 for a simple reflecting paraboloid ($D = 25 \text{ m}$, $f/D = 0.36$), relative to the power density at the focus, in response to a plane wave illumination. L_1 ,	111
6.9	L_2 and L_s refer to our choices for prospective locations (see text) Estimated SEFD for single rings of 74 MHz-, 150 MHz- and 327 MHz- resonant dipoles at the border of Regions 1 and 2 of a simple reflector antenna	112
	with $D = 25$ m and $f/D = 0.36$	114
6.10	Comparison of performance between arrays of dipoles at the border of Re- gions 1 and 2 (SEFD _{co}), and arrays of dipoles in the focal plane (SEFD _{fc}) of a simple reflector antenna with $D = 25$ m and $f/D = 0.36$	115
6.11	Estimated SEFD for the double rings of 74 MHz–, 150 MHz– and 327 MHz– resonant dipoles at the border of Regions 1 and 2 of a simple reflector antenna	110
	with $D = 25$ m and $f/D = 0.36$	116
6.12	Estimated SEFD for the arrays of 74 MHz– and 150 MHz–resonant strut– straddling dipoles in the focal plane of an EVLA antenna	120
6 13	Location selection for the 74 MHz strut-straddling dipole array	120
6.14	Estimated SEFD for strut–straddling arrays of 74 MHz–resonant dipoles. Also shown for comparison is the estimated performance of the existing EVLA 4 m	14
	system	122
6.15	Double ring of strut-straddling dipoles.	123
6.16	Location selection for 150 MHz–resonant strut-straddling dipoles array	125

6.17	Estimated SEFD for strut–straddling arrays of 150 MHz–resonant dipoles on	
	the EVLA. Also shown for comparison is the estimated SEFD for 150 MHz– $$	
	resonant crossed dipole feeds located at the image of the prime focus	126
6.18	Double ring of "dual length" dipoles positioned at $z = -1$ m and $z = -2.25$ m	
	on an EVLA antenna. Also shown is the estimated performance when dual	
	length crossed dipoles are employed at the image of the prime focus of an	
	EVLA antenna.	127
7.1	Segments in the propagation path in a Cassegrain reflector antenna in transmit	
	mode	130
7.2	Directivity of an EVLA antenna at 1.4 GHz, in response to a dipole feed	
	located at the secondary focus, for different blocking conditions. The "No low	
	frequency dipoles" and "Double ring of strut–straddling dipoles" results are	
	essentially the same over most of this range	132
A.1	(a) Magnitude and (b) Phase of the UTD transition function	144
B.1	E-plane pattern for a half-wave dipole $\lambda/4$ above the wire grid ground for	
	different grid parameters. θ is the zenith angle	148
B.2	E-plane pattern for a half-wave dipole $\lambda/4$ above different ground models.	
	Grid parameters used for the disk model are $d = 0.1\lambda$, $s = 1$, $r = \lambda/75$, and	
	$N = 7272. \qquad \dots \qquad $	149
C.1	Comparison between the data points taken from the subreflector drawing and	
	our polynomial model	151

List of Tables

1.1	Comparing the low frequency systems on the existing reflector antennas of several radio telescopes. Comparison is based on a single reflector antenna	
	performance. Note that SEFD is calculated using Equation 1.1, assuming $e_I =$	
	1, and using values for A_e and T_{sus} extracted from the indicated references.	8
1.2	Comparing the performance of the distributed feed array, proposed in this	
	dissertation, with the existing low frequency system on an EVLA antenna	12
2.1	Summary of reflector antenna analysis using aperture theory and MoM	34
2.2	The lengths of the feed dipoles considered in Section 2.5.	44
2.3	Results of analysis for different configurations. Dir. refers to the directivity of the antenna [Note] The computation times reported are based on a 2 GHz	
	Pentium M processor and 1 GB RAM machine.]	49
3.1	Lengths of the feed dipoles in the study described in Section 3.2	57
3.2	Lengths and radii of the feed dipoles considered in Section 3.3	60
4.1	SEFD estimates for a simple 25 m-diameter reflector antenna, fed from the focus with a 74 MHz–resonant dipole, using the (direct) MoM method and	0.0
	the MoM+PO method	88
6.1	Prospective locations for employing dipole rings in the focal plane, expressed as the distance of the dipole centers from the focus. Also shown are the estimated performance of dipole rings employed at these locations and the	
	estimated performance of focus-fed systems	109
6.2	Prospective locations for employing dipole rings at the border of Regions 1	
	and 2 of a simple reflector antenna, expressed as the distance of the dipole	
	centers from the focus along r_d	113
6.3	Strut-straddling dipole arrays in the focal plane of an EVLA antenna	118

7.1	Comparison of blockage at 1.4 GHz presented by the image focus-located dipoles and the double ring of strut-straddling dipoles	134
A.1	Values used for interpolation for the computation of $F(X)$	145
C.1	The residual from the mathematical model of the subreflector shape. \ldots .	152
D.1	Antenna temperature and system temperature data for the reflector–only model, fed with a 29 MHz half–wave dipole, pointing toward Sagittarius	154
D.2	Antenna temperature and system temperature data for the reflector-only	154
D.3	Antenna temperature and system temperature data for the reflector-only	154
D (model, fed with a 38 MHz half–wave dipole, pointing toward Sagittarius	154
D.4	Antenna temperature and system temperature data for the reflector-only model, fed with a 38 MHz half-wave dipole, pointing toward Virgo	155
D.5	Antenna temperature and system temperature data for the reflector-only	
D.6	model, fed with a 47 MHz half-wave dipole, pointing toward Sagittarius Antenna temperature and system temperature data for the reflector-only	155
2.0	model, fed with a 47 MHz half–wave dipole, pointing toward Virgo	155
D.7	Antenna temperature and system temperature data for the reflector-only model fed with a 60 MHz half wave dipole, pointing toward Societarius	155
D.8	Antenna temperature and system temperature data for the reflector-only	100
D o	model, fed with a 60 MHz half–wave dipole, pointing toward Virgo	156
D.9	Antenna temperature and system temperature data for the reflector-only model, fed with a 74 MHz half-wave dipole, pointing toward Sagittarius.	156
D.10	Antenna temperature and system temperature data for the reflector-only	- •
	model, fed with a 74 MHz half–wave dipole, pointing toward Virgo	156

Chapter 1

Introduction

This dissertation considers modifications of large reflector antennas for operation in the range of 10 - 500 MHz. Radio astronomy at frequencies below 500 MHz is poorly explored despite important science being available in this frequency regime [1]. One reason behind this is the complications imposed by the Earth's ionosphere which becomes increasingly refractive as the observing frequency is decreased, and essentially becomes opaque below 10 MHz. Moreover, high-resolution (subarcminute) imaging in this frequency region requires long (> 2 km) baselines; but the phase error due to ionospheric refraction increases with baseline length. Due to these issues, availability of instruments for high-resolution imaging at low frequencies has been limited. Interest therefore shifted to the science accessible to large centimeter-wavelength aperture synthesis radio telescopes, such as the Very Large Array (VLA) of the National Radio Astronomy Observatory (NRAO), commissioned in the early 1980s [2].

However, interest in low frequency radio astronomy is gradually being revived due to several factors. In the early 1990s, a technique was developed which dramatically improved astronomers' ability to mitigate the effects of the ionosphere in aperture synthesis imaging [3]. This technique allowed resolution on the subarcminute scale for the first time in this frequency range. In addition, the possibility of answering an increasing number of questions in astrophysics and cosmology through low frequency radio astronomy has emerged [4].

Now that interest in low frequencies is increasing we find ourselves limited, because existing instruments do not support low frequency observation very well. Two possible ways to overcome this situation are:

a) Construction of new telescopes. Several new large telescopes are therefore being developed. Examples include the Low Frequency Array¹ (LOFAR) in the Netherlands [5], the Long Wavelength Array² (LWA) in New Mexico [6], and the Murchison Wide-Field Array³ (MWA) in Western Australia [7]. These are entirely new instruments using phased array design.

b) Enhancement of the observing capabilities of existing telescopes by including new low frequency systems. For example, it has been proposed in [8] that the Expanded Very Large Array⁴ (EVLA) in New Mexico [9] be equipped with a new low frequency system (LFS) covering the range roughly 50–1000 MHz (excluding 470-700 MHz, which is not accessible through the EVLA electronics).

In this dissertation, we describe and demonstrate design concepts to modify existing reflector antennas for covering as much of the frequency range of 10 - 500 MHz as possible, with a useful level of sensitivity. The upper and lower bounds are not stringent. The lower bound of the range is decided to be 10 MHz, as the ionosphere essentially becomes opaque below this frequency, and also because large reflectors become electrically small and thus unusable – not really behaving "optically". The upper bound of 500 MHz is selected as significant efforts are already being made to develop reflector antenna feeds for frequencies beyond this range. The most popular of these feeds are dense phased arrays [10]. Major phased array feed (PAF) development efforts include APERTIF ("APERture Tile In Focus") (Astron, The Netherlands) [11], ASKAP (CSIRO, Australia) [12], PHAD (DRAO,

¹http://www.lofar.org/

²http://lwa.unm.edu/

³http://www.mwatelescope.org/

⁴http://www.vla.nrao.edu/

Canada) [13], and the BYU/NRAO centimeter-band PAF [14]. Within the range of our interest, radio frequency interference (RFI) from FM broadcast stations in the frequency range of 88–108 MHz, and digital TV broadcast stations in the range of 174-216 MHz, make observing extraordinarily difficult; however, we may not wish to completely rule out the possibility of productive observing in these bands given the possibility of continued progress in development of RFI mitigation techniques.

Our objective is twofold: 1) By modification of existing reflector antenna systems, cover as much of the frequency range of 10–500 MHz as possible with a useful level of sensitivity, and 2) achieve (1) without compromising the operation of the pre–existing higher–frequency systems. In order to fulfill the objective, a new feed design is proposed in this dissertation – namely, the "distributed feed array". The feed consists of an array, with the array elements distributed at locations away from the central axis of the reflector. The signals from the array elements are combined in a beamforming manner to maximize the sensitivity of the system. The advantage of this strategy is that the additional feeds are placed in regions of lower power density, and so blockage at higher frequencies is presumably reduced; thus coexistence with the higher–frequency systems might be possible. This will be discussed in detail later in this dissertation.

The work presented in this dissertation should be applicable to any existing system using large reflector antennas, e.g., the Giant Meterwave Radio Telescope $(GMRT)^5$, the Westerbork Synthesis Radio Telescope $(WSRT)^6$, and the Australia Telescope Compact Array $(ATCA)^7$. Also, this work may contribute to the developments needed to realize the Square Kilometer Array (SKA) [15].

In order to demonstrate an example, the findings of this research will be applied to the development of an improved low frequency system for the EVLA. The EVLA consists of 27 Cassegrain reflector antennas on the Plains of San Agustin, New Mexico. An EVLA

⁵http://gmrt.ncra.tifr.res.in/

⁶http://www.astron.nl/

⁷http://www.narrabri.atnf.csiro.au/

reflector antenna is shown in Figures 1.1 and 1.2. Each reflector antenna is 25 meters in diameter, and has a f/D (focal length to diameter ratio) of 0.36. The original design of this instrument included four frequency bands corresponding to wavelengths of 21 cm (\approx 1.4 GHz), 6 cm (5 GHz), 2 cm (15 GHz), and 1.3 cm (23 GHz) [2]. The first low frequency system was installed in 1989, and operated at 90 cm (bandlimited to about 300–340 MHz in the receiver) [16]. The feeds for this system are shown in Figure 1.3. The success of the "90 cm system" (also variously known as the "P–band system" or "327 MHz system") led to the consideration of a lower frequency observing capability. The "4 m system" (also variously known as the "4–band system" or "74 MHz system") was thus installed to operate in the range 73–74.6 MHz [3], [17]. The feeds for this system are also shown in Figure 1.3. Yet another low frequency system was installed briefly (also shown in Figure 1.3) to operate from 186–202 MHz [18]. However, the feeds of this system unacceptably degraded the performance of the 90 cm system, and were removed before the system was fully installed.

The EVLA LFS would upgrade or replace the existing 4 m and 90 cm systems. Moreover, frequency coverage with the LFS will be contiguous to frequencies as low as 50 MHz. The findings of the research presented in this dissertation are applied in designing candidate low frequency systems for the EVLA LFS. The large tuning range and low degradation of higher–frequency systems desired for the LFS make this project considerably different from previous low frequency modifications, and are primary motivations for the research reported in this dissertation.

The rest of this chapter is organized as follows. Section 1.1 summarizes the state of the art in low frequency systems for reflector antennas. In Section 1.2, we look at some relevant current trends in feed design. The contributions of the work presented in this dissertation are summarized in Section 1.3. The organization of the rest of this dissertation is provided in Section 1.4.



Figure 1.1. An EVLA reflector antenna. Image source: NRAO/AUI and Patrick A. Lofy, Photographer (used with permission, see Appendix E).

1.1 State of the Art

Some of the existing large reflector antennas that have low frequency observing capability include the EVLA, the GMRT and the WSRT. The sensitivity of the low frequency systems on these instruments is summarized in Table 1.1. Sensitivity is also expressed in terms of the *System Equivalent Flux Density* (SEFD), which is defined as the incident flux spectral density (W m⁻² Hz⁻¹) which yields signal-to-noise ratio (SNR) of unity at the output of the system. SEFD can be expressed as

$$SEFD = \frac{2kT_{sys}}{e_I A_e} , \qquad (1.1)$$

where k is the Boltzmann's constant (1.38 × 10⁻²³ J/K), e_I is the impedance mismatch efficiency (IME) between the feed and the receiver, A_e is the effective aperture of the reflector



Figure 1.2. Close up of feeds and struts on an EVLA reflector antenna. Courtesy of Lincoln Greenhill, Harvard/SAO (used with permission, see Appendix E).

antenna, and T_{sys} is the system noise temperature, defined as $T_{sys} = e_I T_A + T_{int}$. In this expression, T_A is the antenna noise temperature, and T_{int} is the internal noise temperature of the system, referred to the feed antenna terminals. A detailed derivation of each term in the SEFD expression is given later in this dissertation. For the values reported in Table 1.1, e_I is assumed to be 1, which corresponds to a matched condition between the feed antenna and the connected electronics; A_e and T_{sys} values are extracted from the references indicated in the table. In this table, SEFD is expressed in units of Jansky which is equivalent to 10^{-26} Wm⁻²Hz⁻¹. In the following paragraphs the feed antennas for these low frequency systems are briefly described.

The EVLA 90 cm system employs crossed dipoles as the feed. Size constraints preclude



Figure 1.3. Low frequency feeds on an EVLA reflector antenna. Only the 90 cm feed is permanent. The 153 cm feed was experimental and never adopted (see text). The 4 m feed is only intermittently used. Courtesy of Lincoln Greenhill, Harvard/SAO (used with permission, see Appendix E).

Cassegrain operation much below 1 GHz; whereas prime focus operation is precluded by the fact that the subreflector cannot be moved out of the optical path. Therefore the dipoles are placed 0.23 m ($\lambda/4$ at 327 MHz) in front of the subreflector which then acts as a crude backplane. The resulting phase center is about one-half wavelength displaced from the prime focus, resulting in relatively low aperture efficiency (40%) [16]; nevertheless, the system achieves a useful level of performance over 300–340 MHz (12.5% bandwidth), and has been scientifically productive.

The EVLA 4 m system also employs crossed dipoles as the feed. The dipoles are located at the image of the prime focus (as reflected by the subreflector). With respect to the 90 cm system, the subreflector is a less efficient reflector; thus, the aperture efficiency is significantly less (somewhere in the range 15% [17] to 25% [3]). Unfortunately, the blockage and scattering from the 4 m system reduces the sensitivity of the 21 cm system by about 6% [17]; therefore, the dipoles are not permanently installed, and are only intermittently used.

The GMRT, located in India, consists of 30 paraboloidal reflectors. Each of the reflectors

Telescope	Dia.	Feed	Nominal Freq.	Aperture	T_{sys}	SEFD
Reference	(m)	Tuning Range	f	Eff. @ <i>f</i>	@f	@~f
EVLA	25	Dipole	327 MHz	40%	130 K	1.8 kJy
[16]		300–340 MHz				
EVLA	25	Dipole	74 MHz	25%, 15%	$1500 \mathrm{K}$	33.9 kJy, 55.0 kJy
[3], [17]		73–74.6 MHz				
GMRT	45	Inv. Vee Dipole Array	55 MHz	65%	4000 K	10.5 kJy
[19]		30–90 MHz				
GMRT	45	Folded Dipole Array	150 MHz	65%	428 K	1.1 kJy
[20]		117–247 MHz				
GMRT	45	Coaxial Waveguide	233 MHz	65%	229 K	0.6 kJy
[20]		221–245 MHz				
GMRT	45	"Kildal Feed"	327 MHz	66%	152 K	0.4 kJy
[20]		286–424 MHz				
WSRT	25	Folded Dipole Array	140 MHz	32%	480 K	8.5 kJy
[21]		115–180 MHz				

Table 1.1. Comparing the low frequency systems on the existing reflector antennas of several radio telescopes. Comparison is based on a single reflector antenna performance. Note that SEFD is calculated using Equation 1.1, assuming $e_I = 1$, and using values for A_e and T_{sys} extracted from the indicated references.

is 45 m in diameter with a f/D ratio of 0.41. A GMRT reflector antenna is shown in Figure 1.4. The GMRT operates in six bands – namely, 1000 - 1450 MHz, 610 MHz, 327 MHz, 233 MHz, 150 MHz and 50 MHz. The 50 MHz system for the GMRT [19] has been designed to cover the frequency range of 30–90 MHz. This system uses four inverted Vee dipoles, each of them 2.4 m in length, in a "boxing ring" configuration, as shown in Figure 1.5. In order to have better bandwidth characteristics, the arms of the Vee dipoles are "folded". The feed point is positioned 1 m above a square reflector with each side being 3 m in length.

The feed for the GMRT 150 MHz system also consists of four dipoles in a boxing ring configuration, placed above an octagonal reflector [20]. These dipoles are "thick" folded dipoles with each arm having a length-to-diameter ratio of 6.48. The largest diagonal of the octagonal reflector is 1.2λ .

The GMRT 233 MHz system employs coaxial waveguide feed. The aperture diameter for



Figure 1.4. A GMRT reflector antenna. Courtesy: Suman Majumdar, IIT Kharagpur, India (used with permission, see Appendix E).

the 233 MHz coaxial feed is 0.85 λ [20].

The feed used for the 327 MHz system on the GMRT is a modified "Kildal Feed" [20]. The Kildal Feed is constructed by locating a dipole between a reflector and a "beam forming ring" [22], [23]. The conducting ring is placed above the dipole in a plane parallel to the reflector, and is supported by dielectric rods. The purpose of the ring is to compress the otherwise broad H-plane pattern of the dipole so that a better balance between the E- and H-plane patterns is obtained. The diameter of the reflector for the 327 MHz feed is 2.20 λ , and the diameter of the beamforming ring is 1.22 λ .

Another radio telescope that offers low frequency capability is the WSRT. WSRT consists of fourteen identical reflector antennas, each having a diameter of 25 m. Figure 1.6 shows the low frequency feed on a WSRT reflector antenna. The low frequency system on the WSRT



Figure 1.5. The GMRT 50 MHz feed [19]. The left image shows the folded V-dipoles in a boxing ring configuration. The right image shows its mounting on the turret of a GMRT antenna (used with permission, see Appendix E).

employs folded dipoles, and operates in the range 115 - 180 MHz [21]. The feed dipoles can be retracted in a way such that they do not affect the performance of the antenna system during observations at other bands. This ensures uncompromised performance for the preexisting (higher-frequency) feed systems. However, the WSRT low frequency receivers are now being decommissioned, as APERTIF will be employed on WSRT antennas to cover the frequency range 1-1.7 GHz [11], and the installation of this array requires the decommission of the low frequency receivers.

Now that the state of the art in low-frequency reflector antenna systems is described, it is a good place in the discussion to present the estimated performance of the new system proposed in this dissertation for the EVLA. The new system is designed to cover the frequency range of 50–250 MHz. The feed for this system consists of two rings of dipoles employed between the adjacent struts of the EVLA. The performance of this system at 74 MHz and 150 MHz is shown in Table 1.2. Also shown in the table is the performance of the existing EVLA 4 m system computed using the same method used to estimate the performance of the proposed systems. It is observed that the proposed double ring of dipoles offers slightly better performance to the existing system at 74 MHz; also a useful level of sensitivity is anticipated at 150 MHz. Moreover, it is shown later in this dissertation that the blockage presented by the proposed double ring system to the existing higher-frequency systems is less than (by about 4.5%) that presented by the existing 4 m system at L-band. This suggests



Figure 1.6. The 150 MHz feed on a WSRT reflector antenna [21]. The left image shows the position of the feed when it is in operation and the right image shows the stowed position of the feed (used with permission, see Appendix E).

the possibility of permanent installation of the proposed feed.

1.2 Current Trends in Low Frequency Feed Development

In this section, we summarize the current trends in the development of low frequency feeds for reflector antennas. The first feed antenna that we address here has a decade bandwidth, and it is called the "Eleven antenna" [24]. This antenna is based on the concept that two parallel dipoles placed $\lambda/2$ distance apart, at a height of $\lambda/4$ over ground, produce almost equal E- and H-plane patterns, thereby reducing cross-polarization. In the Eleven antenna each dipole is replaced by a log-periodic array of dipoles to achieve larger impedance bandwidth.

The Eleven antenna has been optimized for the frequency range of 200–800 MHz for use on the GMRT [25]. Figure 1.7 shows this feed. The dimension of the metal box is 1.46 m \times 1.46 m \times 0.50 m. The aperture efficiency obtained with this feed is better than 47% in the desired band. Using a differently–scaled version of the Eleven antenna, another low frequency system has been tested for operation from 150–1700 MHz [26]. This system was installed on the 43 m-diameter reflector antenna located in Green Bank, West Virginia. The width and the height of the box is 1.64 m and 0.55 m, respectively. The aperture efficiency of the reflector antenna when fed with the Eleven antenna is reported to be better than

Low Frequency System	Nominal Freq.	SEFD
Existing EVLA 4 m System	$74 \mathrm{~MHz}$	55.0 kJy
Proposed Double Ring of Dipoles (Section 6.3.2)	$74 \mathrm{~MHz}$	31.6 kJy
Proposed Double Ring of Dipoles (Section 6.3.3)	150 MHz	4.0 kJy

Table 1.2. Comparing the performance of the distributed feed array, proposed in this dissertation, with the existing low frequency system on an EVLA antenna.

59% in the entire band. However, the physical dimensions of the Eleven antenna become prohibitively large if lower frequency coverage is desired. For example, the design developed by Olsson (2006) [24] uses a ground plane having a width of 0.88 λ_{max} . This corresponds to a ground plane width of 5.3 m at 50 MHz. Therefore coexistence with higher-frequency systems is not possible if the Eleven antenna is used; also the removal and reinstallation of this feed would be complex and labor-intensive. Therefore we no longer consider the Eleven antenna or similar design concepts as a possible candidate for low frequency systems of interest.

Another feed concept that has been suggested for low frequency systems is a focal plane array which employs a phased array of feed antennas in the focal plane of a reflector antenna. Focal plane phased arrays are being considered for the WSRT for operation in the range 1– 1.7 GHz [11]. Also, a phased array of Vivaldi antennas has previously been considered for the EVLA for observation in the frequency range of 250–750 MHz [27]. The dimensions of the square array is 1.2 m on each side. One issue of concern with the focal plane array is that a certain phasing of this array that optimizes performance at one frequency performs well only over a relatively small band around that frequency. For example, in the case of the phased array proposed for the EVLA, sensitivity at 500 MHz, obtained with phasing optimized at 500 MHz, is 55% greater than that observed with phasing optimized at 460 MHz. Moreover, if this phased array is to be used at lower frequencies, then the array size needs to be increased. For example, at 10 MHz and 50 MHz the array size has to be around 30 m and 6 m, respectively, which is not reasonable. Therefore we do not consider focal plane array



Figure 1.7. The optimized Eleven antenna for the GMRT [25] (used with permission, see Appendix E).

as a suitable low frequency feed concept in this dissertation.

Since we want to cover a large frequency range using a reflector antenna, one particular concept of interest is an impulse radiating antenna (IRA) [28]. The concept of a reflector IRA is shown in Figure 1.8. The reflector IRA consists of a paraboloidal reflector fed by a pair of conical coplanar TEM transmission lines. The size and the angles for these conical coplanar plates are designed to achieve a specified characteristic impedance. A reflector IRA can be well-matched over a wide frequency range. For example, for a 200 Ω system, the return loss can be better than 5 dB in the frequency range of 10–1000 MHz [28]. However, the maximum aperture efficiency is only about 35% for a TEM transmission line–fed reflector



Figure 1.8. The concept of a reflector IRA.

IRA (achieved for $\phi_o = 70^\circ$, and $b_1/a = 0.84$ [29]). This corresponds to b_1 and b_2 values of 10.5 m and 14.9 m, respectively for a 25 m-diameter reflector antenna. A lot of structural change is required if an existing reflector antenna is to be fed with this kind of a feed. We consider this concept to be out of scope of this dissertation, but a possible area for future work.

It can be seen from the above discussion that most effort in low-frequency feed design is directed toward achieving broad impedance bandwidth. The resulting feeds are mechanically complex, difficult to construct, and thus expensive. Also, due to the large dimensions associated with these feeds, they introduce large blockage, and hence are not suitable for coexistence with the pre-existing higher-frequency systems. Simple dipole-like feed antennas, on the other hand, are inexpensive, but they have much narrower impedance bandwidth. However, in the frequency range of interest, even simple narrowband feed antennas can offer useful sensitivity over a large bandwidth [30]. This is because, at these frequencies, the external noise can easily dominate over the self-noise of the electronics connected to the feed, even after a large impedance mismatch. Once the instrument is external noise–limited, improvement in the receiver noise temperature or in the impedance matching has little effect on the sensitivity. Unless the impedance mismatch becomes bad enough, such that the system is no longer external noise–limited, the low–frequency system continues to offer external noise–limited sensitivity. Therefore in this dissertation, we try to take advantage of this scenario at low frequencies, and consider simple, inexpensive dipole–like feed antennas for covering large tuning ranges.

It is also observed that the issue of coexistence with the pre-existing higher-frequency systems is not considered in the design of most low-frequency systems. Most of the feeds are employed at the focus, or they lie in the optical path from the reflector to the focus. Moreover, a number of these low-frequency feeds include a large reflector as a part of their geometry, which precludes simultaneous operation at higher frequencies. A possible way to resolve this issue might be adopting a distributed feed array scheme which employs arrays of feed elements distributed in the volume in front of the reflector. An appreciable amount of power is available at locations away from the reflector axis at low frequencies. This is demonstrated in Figure 1.9, where the power density in the focal plane of a 25 m-diameter reflector antenna is calculated at 500 MHz and 50 MHz. It is observed that as the frequency goes down, the available power spreads out from the focal point. This behavior suggests the possibility of absorbing a useful portion of the available power at low frequencies using an array of feeds distributed at locations away from the axis of the reflector. This low-frequency system would be less likely to hinder operation at higher frequencies. Therefore the utility of the distributed feed array is explored in detail in this dissertation.



Figure 1.9. Distribution of power density in the focal plane of a reflecting paraboloid (D = 25 m, f/D = 0.36) relative to the power density at the focus in response to plane wave illumination at (a) 500 MHz, (b) 50 MHz.

1.3 Contributions

This dissertation seeks to improve upon the state of the art in low-frequency systems for reflector antennas. The objective is to develop new dipole-based feed systems which offer a useful level of sensitivity over a large frequency range, and coexist with the pre-existing higher-frequency systems.

Our contributions to this area of research consist of the following elements:

• Electromagnetic Analysis of Reflector Antennas at Low Frequencies: Analysis techniques used for reflector antennas at the traditional high frequencies of operation, e.g., physical optics (PO) and the uniform geometrical theory of diffraction (UTD), are not necessarily accurate at low frequencies, especially when the reflector is not very large compared to a wavelength. However, the computational burden associated with these techniques is much less than other popular analysis methods suitable for these frequencies, e.g., the method of moments (MoM) and finite difference time domain (FDTD). Therefore we attempt to establish low-frequency limits on the usefulness of PO and UTD, so that one may know when they may be employed in computationally-efficient iterative design processes even at low frequencies. Detailed descriptions are provided in Chapter 2 (Electromagnetic Analysis of Large Reflector Antennas at Low Frequencies).

- Analysis of Sensitivity of Reflector Antennas Fed with Array Feeds: In this dissertation, a methodology for estimating the sensitivity of reflector antenna systems that employ array feeds is described. In this methodology, mutual coupling is taken into account. Moreover, this particular methodology accounts for the correlation of external noise between array elements, which can significantly desensitize the system. This is especially important at low frequencies when the external noise can easily dominate over the internal noise of the system. Chapter 3 (Analysis of Sensitivity of Large Reflector Antennas at Low Frequencies) describes this methodology.
- Focus- or Near Focus-Fed Systems: Using the above techniques, we analyze the performance of systems consisting of simple dipoles. Three sets of crossed dipoles namely, 74 MHz-resonant crossed dipoles, 150 MHz-resonant crossed dipoles and 327 MHz-resonant crossed dipoles are identified that can cover the frequency range of 50–470 MHz with a useful level of sensitivity when they are employed at/near (in cases where the focus is not accessible) the focus of large reflector antennas. The performance of these systems serves as a baseline against which new designs are evaluated. The analysis of these systems and the results are described in Chapter 5 (Focus- or Near Focus-Fed Systems).
- Distributed Feed Array: A new feed design concept namely, the distributed feed array is described. In this scheme, feed elements are distributed in the volume in front of the reflector. The signals from these feeds are then combined to achieve performance comparable to focus– or near focus–fed systems. The advantage of this scheme is that the blockage presented to the pre–existing higher–frequency systems is presumably reduced. The design concept is described in Chapter 6 (Distributed

Feed Array).

- *LFS for the EVLA*: As a demonstration of the application of the research presented in this dissertation, a low-frequency system covering the frequency range of 50 470 MHz is designed and analyzed for the EVLA. A complete MoM model of an EVLA antenna is created for this purpose. The description of the proposed LFS is given in Chapter 6 (Distributed Feed Array).
- Assessment of Blockage: A simple method for assessment of blockage presented by low– frequency feeds to the pre–existing higher–frequency systems on the EVLA is described in Chapter 7 (Assessment of Blockage due to Low Frequency Feeds on the EVLA).

The design and analysis techniques presented in this dissertation should be applicable to other large reflector antennas as well.

1.4 Organization

The contents of Chapters 2–7 are as described in the previous section. The findings of this research are summarized in Chapter 8 (**Conclusions**), where suggestions for future work are also identified.

Chapter 2

Electromagnetic Analysis of Large Reflector Antennas at Low Frequencies

The modification of large reflector antennas for low frequency operation is not simply a design challenge. It also poses an analysis challenge. In this chapter, we address the electromagnetic analysis of reflector antennas at low frequencies. At traditional frequencies of operation, the analysis of large reflector antennas can be divided into two parts. First, the feed can be analyzed independently of the reflector antenna. Once the feed pattern is calculated, it can be used to illuminate the reflector antenna, and thereby obtain the final pattern and gain characteristics. However, at the frequencies of interest in this dissertation, especially at the lower end of the desired frequency range, it is not necessarily reasonable to decompose the problem in this way, as the coupling and scattering among the feed, reflector, and supports is significant. Another issue of concern is that commonly used techniques for gain and pattern analysis of reflector antenna systems – in particular, physical optics (PO) and the uniform geometrical theory of diffraction (UTD) – are not certain to be reliable in the frequency region of our interest. It thus becomes necessary to use, instead, a full wave method such as the method of moments (MoM). However, the computational burden involved with MoM is many orders of magnitude greater than either PO or UTD. It will therefore be useful if we can establish a low frequency limit on the usefulness of PO and UTD. This can be done by comparing PO and UTD results with MoM results, especially at the lower frequencies.

In Sections 2.1, 2.2 and 2.3 we briefly describe PO, UTD, and MoM, respectively. A comparison of these techniques at frequencies below 100 MHz is presented based on the analysis of a simple 26 m-diameter reflector antenna in Section 2.4. In Section 2.5, a complete model of an EVLA antenna is developed for MoM analysis. This is useful because the model is required for the work described in the subsequent chapters of this dissertation.

2.1 Physical Optics

In this section, we briefly describe the analysis of a prime focus-fed axisymmetric reflector antenna using PO. The transmitted far field \mathbf{E} of a reflector antenna can be calculated by integrating the contributions from the surface current $\mathbf{J}^{\mathbf{s}}$ over the reflector surface S_r . This can be expressed as [31]:

$$\mathbf{E} = -j\omega\mu \frac{e^{-jkr}}{4\pi r} \int_{S_r} [\mathbf{J}^{\mathbf{s}} - (\mathbf{J}^{\mathbf{s}} \cdot \hat{\mathbf{r}})\hat{\mathbf{r}}] e^{jk\hat{\mathbf{r}} \cdot \mathbf{r}'} dS' , \qquad (2.1)$$

where ω is angular frequency, μ is permeability and k is the propagation phase constant. The distance from the origin to the observation point is expressed as r, $\hat{\mathbf{r}}$ is the unit vector along r, and r' is the distance from the origin to the source point, i.e., the surface of the reflector in this case.

In PO analysis of a reflector antenna, $\mathbf{J}^{\mathbf{s}}$ is found using the PO approximation. In this approximation, we assume that the field incident from the feed behaves locally as a plane wave, and scattering from each point on S_r is equal to that expected due to reflection from an infinite perfect electric conductor (PEC) planar surface tangent to the point of scattering. Boundary conditions then dictate that the equivalent surface current is related to the incident magnetic field $\mathbf{H}^{\mathbf{i}}$ as

$$\mathbf{J}^{\mathbf{s}} = 2(\hat{\mathbf{n}} \times \mathbf{H}^{\mathbf{i}}) , \qquad (2.2)$$

where $\hat{\mathbf{n}}$ is the unit normal to the reflector surface at the point of reflection.

The integral in Equation 2.1 is typically evaluated numerically. In order to perform the integration, the reflector surface can be modeled using patches having an area of $0.01\lambda^2$ where λ is the wavelength under consideration. However, there is a trade-off between computational burden and accuracy while selecting the patch size. Smaller size patches give more accurate results, but at the price of higher computational burden. In a preliminary study to determine a reasonable patch size, we experimented with a 26 m-diameter reflector. This reflector antenna, when modeled using patches of area $0.01\lambda^2$, results in 147.477 patches at 500 MHz. It takes about 30 minutes on a Dell Inspiron 1545 (2.1 GHz processor, 3 GB RAM) machine to compute the pattern at 180 points. The directivity and the sidelobe level are computed to be 37.4 dB and 11.7 dB, respectively. On the other hand, patches of area $0.03\lambda^2$ result in 16,405 patches at 500 MHz, and it takes about 16 seconds to compute the same results. The directivity and the sidelobe level are then calculated to be 37.4 dB and 11.6 dB, respectively; i.e., very close to the previous results. The reduction in computational burden is significant. whereas the difference in the results is negligible. Therefore in this dissertation, a patch area of 0.01 m² (corresponding to $0.03\lambda^2$ at 500 MHz) is used for modeling reflector antennas for PO analysis at or below 500 MHz.

The discontinuity at the rim of the reflector separating the illuminated and shadowed regions is neglected in PO. As a result PO can predict the main beam and the first few sidelobes only, and fails for far-out sidelobes.
2.2 Uniform Geometrical Theory of Diffraction

In order to obtain the far-out sidelobes in the pattern of a reflector antenna we need to consider diffraction from the rim of the reflector. One possible way to do this is to use UTD [32]. In this section, we briefly describe UTD analysis of an axisymmetric reflector antenna. We will consider only single diffractions that occur at the edge of the reflector surface.

The geometry used for UTD analysis of a reflector antenna is shown in Figure 2.1 and Figure 2.2. The edge of the reflector surface lies in the xy-plane with the focus F located at the origin of the coordinate system. According to the "law of edge diffraction", the diffracted field at a point P originates from the two "stationary points" Q_1 and Q_2 . These two points correspond to the intersection of the reflector edge with the plane passing through FCP. The contribution to the total diffracted field from Q_1 at P may then be written as

$$\begin{bmatrix} \mathbf{E}_{||}^{d} \\ \mathbf{E}_{\perp}^{d} \end{bmatrix} = \begin{bmatrix} D_{s} & 0 \\ 0 & D_{h} \end{bmatrix} \begin{bmatrix} \mathbf{E}_{||}^{i}(\mathbf{Q}_{1}) \\ \mathbf{E}_{\perp}^{i}(\mathbf{Q}_{1}) \end{bmatrix} \sqrt{\frac{\rho^{c}}{s^{d}(s^{d}+\rho^{c})}} e^{-jks^{d}} , \qquad (2.3)$$

where s^d is the distance from the observation point P to the diffraction point, $\mathbf{E}_{\parallel}^i(\mathbf{Q}_1)$ and $\mathbf{E}_{\perp}^i(\mathbf{Q}_1)$ refer to the incident fields which are respectively parallel and perpendicular to the plane of incidence, and \mathbf{E}_{\parallel}^d , \mathbf{E}_{\perp}^d refer to the diffracted fields which are respectively parallel and perpendicular to the plane of diffraction. The contribution to the total diffracted field from \mathbf{Q}_2 is calculated similarly. The plane containing the incident ray and the normal to the reflector surface (at the point of incidence) is referred to as the plane of incidence, whereas the plane containing the diffracted ray and the normal to the reflector surface (at the point of as the plane of diffraction. ρ^c is the caustic distance defined by

$$\frac{1}{\rho^c} = \frac{1}{\rho_e^i} - \frac{\hat{\mathbf{n}}_e \cdot (\hat{\mathbf{s}}^i - \hat{\mathbf{s}}^d)}{a_e \sin^2 \beta_0} , \qquad (2.4)$$

where



Figure 2.1. Geometry for UTD analysis of an axisymmetric paraboloidal reflector [33].

 ρ_e^i = radius of curvature of the incident wavefront at the diffraction point (Q₁ or Q₂) in the plane containing the incident ray and the tangent to the edge.

 $\hat{\mathbf{n}}_e$ = unit normal vector to the edge at the point of diffraction (Q₁ or Q₂), directed outward from the center of curvature.

 $\mathbf{\hat{s}}^{i},\,\mathbf{\hat{s}}^{d}$ = unit vectors along the incident and diffracted rays respectively.

 a_e = radius of curvature of the edge at the point of diffraction (Q₁ or Q₂).



Figure 2.2. Illustrating the plane passing through the edge of the paraboloid [33].

 β_0 = angle (in the plane of incidence) between the edge tangent and the incident ray at the point of diffraction.

For UTD analysis, the rim of the reflector can be modeled as a half-plane tangent to the surface, as shown in Figures 2.3 and 2.4. This is a reasonable approximation if the radius of curvature of the reflector surface in this plane is large compared to wavelength. From the geometry of the diffracted rays we find that

$$\hat{\mathbf{n}}_e^{1,2} \cdot \hat{\mathbf{s}}_{1,2}^i = \sin \alpha , \qquad (2.5)$$

$$\hat{\mathbf{n}}_{e}^{1,2} \cdot \hat{\mathbf{s}}_{1,2}^{d} = \pm \sin \theta$$
, and (2.6)

$$\rho_e^i = s^i . (2.7)$$



Figure 2.3. Geometry of incident and diffracted ray at Q_1 [33].

Inserting Equations 2.5, 2.6, and 2.7 in Equation 2.4, we obtain

$$\rho_{1,2}^c = \pm s^i \, \frac{\sin \alpha}{\sin \theta} \,. \tag{2.8}$$

 ${\cal D}_h$ and ${\cal D}_s$ are the UTD diffraction coefficients, and are given by

$$D_{s,h}(L,\phi^{i},\phi^{d}) = \frac{-e^{-j(\pi/4)}}{2n\sqrt{2\pi k}\sin\beta_{o}} \\ \times \left[\cot(\frac{\pi + (\phi^{i} - \phi^{d})}{2n})F[kL^{i}a^{+}(\phi^{i} - \phi^{d})] + \cot(\frac{\pi - (\phi^{i} - \phi^{d})}{2n})F[kL^{i}a^{-}(\phi^{i} - \phi^{d})]\right]$$



Figure 2.4. Geometry of incident and diffracted ray at Q_2 [33].

$$\mp \left\{ \cot(\frac{\pi + (\phi^{i} + \phi^{d})}{2n}) F[kL^{rn}a^{+}(\phi^{i} + \phi^{d})] + \cot(\frac{\pi - (\phi^{i} + \phi^{d})}{2n}) F[kL^{ro}a^{-}(\phi^{i} + \phi^{d})] \right\} \right],$$

$$(2.9)$$

where F is the transition function. If the argument of F is represented by X,

$$F(X) = 2j|\sqrt{X}|e^{jX} \int_{|\sqrt{X}|}^{\infty} e^{-j\tau^2} d\tau .$$
 (2.10)

The details for computation of F(X) are provided in Appendix A. To determine a^{\pm} we use

$$a^{\pm}(\psi) \equiv 2\cos^2\left[\frac{2\pi nN^{\pm} - \psi}{2}\right] , \qquad (2.11)$$

in which N^{\pm} is the integer -1,0,+1 that most nearly satisfies the expression:

$$2\pi n N^{\pm} - \psi = \pm \pi . \tag{2.12}$$

 L^{i} , L^{ro} and L^{rn} are distance parameters which for spherical wavefront becomes

$$L^{i,ro,rn} = \frac{s^i s^d}{s^i + s^d} \sin^2 \beta_o$$

Once we get the fields diffracted from point Q_1 (\mathbf{E}_{Q_1}) and Q_2 (\mathbf{E}_{Q_2}), we add to them the direct-radiated field from the feed (\mathbf{E}_D) to obtain the total field \mathbf{E}_T as

$$\mathbf{E}_{\mathrm{T}} = \mathbf{E}_{\mathrm{Q}_{1}} + \mathbf{E}_{\mathrm{Q}_{2}} + \mathbf{E}_{\mathrm{D}} \ . \tag{2.13}$$

In Section 2.4 we show examples of results computed using UTD.

2.3 Method of Moments

For the purpose of analyzing reflector antennas in this dissertation, we need a method that is certain to be valid even at the lowest frequencies of interest, so that we can be sure of our analysis. Two such low frequency techniques are MoM and the finite difference time domain (FDTD) method [34]. MoM can provide results for small (few wavelengths in size), relatively simple antennas with much less computation time and memory than FDTD, since MoM finds only the currents flowing on the wire or conducting surfaces, whereas FDTD must calculate the fields in the entire computational region which must contain enough cells to allow near-field decay between the antenna and the absorbing boundaries [31]. In our frequency range of interest, a reflector antenna might not be very large compared to wavelength (e.g., a 25 m-diameter reflector is only about 4λ in diameter at 50 MHz). Therefore MoM is considered to be more suitable for low frequency analysis of reflector antennas.

We now give a brief description of MoM. For simplicity, let us consider a wire antenna of length L along the z axis. A generic form for an integral equation describing such an antenna is

$$-\int_{-L/2}^{L/2} I(z')K(z,z')dz' = E^{i}(z) , \qquad (2.14)$$

where I(z) is the current which, in free space, radiates the same electric field produced by the total currents radiating in the presence of the wire; and $E^i(z)$ is the electric field on the surface defined by the wire surface, but due instead to the impressed (source) currents radiating in free space. The kernel K(z, z') depends on the specific integral equation formulation used; e.g., for the popular Pocklington form, the above equation can be expressed as

$$-\int_{-L/2}^{L/2} I(z') \left(\frac{1}{j\omega\epsilon_o} \left[\frac{\partial^2 \psi(r,z')}{\partial z^2} + k^2 \psi(r,z')\right]\right) dz' = E^i(z) , \qquad (2.15)$$

where ϵ_0 is permittivity and $\psi(r, z)$ is the free-space Green's Function. $\psi(r, z)$ is defined as

$$\psi(r,z) = \frac{e^{-jk|\mathbf{r}-\mathbf{z}|}}{4\pi|\mathbf{r}-\mathbf{z}|} .$$
(2.16)

This integral equation is solved numerically for the current I(z). The first step in solving this equation is to divide the wire into N small pieces, referred to as *segments*. The next step is to approximate the unknown current in terms of a sum of basis functions $F_n(z)$ such that

$$I(z') = \sum_{n=1}^{N} I_n F_n(z') , \qquad (2.17)$$

where the I_n 's are complex expansion coefficients, and are the unknowns. Many different approaches for defining $F_n(z)$ are described in [31]; for example, a common approach is to take the expansion functions to be a set of orthogonal pulse functions given by

$$F_n(z) = \begin{cases} 1 & \text{for } z \text{ in segment } n, \\ 0 & \text{otherwise.} \end{cases}$$
(2.18)

Substituting Equation 2.17 into Equation 2.14 gives

$$-\sum_{n=1}^{N} I_n \left[\int_{-L/2}^{L/2} F_n(z') K(z, z') dz' \right] = E^i(z) .$$
 (2.19)

Equation 2.19 can be written compactly as

$$\sum_{n=1}^{N} Z_n(z) I_n = V_n(z) , \qquad (2.20)$$

where

$$Z_n(z) = -\int_{-L/2}^{L/2} F_n(z') K(z, z') dz' , \qquad (2.21)$$

and

$$V_n(z) = E^i(z)$$
 . (2.22)

Now the problem becomes to find N simultaneous equations to allow solution of Equation 2.20 for the I_n 's. To generate N simultaneous equations, a simple approach is to evaluate Equation 2.20 at N points corresponding to the centers of the N pulses in the current expansion. This can be expressed as

$$\sum_{n=1}^{N} Z_n(z_m) I_n = V_n(z_m) \quad , \, m = 1, 2, \dots N.$$
(2.23)

This results in the following system of equations:

$$I_{1}Z_{11} + I_{2}Z_{12} + \dots + I_{N}Z_{1N} = V_{1}(z_{1})$$

$$I_{1}Z_{21} + I_{2}Z_{22} + \dots + I_{N}Z_{2N} = V_{2}(z_{2})$$

$$\vdots \qquad \vdots \qquad \vdots \qquad \vdots \qquad \vdots$$

$$I_{1}Z_{N1} + I_{2}Z_{N2} + \dots + I_{N}Z_{NN} = V_{N}(z_{N})$$

$$(2.24)$$

where $Z_{mn} \triangleq Z_n(z_m)$. This can be written in matrix form as

$$\begin{bmatrix} Z_{11} & Z_{12} & \cdots & Z_{1N} \\ Z_{21} & Z_{22} & \cdots & Z_{2N} \\ \vdots & \vdots & & \vdots \\ Z_{N1} & Z_{N2} & \cdots & Z_{NN} \end{bmatrix} \begin{bmatrix} I_1 \\ I_2 \\ \vdots \\ I_N \end{bmatrix} = \begin{bmatrix} V_1(z_1) \\ V_2(z_2) \\ \vdots \\ V_N(z_N) \end{bmatrix}$$
(2.25)

or in even more compact notation as

$$\mathbf{ZI} = \mathbf{V},\tag{2.26}$$

where \mathbf{Z} , \mathbf{I} and \mathbf{V} are referred to as the impedance matrix, current vector, and excitation vector, respectively. \mathbf{Z} depends only on the geometry and the basis functions $F_n(z)$; \mathbf{V} depends on the excitation used in the problem, and is usually known. Therefore, $\mathbf{I} = \mathbf{Z}^{-1}\mathbf{V}$, and the radiated fields can be computed from these currents.

This process of expanding the unknown current I(z) into a series of basis functions and then generating N equations in N unknowns is commonly referred to in electromagnetics literature as the method of moments [31]. Several codes are available that implement MoM. In this dissertation, we use the NEC2 [35] implementation of MoM, because of our familiarity with it.

A reflector antenna can be modeled using a wire grid or patches for MoM analysis. In this dissertation, we employ wire grid models, because wire grids are appropriate for modeling open surfaces in NEC2 (via the Electric Field Integral Equation (EFIE) [35]). Patches, on the other hand, are appropriate for modeling closed surfaces in NEC2 (via the Magnetic Field Integral Equation [35]).

The successful modeling of a continuous metallic surface using a wire grid depends on the grid spacing (smallest distance between two parallel wires in the grid) d and the radius of the wire r. However, there is a trade-off between computational burden and accuracy in selecting the grid spacing. The computational effort is mostly dominated by the inversion of the matrix **Z**, and is approximately proportional (may vary depending on the algorithm used for the inversion) to N^3 , where N is the number of segments. For example, the computational time may increase by a factor of 64 if the resolution of the grid is doubled; i.e., the number of segments is quadrupled. Therefore it is important to select a grid size which strikes an appropriate balance between accuracy and the associated computational burden. A common approach to select the radius of the wires in the grid is to apply the "Equal Area Rule" [36], which states that the sum of the surface area of the parallel wires in one direction should be approximately equal to that of the solid surface replaced by the grid; i.e., a square patch of side Δ should be modeled using four wires having radius $r = \frac{\Delta}{2\pi}$. However, in this dissertation, the grid spacing and the wire radius for modeling surfaces are selected based on a simple experiment so that we have a better understanding of the effect of the parameters, and which generates results consistent with the Equal Area Rule. The experiment will now be briefly described here.

In this experiment, a wire grid model of a 26 m-diameter flat disk is used as a ground plane, and a half-wave dipole is located one quarter wavelength above this grid at 74 MHz.

The grid spacing and wire radius are then varied to achieve directivity and pattern which is comparable to that observed when the grid is replaced by a perfect solid ground plane. The details of the experiment are described in Appendix B. The experiment shows that a wire grid with spacing of 36.5 cm (about 0.1λ at 74 MHz) and wire radius of 45 mm (about 0.01λ at 74 MHz) sufficiently models the solid surface of the disk. Note that in this case, $2\pi r = 28$ cm which is very close to the selected grid size of 36.5 cm; therefore, the radius of the wires selected for the wire grid model is consistent with that suggested by the Equal Area Rule.

A reflector surface having diameter D = 26 m and consisting of the wire grid described in the previous paragraph is shown in Figure 2.5. The model results in 7,272 segments. In this model, the reflector is centered around the negative z-axis with its perimeter parallel to the xy-plane, and the focus of the reflector lies at the origin.

In order to check the validity of the reflector model, we separately analyze the reflector with MoM (using the wire grid model) and aperture theory (assuming continuous PEC surface, as is usual), and compare the results. In this study, the reflector is fed from the focus with a half–wave dipole feed at frequencies below 100 MHz; specifically, at 10 MHz, 20 MHz, 38 MHz, 74 MHz, and 88 MHz. In order to do the analysis using aperture theory, we first assume a uniform illumination (as opposed to the actual non–uniform illumination by the dipole feed) of the reflector, and estimate the uniform illumination gain (G_u) and half-power beamwidth (θ_u) at the frequencies of interest as described in [31]. This assumes an aperture efficiency ϵ_{ap} value of 1; however, the actual aperture efficiency (when the dipole feed is considered) is less than 1, and depends on the taper efficiency ϵ_a to be 1). ϵ_t is 90%, and ϵ_s is 45% for a half–wave dipole feed located at the focus. This gives $\epsilon_{ap} = \epsilon_t \epsilon_s$ to be 40%. We use this aperture efficiency to estimate the gain ($G = \epsilon_{ap}G_u$) for the actual (non-uniform) illumination, and compare these results with gain (G_{MoM}) and half-power beamwidth (θ_{MoM}) computed using MoM. Table 2.1 summarizes these results. Note the reasonable agreement



Figure 2.5. Wire grid model of a 26 m-diameter reflector surface.

between gain and half-power beamwidth, except at 10 MHz. At 10 MHz, the reflector diameter is 0.87λ . Therefore we expect the aperture theory-based estimate to be inaccurate there, but probably reasonable at higher frequencies. This experiment gives us confidence in the wire-grid model of the reflector that we created; so now we can compare PO and UTD results against MoM to check their validity at low frequencies.

2.4 Comparison of MoM, PO and UTD at Low Frequencies

In this section, we establish a low frequency limit on the usefulness of PO and UTD. For this purpose, we compare PO and UTD results with MoM results. Specifically, a comparison

Freq	D/λ	Aperture Theory			MoM	
		G_u	$G = \epsilon_{ap}G_u$	θ_u	G_{MoM}	$ heta_{MoM}$
10 MHz	0.87	8.7 dB	4.7 dB	67.4°	7.4 dB	100.0°
20 MHz	1.73	14.7 dB	10.7 dB	33.7°	10.6 dB	32.0°
38 MHz	3.30	20.3 dB	16.3 dB	17.7°	15.9 dB	20.0°
74 MHz	6.42	26.1 dB	22.1 dB	9.1°	20.3 dB	8.0°
88 MHz	7.62	27.6 dB	23.6 dB	7.7°	22.1 dB	8.0°

Table 2.1. Summary of reflector antenna analysis using aperture theory and
MoM.

is made between the results of MoM, PO, and UTD analysis of an axisymmetric reflector antenna (D = 26 m, f/D = 0.42) fed with a half-wave dipole located at the focus at frequencies below 100 MHz – namely, 10 MHz, 20 MHz, 38 MHz, 74 MHz, and 88 MHz.

The reflector antenna model used for MoM analysis is the same as described in the previous section. The model is initially designed at 74 MHz, but it satisfies the NEC2 modeling guidelines at least up to 110 MHz. Therefore, we use the same model for all the frequencies considered in this study. The model has 7,272 segments, and for a single frequency it takes about 20 minutes to compute the pattern at 180 points on a SX2800 Gateway computer (2.33 GHz Intel core 2 quad processor, 4GB RAM). The PO model used here, as described in Section 2.1, uses 16,405 patches, and requires 16 seconds to compute the far-field at 180 points for a single frequency. The UTD model uses 2 rim diffractions, and yields results in about 0.7 seconds. We compare the results obtained from these three methods, and determine the range of frequencies and pattern angles for which the approximate methods (PO and UTD) properly work.

Results are summarized in Figures 2.6 through 2.16. It is observed that the combination of PO for main lobe pattern and UTD for sidelobe pattern provides results which are comparable to MoM results at frequencies above 38 MHz. As frequency is lowered, the agreement with MoM results starts to diminish.

Figure 2.16 compares the estimation of main lobe directivity by MoM (assumed accurate) and PO (assumed less accurate, but faster to compute) as a function of frequency. UTD



Figure 2.6. E-plane pattern at 88 MHz using MoM, PO and UTD.

Figure 2.7. H-plane pattern at 88 MHz using MoM, PO and UTD.

is not considered in this case since it does not produce valid results for the main lobe. It is observed that PO is able to estimate the main lobe directivity obtained with MoM to within 1 dB above 35 MHz. Moreover, we see that both methods show some oscillation in directivity with frequency. This is expected as the reflected field from the dish interferes with the direct radiation from the feed. However, the oscillation estimated by PO is large compared to the MoM result, and the results converge with increasing frequency.

In Chapters 4–7, PO will be used to estimate the directivity of reflector antennas at frequencies where MoM analysis is not feasible due to the required wire grid model having too many segments (>11,000) to fit in commonly available computer memory (e.g., 4GB of RAM). Specifically, for a 25 m-diameter reflector, PO is used for frequencies in the range 110–500 MHz. MoM, on the other hand, is used in order to analyze reflector antennas at the lower end of our frequency range of interest, specifically below 88 MHz.

2.5 MoM Model of an EVLA Reflector Antenna

In order to accurately estimate the performance of a reflector antenna using MoM, we need a sufficient model, i.e., one that incorporates all the relevant structural components. Such a complete model for an EVLA reflector antenna is developed in this section. The



Figure 2.8. E-plane pattern at 74 MHz using MoM, PO and UTD.



Figure 2.9. H-plane pattern at 74 MHz using MoM, PO and UTD.

major parts of an EVLA reflector antenna are: main reflector, subreflector, struts, support wires, subreflector mount, and high-frequency feed assembly. Figure 2.17 shows a drawing of these components (except the main reflector) along with their dimensions.

As in the reflector model described in Section 2.3, the main reflector of an EVLA antenna is modeled as a wire grid with a grid spacing of 36.5 cm and wire radius of 45 mm. In this model the main reflector is centered around the negative z-axis of the coordinate system with its rim parallel to the xy-plane, and the prime focus coincides with the origin of the coordinate system. We will refer to this coordinate system in the rest of this dissertation for identifying different locations in the antenna model.

The shape of the Cassegrain subreflector on an EVLA antenna is not known to us in detail. However, a drawing of the subreflector is made available to us by Jim Ruff of NRAO. Data points are extracted from one cross-section of the subreflector in the drawing, and a polynomial equation is derived that fits the data within a residual (difference between the actual data point and the predicted value from the polynomial equation) of 0.01λ at 74 MHz. The details are given in Appendix C. Using the polynomial equation a wire grid model of the subreflector is created, and is shown in Figure 2.18. The same grid spacing as the main reflector is not used in this case, as this spacing is very large for modeling a subreflector with



Figure 2.10. E-plane pattern at 38 MHz using MoM, PO and UTD.



Figure 2.12. E-plane pattern at 20 MHz using MoM, PO and UTD.



Figure 2.11. H-plane pattern at 38 MHz using MoM, PO and UTD.



Figure 2.13. H-plane pattern at 20 MHz using MoM, PO and UTD.

a diameter of 2.35 m, and results in large "quantization error" in the rim of the subreflector. The resulting diameter is 0.04λ less than the original diameter at 74 MHz. Therefore, we used a smaller grid spacing of d = 13 cm (corresponding to 0.03λ at 74 MHz) to reduce the error, and the resulting diameter is only 0.005λ less than the original diameter. The radius of wire in the grid is selected to be 45 mm as before. Also, in the complete antenna model the subreflector is not symmetric about the axis of the reflector; it is slightly tilted. This is because in the actual EVLA antenna the subreflector is tilted in order to point toward the ring of high frequency feeds.



10 MHz using MoM, PO and UTD.

Figure 2.15. H-plane pattern at 10 MHz using MoM, PO and UTD.

The struts are modeled as follows. The largest dimension w of the cross-section of a strut is 40 cm, which at 74 MHz is 0.1λ . We could model the strut as a single wire, where the radius of the wire a is related to w as, a = 0.25w [31]. However, in our case the radius of the wire would be 10 cm, which is larger than $\lambda/(20\pi)$ at 74 MHz, and therefore not appropriate according to the modeling guidelines of NEC2. Instead, we model the strut as two wires with radii of 3 cm, placed 40 cm apart, with connecting segments every 50 cm. The radius of the wires is selected to be 3 cm, based on the Equal Area Rule [36].

The four support wires are modeled using single wires with radius of 1 cm and segment lengths of $\lambda/10$ at 74 MHz.

The subreflector mount is modeled as a wire grid cylinder as shown in Figure 2.19. The top and bottom surfaces of the cylinder are modeled using concentric rings separated by a distance of 0.40 m (λ /10 at 74 MHz). Radial wires with segment lengths of 0.40 m (λ /10 at 74 MHz) are then created to connect the concentric rings. As for the cylindrical surface two more circular rings are created in between the top and the bottom surfaces, and these rings are connected to the top and the bottom surfaces; they are also connected with each other using vertical wires with segment lengths of 0.40 m (λ /10 at 74 MHz). One issue with the model is that the radial wires in the inner most circles of the top and bottom surfaces are



Figure 2.16. Directivity of a 26 m-diameter reflector antenna, fed from the focus with a half-wave dipole, computed using PO and MoM.

connected at acute angles; also, the radius of these wires is about three times less than that of the wires connecting the innermost circles to the outer circles. This may result in loss of accuracy [35]. However, the subreflector mount is mostly shadowed by the subreflector, and therefore this model is not expected to reduce the accuracy of MoM results significantly. Improvement of this model is identified as a future work.

Another important component in an EVLA antenna is the high-frequency feed assembly located at the vertex of the main reflector. As is visible in Figure 1.2, the high-frequency feed assembly is a very complex structure, and thus very difficult to model. We model the feed assembly as a wire grid rectangular box encapsulating its volume. The grid spacing and wire radius of this rectangular box is the same as that used for the main reflector.



Figure 2.17. Detail of the EVLA reflector geometry, showing the subreflector, the subreflector mount, the struts, the support wires, and the feed assembly. The drawing is provided by Jim Ruff of NRAO.

Figures 2.20 shows our model of the EVLA antenna with all the components included. The wire grid model has a total number of 9,062 segments. The model is designed at 74 MHz due to our special interest in the EVLA 4 m system. However, this model satisfies NEC2 modeling guidelines for frequencies at least up to 110 MHz. Therefore, we will use this very model for all the frequencies below 110 MHz. A single frequency simulation of this model, yielding pattern at 180 points, takes about 30 minutes on a 2.33 GHz Intel Core 2 Quad, 4GB RAM machine.

Now that a model with all the components of an EVLA antenna is available, the effects of these various components on directivity and pattern are assessed through an experiment. Those components which do not have much effect might be excluded from the complete reflector antenna model, and thereby the computational burden might be reduced.



Figure 2.18. Wire grid model of an EVLA subreflector.

In this experiment, we started with a reflector-only configuration, and added to this configuration are the various components one by one. The directivity and the pattern for each of these configurations is computed when the reflector is fed with several resonant dipole feeds, namely dipoles resonant at 88 MHz, 74 MHz, 38 MHz and 29 MHz; one at a time. The radius of these dipoles is selected to be 2.38 mm to be comparable to the dipole feeds on the existing EVLA 4 m system, and the lengths of these dipoles are as shown in Table 2.2. In configurations where the subreflector is not present, the feed antenna is located at the true focus. For configurations which include the subreflector, the feed is located at the image of the focus, i.e., the dipole is located the same distance in front of the subreflector as the prime focus is behind it. The lowest frequency in this experiment is selected to be



Figure 2.19. Wire grid model of the subreflector mount.

29 MHz, as the 29 MHz-resonant dipole is the longest dipole that can be mounted between the opposite struts of an EVLA antenna when the dipole is located at the image of the prime focus. Again, due to space constraints, the 29 MHz- and 38 MHz-resonant dipoles cannot be mounted between the adjacent struts when they are located at the focus, and therefore they are not included in the main reflector+strut+mount configuration of this experiment. The results of this experiment are shown in Figures 2.21 through 2.24, where the co-polarized E-plane pattern for the different configurations of the antenna are plotted.

Our findings are summarized in Table 2.3. It is observed that the directivity increases as the struts and the mount are added to the reflector-only configuration. The addition of the subreflector to the main reflector-only configuration also increases directivity at 74 MHz and 88 MHz, but reduces directivity at 29 MHz and 38 MHz. On the other hand, the high-frequency feed assembly reduces the directivity at 88 MHz, 74 MHz and 38 MHz, but increases the directivity at 29 MHz. However, when all the components are present, the directivity is increased from the reflector-only configuration at all frequencies considered.



Figure 2.20. Wire grid model of an EVLA antenna including main reflector, subreflector, struts, support wires, mount, and feed assembly.

Thus, all the components of the reflector antenna appear to be important. Therefore, this last, most-detailed model is used in subsequent work in this dissertation.

2.6 Summary

This chapter addressed electromagnetic analysis techniques for reflector antennas. Upon comparison of the three techniques, namely PO, UTD, and MoM, we can make the following conclusions. MoM gives accurate results, but is computationally expensive. PO on the other hand, is fast, but is good only for the main lobe and first few sidelobes. In order to complement PO results with far-out sidelobes, UTD can be used. As an example, it was demonstrated that for a 26 m-diameter reflector antenna, the combination of PO and UTD

Frequency	Dipole Length
$29 \mathrm{~MHz}$	4.96 m
$38 \mathrm{~MHz}$	3.78 m
$74 \mathrm{~MHz}$	1.94 m
88 MHz	1.62 m

Table 2.2. The lengths of the feed dipoles considered in Section 2.5.

offers comparable results to MoM at frequencies above approximately 38 MHz. Generalizing, PO and UTD should not be assumed valid for reflectors smaller than about 3.25λ in diameter.

A MoM model of an EVLA antenna was developed. The effects of the various components (e.g., the subreflector, the mount, the struts, the high frequency feed assembly) on antenna directivity and pattern were considered, and was observed that all the components make significant contribution to the results. The final model has 9,062 segments, and it requires about 2415 seconds to compute the pattern of the antenna (on a 2 GHz Pentium processor) at 180 points using this model.



Figure 2.21. E-plane co–pol pattern at 29 MHz for different configurations.



Figure 2.22. E-plane co-pol pattern at 38 MHz for different configurations.



Figure 2.23. E-plane co-pol pattern at 74 MHz for different configurations.



Figure 2.24. E-plane co-pol pattern at 88 MHz for different configurations.

Components	main ref.	main ref.	main ref.	main ref.	main ref.	main ref.
		+strut	+subref.	+subref.	+feed assy.	+ subref.
		+ mount		+mount		+ mount
				+struts		+struts
						+feed assy.
No. of	7231	7955	7837	8558	7735	9062
Segments						
Run Time	1313 s	$1700~{\rm s}$	1615 s	2079 s	$1674~\mathrm{s}$	2415 s
Freq	Dir.	Dir.	Dir.	Dir.	Dir.	Dir.
29 MHz	13.8 dB	NA	13.7 dB	15.7 dB	14.2 dB	16.5 dB
38 MHz	13.8 dB	NA	13.3 dB	15.9 dB	$11.3 \mathrm{~dB}$	14.4 dB
74 MHz	20.7 dB	21.9 dB	23.3 dB	22.2 dB	20.0 dB	21.3 dB
88 MHz	21.8 dB	24.3 dB	24.7 dB	23.9 dB	21.6 dB	23.6 dB

Table 2.3. Results of analysis for different configurations. Dir. refers to the directivity of the antenna. [Note. The computation times reported are based on a 2 GHz Pentium M processor and 1 GB RAM machine.]

Chapter 3

Analysis of Sensitivity of Large Reflector Antennas at Low Frequencies

In this chapter, we describe the analysis of sensitivity of a large reflector antenna. The sensitivity of a radio telescope is defined as the weakest level of radio emission that can be detected by it. In this dissertation, sensitivity is expressed in terms of SEFD, and is used to compare the performance of different low frequency systems. SEFD is defined as the incident flux spectral density which yields a signal-to-noise ratio (SNR) equal to unity at the output of the system.

In Section 3.1, a methodology for the analysis of sensitivity of a reflector antenna employing a single feed element is described. This is the most common form of feeding for a reflector antenna. A system model of the signal path from the incident signal through the receiver is developed. Using this model, an expression is formulated for the sensitivity in terms of SEFD. An example of this sensitivity analysis methodology is then presented in Section 3.2. Specifically, the analysis of sensitivity is presented for a simple axisymmetric reflector antenna (D = 25 m, f/D=0.36) when fed from the prime focus with a half-wave dipole feed. Section 3.3 addresses the effects of the various system model parameters on the sensitivity of a system. Also, we investigate how much of the desired range of 10–500 MHz can be covered with a useful level of sensitivity using a minimum number of simple dipole feeds. In Section 3.4, a different methodology for analysis of sensitivity, useful for reflector antennas employing array feeds, is described. At low frequencies, the correlation of the external noise between the array elements can desensitize the system, and therefore a different treatment for the array feed is required to estimate sensitivity. Low frequency systems consisting of dipole array feeds are proposed in Chapter 6 of this dissertation, and therefore this formulation becomes necessary in analyzing these systems. This methodology is verified in Section 3.5 by comparing the analysis results of a single feed system with that obtained with the method of Section 3.1. An example of sensitivity analysis of a reflector antenna employing an array feed is then presented in Section 3.6.

3.1 System Model and Sensitivity Analysis of a Reflector Antenna System Consisting of a Single Feed Element

In this section, we describe the analysis of sensitivity of a reflector antenna system which employs a single feed element. The system can be modeled as shown in Figure 3.1 as: 1) a feed; 2) a passive balun (since most of the feed antennas that will be investigated are balanced, and a passive balun is desired to connect them to a coaxial (unbalanced) feedline); 3) feedline connecting the balun to a preamplifier; 4) a preamplifier; and 5) feedline connecting the preamplifier to the receiver.

The reflector antenna can be described in terms of its noise temperature T_A . T_A can be found by integrating over the external (sky) noise brightness temperature distribution $T_b(\theta, \phi)$, weighted by the normalized power pattern of the reflector antenna $P(\theta, \phi)$, as follows:

$$T_A = \frac{1}{\Omega_A} \int_0^{\pi} \int_0^{2\pi} T_b(\theta, \phi) P(\theta, \phi) d\Omega , \qquad (3.1)$$



Figure 3.1. System model of the signal path from the incident signal through the receiver.

where θ is zenith angle, ϕ is azimuth angle and Ω_A is the beam solid angle of the reflector antenna. Ω_A is defined as

$$\Omega_A = \iint_{4\pi} P(\theta, \phi) d\Omega .$$
(3.2)

 $P(\theta, \phi)$ can be computed from an electromagnetic analysis of the reflector antenna as described in Chapter 2. $T_b(\theta, \phi)$ can be obtained from existing astronomical surveys (e.g., [37]) or sky models (e.g., [38], [39]).

Let the balun in the system model be described in terms of its input impedance Z_b and gain G_b . Also, let the first and second feedlines be described in terms of their gains G_{f1} and G_{f2} . G_b , G_{f1} and G_{f2} are defined such that each of these has a maximum value of 1 corresponding to the case where there is no loss. The physical temperature of the components is T_e . Furthermore, the preamplifier is described in terms of its gain G_p and input-referred noise temperature T_p .

The self-impedance of the feed antenna is denoted as Z_A . When Z_A is conjugate matched to the balun impedance $(Z_A = Z_b^*)$, the power spectral density (PSD) S_1 delivered to the balun due to a source of solid angle Ω_s is

$$S_1 = \frac{1}{2} A_e \iint_{source} I_v(\theta, \phi) P(\theta, \phi) d\Omega \qquad [W \text{ Hz}^{-1}], \qquad (3.3)$$

where A_e is the effective aperture of the reflector antenna and $I_v(\theta, \phi)$ is the intensity of the source (having units of W m⁻² Hz⁻¹ sr⁻¹). The factor of 1/2 accounts for the fact that if the signal is unpolarized, any single polarization captures about half of the available power. Moreover, if the source is of sufficiently small extent, so that $P(\theta, \phi) \approx 1$, and I_v can be assumed constant over the source, we have

$$S_1 = \frac{1}{2} A_e I_v \Omega_s. \tag{3.4}$$

where $\Omega_s = \iint_{source} d\Omega << \Omega_A.$

The effective aperture of the reflector antenna under conjugate matching is

$$A_e = \frac{\lambda^2}{4\pi} G , \qquad (3.5)$$

where λ is the wavelength and $G = e_r D_A$ is the gain of the antenna, with D_A being the directivity and e_r being the radiation efficiency. D_A can be determined from an electromagnetic analysis of the reflector antenna. e_r is assumed to be 1; i.e., the antenna is assumed lossless, and it is also assumed that there is no ground loss. The assumption of no ground loss is verified for pointing directions close to the zenith through a simple study. The total power in the far field of an EVLA reflector antenna is computed at 74 MHz (using MoM) when it is fed from the image of the focus with a 74 MHz half–wave dipole. The reflector antenna is pointed toward the zenith in this case. The study is separately performed for the following cases: 1) reflector antenna's vertex is 16 m (4 λ at 74 MHz) above a perfect ground, and 2) reflector antenna's vertex is 16 m above a realistic lossy ground (conductivity $\sigma = 5 \times 10^{-3}$ S/m and relative permittivity $\epsilon_r = 13$). The height of the reflector above ground is selected to be 16 m as the vertex of the EVLA is about this high above the ground in practice while pointing toward the zenith. In transmit mode, the difference in the total power in the far field for these two cases is observed to be less than 1%. The reason behind this is the fact that the feed dipole is mostly shielded by the main reflector when pointing toward zenith; if the antenna pointing direction is changed to higher zenith angles, ground loss may not be neglected; e.g., when the above study is repeated for antenna pointing direction of 60° zenith angle, the total power in the far field is observed to be reduced by about 16% for the realistic ground (compared to the perfect ground) scenario. However, in the work presented in this dissertation, we will assume smaller zenith angles, and negligible ground loss.

An impedance mismatch between frequency-variable Z_A and the nominally frequencyfixed Z_b reduces the effective aperture at the input of the balun to $e_I A_e$, where e_I is the impedance mismatch efficiency (IME), given by

$$e_I = 1 - |\Gamma|^2 . (3.6)$$

 Γ is the voltage reflection coefficient given as

$$\Gamma = \frac{Z_b - Z_A}{Z_b + Z_A} \,. \tag{3.7}$$

The PSD associated with the source (alone) at the output of the second feedline is then

$$S_2 = \frac{1}{2} A_e I_v \Omega_s e_I G_b G_{f1} G_p G_{f2} . aga{3.8}$$

The PSD associated with the noise contributions at the output of the second feedline is

$$N = k(e_{I}T_{A}G_{b}G_{f1}G_{p}G_{f2} + T_{e}[1 - G_{b}]G_{f1}G_{p}G_{f2} + T_{e}[1 - G_{f1}]G_{p}G_{f2} + T_{p}G_{p}G_{f2} + T_{e}[1 - G_{f2}]).$$
(3.9)

The terms on the right side of Equation 3.9 are the contributions from external noise (T_A) ,

balun, first feedline, preamplifier (T_p) , and second feedline, respectively. Now the signal to noise ratio (SNR) at the output of the second feedline is S_2/N , which after some algebraic manipulation becomes

$$SNR = \frac{e_I A_e I_v \Omega_s}{2k(e_I T_A + T_{int})} , \qquad (3.10)$$

where T_{int} is defined as the input-referred internal noise temperature of the system, and is given by

$$T_{int} = T_e \left[\frac{(1 - G_b)}{G_b} + \frac{(1 - G_{f1})}{G_b G_{f1}} + \frac{(1 - G_{f2})}{G_b G_{f1} G_p G_{f2}} \right] + \frac{T_p}{G_b G_{f1}} .$$
(3.11)

The value of $I_v \Omega_s$ (i.e., source flux) that yields a SNR value of 1 is the SEFD, which can now be expressed as

$$SEFD = \frac{2k(e_I T_A + T_{int})}{e_I A_e} . \qquad (3.12)$$

This metric can be used to quantify the sensitivity of an antenna system. Note that if $e_I T_A >> T_{int}$; i.e., the external noise after impedance mismatch is much greater than the input-referred internal noise of the system, then SEFD $\approx 2kT_A/A_e$. This implies that under this condition, sensitivity can be improved only by increasing the effective aperture. This is a valuable insight, as it infers that large feed antenna impedance mismatch may be accommodated without significant loss of sensitivity under external noise–dominated operation. As a result, simple dipole–like antennas (having intrinsically narrow impedance bandwidth) can offer large usable bandwidth from a sensitivity perspective. However, the frequency range over which external noise can dominate the internal noise depends on the various system model parameters. The effects of the various system model parameters on the level of external noise dominance are discussed in Section 3.3. In the next section, an example of the analysis methodology described here is presented.

3.2 Example: Analysis of Sensitivity of a Reflector Antenna Fed with a Single Feed Element

In this section, the analysis of sensitivity for a prime focus-fed axisymmetric reflector antenna with D = 25 m and f/D = 0.36 is presented. The reflector antenna is fed from the focus with dipoles half-wavelength long at 29 MHz, 38 MHz, 47 MHz, 60 MHz, and 74 MHz; one at a time. The radius of these dipoles is 2.38 mm, and their lengths are as shown in Table 3.1. In this study, T_{int} and Z_b are assumed to be 250 K and 100 Ω respectively, which resembles existing dipole-based low frequency systems operating in this range; e.g., LWA. The sensitivity of each system is analyzed at 10 MHz, 20 MHz, 38 MHz, 74 MHz and 88 MHz.

NEC2 (in contrast to PO or UTD) is used in this study. The reflector antenna model used in this analysis consists of the wire grid developed in Section 2.3 having d = 36.5 cm and r = 45 mm. The dipoles are modeled as single wires, having 11 segments.

 Z_A is obtained from the calculated current at the feed point, and D_A is found by integrating over the far field pattern $P(\theta, \phi)$. Using these, A_e (assuming $e_r = 1$) and e_I are computed from Equations 3.5 and 3.6, respectively.

 T_A is calculated by evaluating Equation 3.1, which requires values for $T_b(\theta, \phi)$ and $P(\theta, \phi)$. $T_b(\theta, \phi)$ is found at 38 MHz using a sky brightness temperature map generated using LFmap [38], and is extrapolated to other frequencies simply by scaling the temperatures by a factor of $(v/38 \text{ MHz})^{-2.55}$, where v is the frequency of interest in MHz. The spectral index of -2.55 is commonly used for extrapolating brightness temperatures from one frequency to another [40, 41]. $P(\theta, \phi)$ depends on the pointing direction of the antenna. In this study, two pointing directions are considered: 1) A point (right ascension (RA): 19 h, declination (DEC): -25° , $T_b = 199 \times 10^3$ K at 38 MHz) in Sagittarius, and 2) a point (RA: 12.5 h, DEC: 13.5°, $T_b = 7000$ K at 38 MHz) in Virgo. The two distinct pointing directions are selected to demonstrate the performance of the reflector antenna at two extreme cases of sky temperature (Sagittarius is "hot", and Virgo is "cold").

Frequency	Dipole Length
$29 \mathrm{~MHz}$	4.97 m
38 MHz	3.79 m
$47 \mathrm{~MHz}$	3.06 m
60 MHz	2.39 m
$74 \mathrm{~MHz}$	1.94 m
88 MHz	1.62 m

Table 3.1. Lengths of the feed dipoles in the study described in Section 3.2.

In order to compute T_A using Equation 3.1, a computer program developed by Kshitjia Deshpande of Virginia Tech [42] is used. The results of T_A calculation are provided in Appendix D.

The results are shown in Figures 3.2 and 3.3. It is observed that SEFD for pointing toward Virgo is lower (i.e., better) than SEFD for pointing toward Sagittarius, as expected. Note that several of the dipoles appear to show comparable performance in the range 38–74 MHz, even though they are nominally "narrowband" feeds. In fact, the overall downward (improving) trend in SEFD with increasing frequency is determined primarily by the sky brightness temperature, as opposed to the dipole's resonant frequency. This behavior again highlights the fact that at low frequencies, even dipole feeds can offer large usable bandwidth from a sensitivity perspective.

3.3 Prospects for Broad Low Frequency Coverage Using Single Dipole Feeds

In this section, we want to identify how well the desired range of 10–500 MHz can be covered with a minimum number of dipoles. To serve this purpose we do a simple study where the external noise captured by dipole antennas is compared to the internal noise of the electronics connected to them. Three dipoles, resonant at 74 MHz, 150 MHz and 327 MHz, are selected for this study. The frequencies are selected as they are in bands that are allocated for radio astronomy [43]. The radius of these dipoles is selected to be


Figure 3.2. Results of sensitivity analysis for a 25 m-diameter reflector antenna pointing toward Sagittarius. Note: HWDP stands for half-wave dipole.

2.38 mm, 3.37 mm and 4.76 mm respectively. The radius of 2.38 mm is selected for the 74 MHz–resonant dipole to be comparable to the existing EVLA 4 m system. For the other two dipoles, the radius is increased to improve their bandwidth characteristics. The radius is made 1.4 times larger each time the frequency is doubled, so that the three resonant dipoles have almost equal mass (implying equal installation effort). The lengths of these dipoles are shown in Table 3.2.

In this section, the external noise spectral density kT_Ae_I delivered to a load of 100 Ω connected across the dipole terminals is calculated, and compared to the internal noise of the system referred to the dipole terminals. T_A is computed using Equation 3.1 assuming



Figure 3.3. Results of sensitivity analysis for a 25 m-diameter reflector antenna pointing toward Virgo. Note: HWDP stands for half-wave dipole.

 $T_b(\psi)$ to be uniform over the sky ($\theta < \pi/2$) and zero for $\theta > \pi/2$. Through the Rayleigh-Jeans law, T_b can be expressed as

$$T_b = \frac{1}{2k} I_v^{sky} \lambda^2 , \qquad (3.13)$$

where I_v^{sky} is the sky brightness intensity (having units of W m⁻² Hz⁻¹ sr⁻¹). I_v^{sky} can be obtained using a simplified version of Cane's empirical model for the spectrum of the external noise background [44], as explained in [30]. In this model,

$$I_v^{sky} \approx I_g v^{-0.52} + I_{eg} v^{-0.80}$$
, (3.14)

Resonant Frequency	Radius	Dipole Length
74 MHz	$2.38 \mathrm{~mm}$	1.94 m
150 MHz	$3.37 \mathrm{~mm}$	0.94 m
327 MHz	4.76 mm	0.42 m

Table 3.2. Lengths and radii of the feed dipoles considered in Section 3.3.

where $I_g = 2.48 \times 10^{-20}$, $I_{eg} = 1.06 \times 10^{-20}$ and v is frequency in MHz. In practice, I_v^{sky} , and consequentially $T_b(\psi)$, vary considerably both as a function of ψ and as a function of time of the day, as already described in Section 3.1. However, the advantage of using this simplified model is that it provides a reasonable standard condition for comparing different low frequency systems. e_I is calculated using Equations 3.6 and 3.7 with the impedance of the dipoles Z_A obtained from MoM analysis.

In computing the internal noise, we assume $G_b = -1$ dB and $G_p = 36$ dB (comparable to LWA). G_{f1} and G_{f2} are calculated assuming 10 m and 1 m long RG–58 cables, respectively. The length of 10 m approximately equals the length of the cable that carries the signals from the existing EVLA 4 m system to the room beneath the main reflector vertex. The length of the second feedline for the EVLA 4 m system is not known to us; however, it is assumed to be 1 m.

A number of T_p values, namely, 60 K, 100 K, 215 K and 250 K are considered one at a time for this study. These noise temperatures are selected to be comparable to commonly available low noise amplifiers operable in our desired frequency range¹.

The system described above is henceforth referred to as Scheme 1. The results of our analysis are shown in Figure 3.4. It is observed that T_p primarily defines the bandwidth over which external noise can dominate the internal noise; and secondarily the sensitivity of the system. The system noise for the 74 MHz resonant dipole is found to be external noise dominated by at least 3 dB in the range 57–86 MHz for a T_p value of 250 K. This range is wider for alternative designs with lower T_p values. However, the compression point of such

¹Mini–Circuits PMA–5456, PSA–5454, LEE–39 and GALI–74 offer T_p values of 60 K, 100 K, 215 K and 250 K, respectively.



Figure 3.4. Comparing the power density at the dipole terminals due to external noise and internal noise.

designs is suspected to be correspondingly lower, which increases the risk of being driven into compression by RFI². As for the 150 MHz–resonant dipoles, the external noise is greater than the internal noise by at least 3 dB only in the range of 139–147 MHz for a T_p value of 60 K; however, as mentioned, the low value of $T_p = 60$ K may result in unacceptable linearity of the system. The external noise delivered by the 327 MHz dipole, on the other hand, is much weaker than the internal noise for all the T_p values in the entire region investigated.

The ratio of external noise to internal noise δ determines the required observation ("integration") time: For a given source flux and internal noise of the system, decreasing δ

 $^{^{2}}$ A detailed analysis of linearity for low frequency systems can be found in [45].

increases the integration time required to achieve a specified sensitivity. For example, integration time is increased by 57% and 21% over the ideal (zero internal noise) values for δ of 6 dB and 10 dB, respectively [46].

The system design of Scheme 1 cannot offer external noise-dominated system temperature over large ranges mostly due to the noise associated with the first feedline. This is especially true for the systems with 150 MHz– and 327 MHz–resonant dipoles. An alternative design is therefore considered. The preamplifier ("active balun") is directly connected to the dipole, and the cable between them is eliminated. This design is referred to as Scheme 2. In terms of our model parameters, it is the same design as Scheme 1, but with $G_{f1} = 1$, and G_{f2} now corresponds to 10 m RG–58 cable. We repeated the study described above for Scheme 2, and the results are shown in Figure 3.4. It is observed that even for a T_p value of 250 K the external noise dominates the internal noise at least by 3 dB over the range 55-88 MHz. Also, for the system with 150 MHz–resonant dipoles, external noise can dominate the internal noise by 3 dB in the range 125-165 MHz when $T_p = 100$ K is used. On the other hand, the system noise for the system with 327 MHz–resonant dipoles is still completely defined by the internal noise in the intended frequency range of operation regardless of T_p . However, this is not completely useless, since the integration time required to achieve a specified sensitivity is still reduced compared to Scheme 1. In fact, the system temperature of the existing EVLA P-band receiver is about 130 K; thus the result presented here represent practical limits.

Another important system parameter that effects the degree to which external noise can dominate the internal noise of the system is the impedance mismatch between the dipole (Z_A) and the rest of the system (R_L) . It has been shown in [30] that at the expense of reduced peak IME, the frequency range of operation can be broadened; in other words, a higher value of R_L may reduce the degree by which external noise dominates internal noise, but it might extend the range over which external noise dominates. However, this is only of interest when the system sensitivity is external noise–dominated. Once internal noise starts to dominate, a higher value of R_L only decreases sensitivity. The above argument suggests the use of a higher value for R_L than 100 Ω ; at least for the 74 MHz system. However, from an implementation point of view, a R_L of 100 Ω is very easy to obtain as opposed to higher values; e.g., two 50 Ω low noise amplifiers differentially connected to the dipole terminal easily give a R_L of 100 Ω as described in [41]. Pros and cons of higher values of R_L require a comprehensive design study, and therefore this is identified as a possible area of future work. As for the work presented in this dissertation, a R_L value of 100 Ω is used.

Based on the discussion presented in this section, it is observed that three resonant dipoles (74 MHz, 150 MHz and 327 MHz) are able to cover at least the frequency range of 50–470 MHz with a useful level of sensitivity. The intended frequency coverage with the individual dipoles would be 50-88 MHz for the 74 MHz–resonant dipole ($\delta \geq 3$ dB), 110-250 MHz for the 150 MHz–resonant dipole ($\delta \geq 1$ dB in the range 110–180 MHz) and 250-470 MHz for the 327 MHz–resonant dipole. Inability to cover the range of 88–110 MHz is perhaps not a significant problem, since observation in the range 88-108 MHz is particularly very difficult due to the presence of the FM broadcast band.

3.4 Analysis of Sensitivity of a Reflector Antenna Fed with an Array Feed

In this section, we describe a method for analyzing the sensitivity of a reflector antenna that employs an array feed. This analysis is useful, as we consider arrays of dipoles as low frequency feeds in Chapter 6. The analysis methodology presented here is different from that presented in Section 3.1. In an array environment, the correlation of sky noise between array elements can be significant. This can have major effect on the sensitivity of the system [47,48]; especially at low frequencies, when the external noise of the system due to the large sky noise can dominate the internal noise of the system as demonstrated in the previous section. The methodology for analyzing sensitivity described in Section 3.1 does not take the correlation of external noise into consideration, and therefore is not suitable for analyzing systems consisting of array feeds. Moreover, the mutual coupling between the elements of the array feed may also be important, as it can have significant effect on the gain and pattern of the reflector antenna system, and therefore needs to be considered.

A different approach to estimate sensitivity is therefore described in this section. This method closely follows the methodology described in [48], and is able to account for correlation of sky noise between array elements and also, the mutual coupling between them.

The system model assumes an array of N feed antennas whose signals are combined. Let $E_{\theta}(t)$ and $E_{\phi}(t)$ be the θ - and ϕ -polarized components of the electric field associated with the signal of our interest, incident from a direction of $\{\theta_0, \phi_0\}$, having units of V m⁻¹ Hz^{-1/2}. The direction $\{\theta_0, \phi_0\}$ will be expressed as ψ_0 henceforth. The resulting voltage at the n^{th} feed antenna terminal, having units of V Hz^{-1/2}, is

$$x_n(t) = a_n^{\theta}(\psi_0) E_{\theta}(t) + a_n^{\phi}(\psi_0) E_{\phi}(t) + z_n(t) + u_n(t) , \qquad (3.15)$$

where $a_n^{\theta}(\psi_0)$ and $a_n^{\phi}(\psi_0)$ are the effective lengths, having units of meters, associated with the θ and ϕ polarizations respectively, for the n^{th} feed antenna for a plane wave incident from ψ_0 ; $z_n(t)$ is the contribution from noise external to the system; and $u_n(t)$ is the contribution from noise internal to the system. When the signals from all the antennas are combined, the output can be expressed as:

$$y(t) = \sum_{n=1}^{N} b_n x_n(t) , \qquad (3.16)$$

where the unitless b_n 's are the combining coefficients.

Assuming root-mean-square voltages, the power after combining the signals from different feed antennas is

$$P_y = \langle y(t)y^*(t) \rangle R_0^{-1} , \qquad (3.17)$$

where $\langle . \rangle$ denotes time-domain averaging, "*" denotes conjugation, and R_0 is the impedance looking into the system as seen from the terminals across which y(t) is measured.

We will now expand Equation 3.17 with the assumptions that the signal of interest, $z_n(t)$, and $u_n(t)$ are mutually uncorrelated for any given n, i.e., for any n and m

$$\langle E_{\theta}(t)z_{n}^{*}(t) \rangle = \langle E_{\phi}(t)z_{n}^{*}(t) \rangle = 0$$
, (3.18)

$$< E_{\theta}(t)u_n^*(t) > = < E_{\phi}(t)u_n^*(t) > = 0$$
, (3.19)

$$\langle z_n(t)u_m^*(t) \rangle = 0$$
. (3.20)

The correlation of external noise between antennas is not precluded by the above assumption since $\langle z_n(t)z_m^*(t) \rangle$ can be $\neq 0$ for $n \neq m$. Furthermore no assumption has been made about the correlation between $E_{\theta}(t)$ and $E_{\phi}(t)$. Upon expanding Equation 3.17 we get

$$P_{y}R_{0} = \mathbf{b}^{H}\mathbf{A}_{\theta\theta}\mathbf{b}P_{\theta\theta} + \mathbf{b}^{H}\mathbf{A}_{\phi\phi}\mathbf{b}P_{\phi\phi}$$
$$+\mathbf{b}^{H}\mathbf{A}_{\theta\phi}\mathbf{b}P_{\theta\phi} + \mathbf{b}^{H}\mathbf{A}_{\phi\theta}\mathbf{b}P_{\phi\theta}$$
$$+\mathbf{b}^{H}\mathbf{P}_{z}\mathbf{b} + \mathbf{b}^{H}\mathbf{P}_{u}\mathbf{b} \qquad (3.21)$$

where " H " denotes the conjugate transpose operator;

$$\mathbf{b} = [b_1 \ b_2 \ \dots \ b_N]^T , \qquad (3.22)$$

where " T " denotes the transpose operator; and

$$\mathbf{A}_{\theta\theta} = \mathbf{a}_{\theta}^{*}(\psi_{0})\mathbf{a}_{\theta}^{T}(\psi_{0}) \quad P_{\theta\theta} = \langle |E_{\theta}(t)|^{2} \rangle \quad , \tag{3.23}$$

$$\mathbf{A}_{\phi\phi} = \mathbf{a}_{\phi}^{*}(\psi_{0})\mathbf{a}_{\phi}^{T}(\psi_{0}) \quad P_{\phi\phi} = \langle |E_{\phi}(t)|^{2} \rangle \quad , \tag{3.24}$$

$$\mathbf{A}_{\theta\phi} = \mathbf{a}_{\phi}^{*}(\psi_{0})\mathbf{a}_{\theta}^{T}(\psi_{0}) \quad P_{\theta\phi} = \langle E_{\theta}(t)E_{\phi}^{*}(t) \rangle \quad , \qquad (3.25)$$

$$\mathbf{A}_{\phi\theta} = \mathbf{a}_{\theta}^{*}(\psi_{0})\mathbf{a}_{\phi}^{T}(\psi_{0}) \quad P_{\phi\theta} = \langle E_{\phi}(t)E_{\theta}(t)^{*} \rangle \quad , \qquad (3.26)$$

$$\mathbf{a}_{\theta}(\psi_0) = [a_1^{\theta}(\psi_0) \ a_2^{\theta}(\psi_0) \ \dots \ a_N^{\theta}(\psi_0)]^T$$
, and (3.27)

$$\mathbf{a}_{\phi}(\psi_0) = [a_1^{\phi}(\psi_0) \ a_2^{\phi}(\psi_0) \ \dots \ a_N^{\phi}(\psi_0)]^T ; \qquad (3.28)$$

 \mathbf{P}_{z} is the external noise covariance matrix whose $(n, m)^{th}$ element is $\langle z_{n}^{*}(t)z_{m}(t) \rangle$, and \mathbf{P}_{u} is the internal noise covariance matrix whose $(n, m)^{th}$ element is $\langle u_{n}^{*}(t)u_{m}(t) \rangle$.

In this work, \mathbf{P}_u is computed assuming that the internal noise of one feed antenna is not correlated with the internal noise of another feed antenna. Therefore, \mathbf{P}_u becomes a diagonal matrix whose non-zero elements are:

$$\mathbf{P}_{u}^{[n]} = kT_{int,n}R_L \tag{3.29}$$

where $T_{int,n}$ is the input-referred internal noise temperature associated with the n^{th} antenna, and R_L is the impedance seen looking out the feed antenna terminal.

We now derive a simple expression for the $(n, m)^{th}$ element of \mathbf{P}_z , $P_z^{[n,m]}$. The power in the two polarizations of the electric field associated with the unpolarized external noise incident from a region of solid angle $\Delta\Omega$ around ψ can be expressed as

$$< |\Delta E_{\theta}(\psi, t)|^{2} > = < |\Delta E_{\phi}(\psi, t)|^{2} > = \frac{\eta}{2} \Delta S(\psi) ,$$
 (3.30)

where η is the impedance of free space and $\Delta S(\psi)$ is the associated flux density having units of W m⁻² Hz⁻¹. $\Delta S(\psi)$ can also be obtained from the Rayleigh-Jeans Law:

$$\Delta S(\psi) = \frac{2k}{\lambda^2} T_b(\psi) \Delta \Omega \tag{3.31}$$

where $T_b(\psi)$ is the sky brightness temperature in the direction ψ , and λ is the wavelength under consideration. We can thus model $\Delta E_{\theta}(\psi, t)$ and $\Delta E_{\phi}(\psi, t)$ as follows:

$$\Delta E_{\theta}(\psi, t) = g_{\theta}(\psi, t) \sqrt{\frac{k\eta}{\lambda^2} T_b(\psi) \Delta \Omega}$$
(3.32)

$$\Delta E_{\phi}(\psi, t) = g_{\phi}(\psi, t) \sqrt{\frac{k\eta}{\lambda^2} T_b(\psi) \Delta \Omega}$$
(3.33)

where $g_{\theta}(\psi, t)$ and $g_{\phi}(\psi, t)$ are Gaussian-distributed independent random variables with zero

mean and unit variance. Now, in order to obtain $z_n(t)$ we sum the contributions received over a sphere:

$$z_n(t) = \sum_{\psi} [a_n^{\theta}(\psi) \Delta E_{\theta}(\psi, t) + a_n^{\phi}(\psi) \Delta E_{\phi}(\psi, t)] . \qquad (3.34)$$

Applying the definition of \mathbf{P}_z and exploiting the statistical properties of $g_{\theta}(\psi, t)$ and $g_{\phi}(\psi, t)$ we find

$$\mathbf{P}_{z}^{[n,m]} = \frac{k\eta}{\lambda^{2}} \sum_{\psi} [a_{n}^{\theta*}(\psi)a_{m}^{\theta}(\psi) + a_{n}^{\phi*}(\psi)a_{m}^{\phi}(\psi)]T_{b}(\psi)\Delta\Omega , \qquad (3.35)$$

which can now be written in integral form:

$$\mathbf{P}_{z}^{[n,m]} = \frac{k\eta}{\lambda^{2}} \int [a_{n}^{\theta*}(\psi)a_{m}^{\theta}(\psi) + a_{n}^{\phi*}(\psi)a_{m}^{\phi}(\psi)]T_{b}(\psi)d\Omega . \qquad (3.36)$$

For an unpolarized signal of interest $P_{\theta\phi} = P_{\phi\theta} = 0$ and $P_{\theta\theta} = P_{\phi\phi} = \eta S(\psi)/2$, where $S(\psi)$ is the flux density associated with the signal. Returning to Equation 3.21 we obtain the following expression for the signal to noise ratio (SNR) for an unpolarized signal of interest:

$$SNR = \frac{\eta}{2} S(\psi) \frac{\mathbf{b}^{H} (\mathbf{A}_{\theta\theta} + \mathbf{A}_{\phi\phi}) \mathbf{b}}{\mathbf{b}^{H} (\mathbf{P}_{z} + \mathbf{P}_{u}) \mathbf{b}}$$
(3.37)

Now, regardless of the choice of **b**, the sensitivity of the antenna system can be expressed in terms of SEFD, which is the value of $S(\psi)$ that gives an SNR of 1; i.e.,

$$SEFD = \frac{2}{\eta} \frac{\mathbf{b}^{H}(\mathbf{P}_{z} + \mathbf{P}_{u})\mathbf{b}}{\mathbf{b}^{H}(\mathbf{A}_{\theta\theta} + \mathbf{A}_{\phi\phi})\mathbf{b}}.$$
(3.38)

Note this expression gives SEFD directly in terms of the feed elements' effective lengths (i.e., patterns), external and internal noise powers, and the combining coefficients. The maximum possible SNR is equal to the maximum eigenvalue of $\mathbf{R_n}^{-1}\mathbf{R_s}$ [49], where

$$\mathbf{R}_n = \mathbf{P}_z + \mathbf{P}_u$$
, and (3.39)

$$\mathbf{R}_s = \mathbf{A}_{\theta\theta} P_{\theta\theta} + \mathbf{A}_{\phi\phi} P_{\phi\phi} . \qquad (3.40)$$

The maximum SNR is achieved by selecting **b** to be the eigenvector corresponding to the maximum eigenvalue of $\mathbf{R_n}^{-1}\mathbf{R_s}$. The analysis methodology described here is compared against the method of Section 3.1 in the next section.

3.5 Comparison of the Methods of Sections 3.1 and 3.4 for Analysis of a Single Dipole Feed System

In order to verify the methodology presented in Section 3.4, we apply this method to analyze the sensitivity of the same reflector antenna as described in Section 3.2 when fed with a single 74 MHz–resonant dipole of radius of 2.38 mm, located at the focus. The results of this analysis are compared to those computed using the method described in Section 3.1.

In order to compute the effective length of a dipole feed, the system is illuminated with a θ -polarized 1 V/m plane wave incident from some direction ψ , and the resulting current I_L across a series resistance R_L at the feed terminal is determined using MoM. $a_{\theta}(\psi)$ for the feed antenna is then simply the resulting voltage across the dipole terminals, i.e., $I_L R_L$. The process is repeated for a ϕ -polarized plane wave and iterated over ψ .

 \mathbf{P}_u is computed using Equation 3.29, assuming $R_L = 100 \ \Omega$ and $T_{int} = 250 \ \text{K}$. \mathbf{P}_z is computed using Equation 3.35 where $T_b(\psi)$ is found (using Equation 3.13) as explained in Section 3.3. Using this model, $T_b(\psi)$ toward the sky is found to be 4835 K, 3035 K, 1777 K, and 1142 K at 50 MHz, 60 MHz, 74 MHz and 88 MHz, respectively. The actual contributions to system temperature are less due to the impedance mismatch between the antenna self-impedance and R_L . This mismatch is automatically taken into account in the way we determine the effective lengths of the feed antennas using MoM.

The results are shown in Figure 3.5. SEFD is calculated for several frequencies below 100 MHz, namely 50 MHz, 60 MHz, 70 MHz, 74 MHz, 80 MHz, 88 MHz and 88 MHz. It is observed that the SEFD estimates of the two methods agree to within < 0.5 dB in the range 50–88 MHz.



Figure 3.5. SEFD for a 25 m-diameter reflector antenna system fed with a 74 MHz resonant dipole from focus.

Note that as a part of the analysis methodology presented in Section 3.4, the effective aperture $e_I A_e$ of the antenna can be derived. If the power delivered to the load (R_L) is P_L , then $P_L = \frac{1}{2} |I_L|^2 R_L$. Also, $P_L = \frac{1}{2} S_d(\psi) e_I A_e(\psi)$ (only half of the total power is available in one polarization for an unpolarized signal of interest) where $S_d(\psi)$ is the power density having units of W m⁻². Since $S_d(\psi) = 2|E_{\theta}^i(\psi)|^2/2\eta = 2|E_{\phi}^i(\psi)|^2/2\eta$, the effective aperture is:

$$e_I A_e(\psi) = \eta \frac{|I_L|^2}{|E^i_{\theta,\phi}(\psi)|^2} R_L .$$
(3.41)

Using this method we calculated $e_I A_e$ for the reflector antenna system from the above example, and the results are found to be 10.2 m², 150.0 m², and 76.6 m² at 50 MHz, 74 MHz and 88 MHz respectively. On the other hand, $e_I A_e$ computed at these frequencies using Equation 3.5 and 3.6 are found to be 10.1 m², 149.1 m² and 75.9 m², which are in good agreement to the aforementioned values. This gives us confidence in the analysis methodology of Section 3.4. We now study the utility of this method in analyzing reflector antenna systems consisting of array feeds.

3.6 Example: Analysis of Sensitivity of a Reflector Antenna Fed with an Array Feed

An application of the methodology presented in Section 3.4 for analyzing the sensitivity of reflector antennas employing array feeds is demonstrated in this section. We now analyze the sensitivity of the same reflector antenna as described in Section 3.2, when fed with an array feed. The array feed consists of 74 MHz-resonant dipoles having radius of 2.38 mm, arranged in a "boxing ring" configuration around the focus, in the focal plane (see Figure 6.3). The dipole centers are 1 m away from the focus; this is the smallest distance from the focus where the dipole ring can fit. This particular array feed is the first example considered in this dissertation of a class of low-frequency feed arrangement – namely, the distributed feed array. An entire chapter, Chapter 6, is dedicated to describe this class of feeds. As for now, the boxing ring of dipoles around focus is selected to demonstrate an example of estimation of sensitivity for array feeds.

The analysis is performed in the same manner as described in the previous section. The results are shown in Figure 3.6. SEFD estimates for the system are calculated at several frequencies below 100 MHz, namely 50 MHz, 55 MHz, 60 MHz, 70 MHz, 74 MHz, 80 MHz, and 88 MHz. It is observed from Figure 3.6 that the lowest SEFD (i.e., the best sensitivity) occurs at 74 MHz (the resonant frequency of the dipole), as expected. In the range about 60–80 MHz SEFD is within 1 dB of this value. However, at frequencies beyond this range SEFD starts to increase rapidly. The reason for this is the fact that the



Figure 3.6. SEFD for a 25 m-diameter reflector antenna system, fed with an array of 74 MHz resonant dipoles employed around the focus in the focal plane.

impedance mismatch worsens with change in frequency from the resonant frequency of the dipole, and as a result the level of external noise dominance decreases; internal noise of the system then starts to define the sensitivity. Also note that as frequency is increased from the resonant frequency, the electrical separation of the dipoles from the focus increases, and consequentially performance worsens.

The ratio of $Tr\{\mathbf{P}_z\}/Tr\{\mathbf{P}_u\}$ (where "Tr" denotes the trace operation; i.e., the sum of the diagonal terms.) expresses the degree by which external noise dominates the internal noise at the output of the system, and is shown for the frequencies considered in this study in

the inset of Figure 3.6. It can be observed that indeed the level of external noise dominance is decreased as frequency is changed from the resonant frequency.

In order to demonstrate the importance of including spatial correlation, we again estimated the sensitivity of this distributed feed array system, but ignoring the correlation of external noise between array elements. The off-diagonal elements of the external noise covariance matrix $\mathbf{P}_{\mathbf{z}}$ are forced to 0 to accomplish this. The results are shown in Figure 3.6. It is observed that the SEFD appears to be better (smaller) when the correlation of external noise is neglected; specifically, in the range 50–74 MHz SEFD appears to be better by as much as 0.5 dB. Therefore it is important to consider the correlation of external noise between array elements to avoid over–estimating the sensitivity of the low frequency systems.

3.7 Summary

In this chapter, methods for analysis of sensitivity of reflector antenna systems are described, and examples of analysis are presented. A set of three dipoles resonant at 74 MHz, 150 MHz and 327 MHz, is identified to be useful at least in the range 50–470 MHz. The specifications of the system design – namely, the input referred internal noise temperature T_{int} and impedance seen looking out the dipole terminals R_L are also identified to ensure a usable level of performance. R_L is selected to be 100 Ω ; T_{int} is selected to be 250 K for the system with the 74 MHz–resonant dipoles, and 100 K for the 150 MHz– and 327 MHz–resonant dipoles.

A method for sensitivity analysis of reflector antenna systems consisting of array feeds is described. It was shown based on the analysis of an array–fed reflector antenna that the correlation of external noise amongst array elements can desensitize a system at low frequencies, and therefore needs to be considered while estimating sensitivity. Applying the analysis methodology described in this chapter for array feeds, the existing EVLA low frequency systems, namely the 4 m system and the 90 cm system are analyzed in the next chapter.

Chapter 4

Analysis of Existing Low Frequency Systems on the EVLA

In this chapter, the performance of the existing low frequency systems on the EVLA, namely the 4 m system and the 90 cm system, are analyzed. The EVLA is selected to demonstrate the application of the research presented in this dissertation; i.e., modification of large reflector antennas for low frequency operation. It is useful to characterize the performance of this system as it currently exists, before proposing replacements or modifications as is done in the following chapters.

The required background to carry out the study in this chapter, specifically, electromagnetic analysis techniques and sensitivity analysis methodology, are already described in Chapters 2 and 3, respectively. Based upon this knowledge, in Sections 4.1 and 4.2 the EVLA 4 m system and the EVLA 90 cm system are analyzed, respectively; such analysis has not been previously reported to the best of our knowledge.

4.1 Existing 4 m System on the EVLA

In this section, we estimate the performance of the existing EVLA 4 m system. A brief overview of this system has already been provided in Sections 1.1 and 2.5. The feed for this system consists of crossed dipoles of radius 2.38 mm, located at the image of the prime focus, i.e., the dipoles are the same distance in front of the subreflector vertex as the prime focus is behind it. Figure 4.1 shows this arrangement for better understanding.

The first part of our analysis of this system consists of directivity and pattern calculations at 74 MHz. The support wires may be oriented in two different ways with respect to a dipole feed, as shown in Figure 4.2, and have different impacts on the directivity and the pattern. One of these orientations is when the wires appear to be aligned with the dipole when a bottom view of the antenna (looking along the reflector axis toward the subreflector) is considered; this will be referred to as the co-aligned orientation (Figure 4.2(a)). The other orientation is when the wires appear to be cross-aligned with the dipole when a bottom view of the antenna is considered; this will be referred to as the cross-aligned orientation (Figure 4.2(b)). For this reason, only one of the crossed dipole feeds is considered at a time. For both the dipole orientations, we analyzed the directivity and the pattern of an EVLA antenna using MoM. For this study, the EVLA antenna model developed in Section 2.5 is used.

The results are shown in Figure 4.3 for the two possible orientations of the support wires. First, we consider the performance of the EVLA antenna when fed with a dipole which is co-aligned with the support wires (shown in Figure 4.3(a)). Note that the directivity (D_A) at 74 MHz is calculated to be 20.4 dB, which corresponds to an aperture efficiency ϵ_{ap} of 29%. ϵ_{ap} is calculated as the ratio of the effective aperture A_e ($A_e = D_A \lambda^2 / 4\pi$) and the physical aperture $(\pi D^2/4)$ of the antenna. The directivity of the EVLA antenna when fed with a cross-aligned dipole feed (Figure 4.3(b)) is found to be 16.2 dB at 74 MHz, i.e., 4 dB worse, corresponding to $\epsilon_{ap} = 11\%$. This is apparently due to an unfavorable interaction with the support wires in this configuration. This difference between the co- and cross-aligned dipole feeds has been observed in practice, but has been difficult to quantify [50]. It is interesting to note that our estimates of 29% and 11% are consistent with the estimates of 25% and 15%, reported in [3] and [17], respectively; although it is not clear if the latter pertains to



Figure 4.1. Feed dipole located at the image of the prime focus with respect to the subreflector. Tilt of the subreflector with respect to the main reflector axis is 2.4° .

the different feed dipole orientations, as considered here.

Also visible in Figure 4.3 is the fact that the pattern of the antenna system has slight asymmetry. This is because the subreflector is tilted as described previously.

The next part of our analysis comprises of estimation of the sensitivity of the system. The sensitivity of the EVLA 4 m system is estimated in the frequency range of 50–88 MHz using MoM as described in Section 3.4. As before, the series load connected across the dipole terminals R_L is assumed to be 100 Ω , and the total internal noise temperature referred to the dipole terminals T_{int} is assumed to be 250 K. Furthermore, a uniform sky brightness temperature T_b is assumed, and is computed as described in Section 3.3. The beamforming coefficients are selected to maximize SNR (as described in Section 3.4).

Based on these assumptions, the sensitivity of the system is analyzed, and the results are shown in Figure 4.4. It is observed that SEFD increases (i.e., becomes worse) below 70 MHz. Again, the reason for this is the fact that the impedance mismatch starts to



Figure 4.2. Orientation of the support wires in the EVLA antenna model with respect to a single (a) co-aligned, and (b) cross-aligned dipole feed.

become so bad that the system no longer remains external-noise limited, and the internal noise starts to define the sensitivity. The ratio of $Tr\{\mathbf{P}_z\}/Tr\{\mathbf{P}_u\}$ is shown in the inset of Figure 4.4 to indicate the degree to which the external noise dominates the internal noise of the system at each frequency. Note that the SEFD doesn't rapidly get worse (i.e., increase) as frequency is increased above 74 MHz. This behavior can be explained using Equation 3.12 and Figure 4.5. From Equation 3.12, SEFD is the ratio of $2k(e_I T_A + T_{int})$ and $e_I A_e$. It is observed from Figure 4.5 that if T_{int} is assumed to be 0, SEFD decreases monotonically with increasing frequency; since T_A decreases with frequency and A_e increases with frequency. On the other hand, if T_A is assumed to be 0, SEFD varies with increasing frequency in the same manner as $1/e_I A_e$ (Figure 4.5(b)); since T_{int} doesn't change with frequency. Therefore it can be summarized that SEFD doesn't get rapidly worse with increasing frequency because decreasing T_A and increasing A_e dominate over decreasing e_I .

One major concern with the 4 m system is that it reduces the sensitivity of the L-band system by about 6% [17]. As a result the dipole feeds are only intermittently mounted. An alternative feed design – namely, the distributed feed array – is therefore proposed in



Figure 4.3. Calculated E–plane co–pol directivity of the EVLA 4 m system for (a) co-aligned support wires, and (b) cross-aligned support wires.

Chapter 6. This new system demonstrates comparable performance to the existing system, but with reduced blockage presented to the higher–frequency systems.

4.2 Existing 90 cm System on the EVLA

In this section, we analyze the existing 90 cm system on the EVLA. The feed for the 90 cm system consists of crossed dipoles of radius of 2.38 mm, and they are located 0.23 m ($\lambda/4$ at 327 MHz) in front of the subreflector vertex. The subreflector serves as a crude backplane. The resulting phase center is about one-half wavelength displaced from the prime focus, resulting in relatively low aperture efficiency (40%) [16]; nevertheless, the system achieves a useful level of performance over the frequency range of 300–340 MHz (12.5% bandwidth), and has been scientifically productive.

The study of the 90 cm system is divided into two parts. In Section 4.2.1 the pattern and the effective aperture of the antenna is calculated, and in Section 4.2.2 the sensitivity is estimated.



Figure 4.4. Estimated SEFD for the existing EVLA 4 m system. Also, shown in inset is the degree to which external noise dominates the internal noise.

4.2.1 Pattern and Effective Aperture

The first part of the analysis presented in this section includes the estimation of aperture efficiency. In the frequency range of interest, a MoM analysis is not feasible, as it requires a wire grid model which will have too many segments (115,696) to fit in commonly-available computing memory (e.g., 4 GB of RAM). Therefore we employ a "hybrid" method to do the analysis. In order to compute the reflector antenna directivity, first, we obtain the radiated field from the feed ignoring the reflector, using MoM. Everything except the reflector and the high–frequency feed assembly in the EVLA antenna model – namely, the struts, the subreflector mount, the support wires, the subreflector and the feeds – are included in this part of the analysis. Also, the wire grid models of these components have been modified



Figure 4.5. SEFD for different noise conditions, and reduced effective aperture for the EVLA 4 m system in the range 50–170 MHz.

from that developed in Chapter 2 to meet the NEC2 modeling requirements at the higher frequencies. The lengths and the radius of the wire segments in the previous model are scaled with frequency, such that the new model is valid at least up to 470 MHz. Once the radiated field is obtained using MoM, the resulting PO current on the reflector surface is computed. These currents are then radiated to find the far-field. The direct radiated field from the feed is added to the PO reflected far-field, and the directivity is computed from the resulting total field. Note that in this method the possibility of a standing wave due to multiple reflections between the feed and the reflector is not considered. A standing wave would cause the the gain of the reflector antenna to exhibit ripple as a function of frequency. However, in the work presented in this dissertation, our interest is on sensitivity over large bandwidth, and the variation due to the standing wave effect are small compared to variation in sensitivity over large bandwidth.

In order to check the validity of this method, we compared the results obtained with this method with the results from a complete MoM analysis of a reflector antenna at 50 MHz, 74 MHz, 80 MHz, 88 MHz, 100 MHz and 110 MHz; the argument being that if the hybrid method gives reasonable results at these low frequencies, then it should perform better as frequency is increased. A half-wave dipole located at the image of the prime focus of an EVLA antenna is used in this study. The results are shown in Figure 4.6. It is found that the directivity obtained with the hybrid method agrees within 0.5 dB with the "direct" MoM method at all the frequencies considered. Therefore, it seems reasonable to use the hybrid method at those frequencies of our interest at which using MoM is not feasible from a computing–hardware requirement point of view.

Like the 4 m system, the estimation of aperture efficiency of the EVLA 90 cm system has been performed for both orientations of a feed dipole with respect to the support wires. The results are shown in Figure 4.7. It is found that unlike the 4 m system, the orientation of the dipoles with respect to the support wires does not have any significant effect. The analysis indicates directivities of 36.0 dB and 36.1 dB for the co-aligned and cross-aligned orientations, respectively, corresponding to an aperture efficiency of about 54%. This estimation of efficiency is somewhat higher than the reported efficiency of 40% [16] for the EVLA 90 cm system. The difference might be attributed to the interaction of the fields radiated from the



Figure 4.6. Comparison of directivity calculated using the MoM method and the hybrid method as a function of frequency

reflector with the structures present in front of it; this interaction is not accounted for in computing the directivity using our hybrid method. However, field measurement of A_e is very difficult and it could simply be that measurement error is large [51].

4.2.2 Sensitivity and Bandwidth

The second part of our study in this section includes the broadband performance analysis of the EVLA 90 cm system from a sensitivity perspective; specifically, in the range of 250-470 MHz. The analysis methodology developed in Section 3.4 is used for this purpose. However, computing the effective lengths $a_n^{\theta,\phi}(\psi)$ of the feed elements is not as straightforward as for the 4 m system. To calculate the effective lengths of the dipole feeds of the 4 m system, the EVLA antenna model (developed in Section 2.5) is illuminated with a 1 V/m (peak)



Figure 4.7. Calculated E–plane co–pol directivity of the EVLA 90 cm system for (a) co-aligned support wires, and (b) cross-aligned support wires.

plane wave, the resulting current I_L in the load R_L is calculated using MoM, and the effective length is then found to be $I_L R_L$. However, as mentioned, the antenna model needed for doing the same analysis in the range 250–470 MHz requires too many segments (about 115,696) to fit in commonly-available computing memory. Therefore an alternative methodology is used in this section to compute the effective lengths (i.e., $I_L R_L$ in response to a 1 V/m plane wave illumination) of the dipole feeds. This method is referred to as "MoM+PO" in this dissertation, and is described here.

Before describing the detailed steps associated with the MoM+PO method, the concept behind this methodology is introduced. Let's identify the time average power P_L delivered to R_L , in response to field incident from ψ direction. If the peak current is I_L , then

$$P_L(\psi) = \frac{|I_L(\psi)|^2 R_L}{2} .$$
(4.1)

The power can also be expressed as

$$P_L(\psi) = e_I(\psi) A_e(\psi) S_d(\psi) , \qquad (4.2)$$

where e_I is the IME, $A_e(\psi)$ is the effective aperture, and $S_d(\psi)$ is the incident power density from the direction ψ having units of W m⁻². The incident power density can be expressed as $S_d = |E_i(\psi)|^2/2\eta$, where $E_i(\psi)$ is the peak incident electric field (having units of V/m) from ψ direction and η is the intrinsic impedance of free space. From Equation 4.1 and 4.2 we can find that

$$|I_L(\psi)|R_L = \sqrt{\frac{e_I(\psi)A_e(\psi)|E_i(\psi)|^2 R_L}{\eta}} .$$
(4.3)

If $|E_i(\psi)| = 1$ V/m,

$$|I_L(\psi)|R_L = \sqrt{\frac{e_I(\psi)A_e(\psi)R_L}{\eta}} .$$
(4.4)

Thus, we can find the magnitude of the effective length, $|I_L(\psi)|R_L$, from Equation 4.4; however, we have to devise a way to find $e_I(\psi)$ and $A_e(\psi)$. As for the phase of the effective length, we can compute this by illuminating the antenna geometry excluding the main reflector and the feed assembly with a plane wave incident from ψ direction, and find the resulting current across R_L . The phase of the current γ is then taken to be the phase of the effective length of the corresponding feed. Again, this method of computing the phase of the effective length is verified by comparison with the results from a MoM analysis of a model that includes the main reflector and the feed assembly. The result is shown in Figure 4.8. It is observed that the error in phase calculation is not significant (less than 2.6%), and therefore the method for calculating the phase is reasonable. Improvement in accuracy in the phase calculation is considered as a part of future work.

The first concern in computing IME e_I is that how much effect the reflector has on the impedance of a feed dipole. If the effect is negligible, then the impedance of the feed dipole can easily be determined using a MoM analysis which excludes the reflector in the model. To explore this possibility, we did a study where the impedance of a feed dipole located at the image of the focus of an EVLA antenna is computed both in the presence and in the absence of the reflector at 50 MHz, 74 MHz, 100 MHz and 110 MHz. The concept being that if the



Figure 4.8. Error in calculation of effective length's phase when the main reflector is not considered.

error is small at 74 MHz, we expect it to be even smaller at higher frequencies (specifically, at frequencies where we cannot do a MoM analysis of the reflector), because of the increased electrical distance from the reflector. The open circuit voltage, V_{oc} (approximated as the voltage across a 10 M Ω resistor), and the short circuit current, I_{sc} , across the feed terminals in response to a 1 V/m plane wave incidence is obtained from a MoM simulation both with and without the reflector being present in the model. The dipole impedance is then V_{oc}/I_{sc} . The IME is then calculated for a 100 Ω load, and the results are shown in Figure 4.9. It is observed that less than 2% error is introduced into the IME calculation by neglecting the main reflector. Therefore, it seems reasonable to compute IME without considering the reflector at higher frequencies where MoM analysis of the reflector is not practical.



Figure 4.9. Error in calculation of IME when the main reflector is excluded from the MoM analysis.

The effective aperture A_e can be computed as $A_e = D_A \lambda^2 / 4\pi$, where D_A is the directivity of the reflector antenna. D_A can be obtained using the same method as described in the previous section. To make sure that there is no significant difference in the directivity computed in the transmit case and in the receive case, a simple study is performed. A 25 m-diameter reflector antenna is fed from the focus with a 74 MHz resonant dipole. The directivity of the reflector antenna D_A is computed in transmit mode using MoM, and is found to be 20.7 dB. The directivity is then computed in the receive case. A conjugatematched load is connected across the dipole terminals, and the current I_L in this load in response to a 1 V/m (peak) plane wave (co-polarized) illumination of the reflector antenna is found using MoM. A_e is then calculated using Equation 4.4. In this case, $D_A = A_e 4\pi/\lambda^2$ is found to be 20.8 dB which is within 0.1 dB of the result for the transmit case. Summarizing, the MoM+PO method used to calculate the sensitivity of the EVLA 90 cm system is as follows:

- 1. Obtain feed pattern using MoM with all the antenna components present in the model except for the main reflector.
- 2. Using the feed pattern compute the PO currents on the reflector surface and radiate these currents to compute the far fields.
- 3. Compute $A_e(\psi)$ using Equation 3.5.
- 4. Compute $Z_A(\psi)$ (i.e., V_{oc}/I_{sc}) using MoM without the reflector in the model. Using $Z_A(\psi)$ compute $e_I(\psi)$ from Equations 3.6 and 3.7.
- 5. Using MoM calculate the phase γ of the current I_L across the series resistance R_L in the absence of the main reflector in the model.
- 6. Compute $|I_L|R_L$ using Equation 4.3.
- 7. Compute $|I_L|R_L e^{j\gamma}$ to be the effective length.
- 8. Once the effective lengths of the array elements are obtained, compute $\mathbf{A}_{\theta\theta}$, $\mathbf{A}_{\phi\phi}$, $\mathbf{P}_{\mathbf{u}}$ and $\mathbf{P}_{\mathbf{z}}$ from Equations 3.23, 3.24, 3.29 and 3.35, respectively.
- 9. Compute the beamforming coefficients b_n 's using Equations 3.39 and 3.40.
- 10. SEFD can now be computed using Equation 3.38.

In order to further validate this method, a simple study is performed. The performance of a simple 25 m-diameter reflector antenna, fed from the focus with a 74 MHz-resonant dipole, is estimated using the MoM+PO method, and the results are compared with the results obtained directly with MoM (as described in Section 4.1). The results are shown in Table 4.1. It is observed that the SEFD obtained with the two methods agree within 1 dB. We expect the agreement to be better as frequency is increased (errors in the intermediate

Freq.	SEFD using MoM	SEFD using MoM+PO	Error	
			%	dB
50 MHz	147.9 kJy	125.9 kJy	-15%	$-0.7 \mathrm{dB}$
74 MHz	32.4 kJy	41.7 kJy	+29%	+1.0 dB
88 MHz	27.5 kJy	22.9 kJy	-17%	-0.8 dB

Table 4.1. SEFD estimates for a simple 25 m-diameter reflector antenna, fed from the focus with a 74 MHz-resonant dipole, using the (direct) MoM method and the MoM+PO method.

steps of the MoM+PO method are expected to decrease with increasing frequency), which gives confidence in the use of this method for the EVLA 90 cm system.

We now estimate the sensitivity of the EVLA 90 cm system using the MoM+PO method. Again, R_L and T_{int} are assumed to be 100 Ω and 100 K, respectively. $T_b(\psi)$ is assumed to be uniform as described in Section 3.3. Results are shown in Figure 4.10. Note the broad minimum around 327 MHz (SEFD varies by less than 0.5 dB in the range 290–340 MHz). Also shown in Figure 4.10 is the ratio of $Tr\{\mathbf{P}_z\}/Tr\{\mathbf{P}_u\}$ at selected frequencies. Since the system noise is not external noise–dominated anywhere in the range 250-470 MHz, the effect of the decrease in IME with change in frequency from the resonant frequency has a greater impact on the estimated SEFD than in the 74 MHz (external noise–dominated) scenario.

4.3 Summary

In this chapter, the existing low frequency systems of the EVLA – namely, the 4 m system and the 90 cm system – are analyzed, and the sensitivity of these systems is estimated. The results are summarized in Figures 4.4 and 4.10, respectively. These results are consistent with the best available information about EVLA sensitivity; however, lack of precise measurement data or the absence of some other independent check (e.g., another EM modeling method) means results should be interpreted with caution.



Figure 4.10. Estimated SEFD for the existing EVLA 90 cm system. Also, shown in inset is the degree to which external noise dominates the internal noise.

Chapter 5

Focus- or Near Focus-Fed Systems

In this chapter, we investigate low frequency reflector antenna systems consisting of crossed dipole feeds located at the focus or near the focus (in cases where the focus is not accessible) of reflector antennas. A preliminary study has been performed in Section 3.3 (based on the system model shown in Figure 3.1) where the external noise delivered by simple dipoles to the electronics connected across their terminals is computed, and compared to the internal noise of these electronics. A set of three dipoles, resonant at 74 MHz, 150 MHz and 327 MHz (see Table 3.2), was selected for the study. The study offered insight about the usefulness of dipole antennas in covering the frequency range of 50–470 MHz. In this chapter, we estimate the performance when these dipoles are employed at the focus to feed reflector antennas. This is described in detail in Section 5.1. The objective is to find out how well the desired frequency range can be covered using these systems. This establishes a baseline of performance against which new designs can be evaluated.

In Section 5.2, the study of Section 5.1 is applied to the EVLA. The existing EVLA 4 m system and 90 cm system consist of dipole feeds located near the focus, and these systems have already been described in Chapter 4. It was shown that the 4 m and 90 cm system offer a useful level of sensitivity in the range 50–88 MHz and 250–470 MHz, respectively. In order to introduce observing capability on the EVLA in the range of 110–250 MHz,

150 MHz–resonant dipoles are employed near the focus of the EVLA, and the performance of this system is estimated in Section 5.2. Again, the objective of this section is to create a baseline of performance of the EVLA in the range of 50–470 MHz so that future designs or modifications can be evaluated against it.

In Section 5.3, a new arrangement of dipole feeds for the EVLA – namely, "dual length dipoles" – is described which simultaneously employs the 74 MHz– and 150 MHz–resonant dipoles at the same location. The advantage of this scheme is that no *additional* blockage is introduced to higher–frequency systems due to the employment of the 150 MHz–resonant dipoles.

5.1 Dipole Feeds Located at the Focus of a Simple Reflector Antenna

In this section, we attempt to establish a performance baseline for a simple reflector antenna in the frequency range of 50–470 MHz. Three sets of crossed dipoles – specifically, 1) 74 MHz–resonant dipoles with radius of 2.38 mm, 2) 150 MHz–resonant dipoles with radius of 3.37 mm, and 3) 327 MHz–resonant dipoles with radius of 4.76 mm – are separately employed at the focus of a simple reflector for this purpose. The study is performed assuming a simple 25 m–diameter reflector antenna with f/D = 0.36.

The broadband performance of the low frequency systems consisting of these crossed dipoles is separately (i.e., only one set of crossed dipoles at a time) evaluated in terms of SEFD in the same manner as explained in Section 4.1. Specifically, the system consisting of the 74 MHz–resonant dipoles is evaluated in the range 50–88 MHz, the system with the 150 MHz–resonant dipoles is evaluated in the range 110–250 MHz, and the system with the 327 MHz–resonant dipoles is evaluated in the range 250–470 MHz. Again, due to the presence of the FM broadcast band, the range 88–108 MHz is not commonly used for astronomical observations and thus is not considered in this study. The system with 74 MHz–resonant

dipoles is analyzed using MoM as described in Section 4.1. On the other hand, the systems with 150 MHz– and 327 MHz–resonant dipoles are analyzed using the MoM+PO method as described in Section 4.2.2. In this study, R_L is assumed to be 100 Ω ; T_{int} is assumed to be 250 K for the 74 MHz–resonant dipoles and 100 K for the 150 MHz– and 327 MHz– resonant dipoles. The signals from the dipoles are combined using beamforming coefficients that maximize the SNR (as explained in Section 3.4).

The results are shown in Figure 5.1. It is observed that the large tuning range of 50–470 MHz can be covered with a useful level of sensitivity using only these three sets of crossed dipole feeds. However, these dipoles lie at the focus, and therefore are not suitable for coexistence with higher–frequency systems. Nevertheless, a baseline of performance is established from the analysis of these focus-fed systems. Our proposed low frequency systems for reducing blockage at higher frequencies, as shall be described in Chapter 6, are evaluated against this baseline.

5.2 Application Example: Dipole Feeds Located near the Focus of an EVLA Antenna

In this section, we study the utility of dipole feeds employed near the focus of an EVLA antenna, and establish a baseline of performance in the range 50–470 MHz. Performance in the range 50–88 MHz, covered by the EVLA 4 m system, has already been described in Section 4.1. In order to cover the range 110–250 MHz with the EVLA, 150 MHz–resonant dipole feeds of radius 3.37 mm are employed, and the performance is estimated in this section. For the range of 250–470 MHz, 327 MHz–resonant dipoles of radius 4.76 mm are employed. Note that the radius of the dipoles on the existing EVLA 90 cm system is 2.38 mm. The radius of the dipoles is increased for this study in order to widen the impedance bandwidth of the dipoles, and observe the effect it has on the sensitivity of the system. However, the same modification for the dipoles of the existing 4 m system is not expected to be



Figure 5.1. Estimated SEFD for a simple 25 m-diameter reflector antenna fed with 74 MHz-, 150 MHz- and 327 MHz-resonant crossed dipole feeds located at the focus. Note that the frequency range covered by each set of dipoles is identified.

very fruitful, since that system is mostly external noise–dominated in its intended frequency range of operation (i.e., 50–88 MHz); the level of external noise dominance is already shown in Figure 4.4.

As mentioned earlier, the crossed dipole feeds cannot be located at the prime focus of an EVLA reflector, as it is not accessible due to the subreflector. In this study, the 327 MHz– resonant dipoles are employed 0.23 m ($\lambda/4$ at 327 MHz) in front of the subreflector vertex as is done for the EVLA 90 cm system. In this case, the subreflector acts as a crude backplane. Likewise, the 150 MHz–resonant dipoles are employed 0.50 m ($\lambda/4$ at 150 MHz) in front of
the subreflector. The analysis of these systems are performed using the MoM+PO method; R_L and T_{int} are assumed to be 100 Ω and 100 K, respectively. Note that the 327 MHz– and 150 MHz–resonant dipoles are separately employed on an EVLA antenna to perform the analysis. The results are shown in Figure 5.2. Also shown in the figure is the performance of the existing 4 m and 90 cm systems (from Chapter 4). It is interesting to note that the 2 m and 90 cm systems perform significantly better on EVLA than on a non-Cassegrain system (as shown in Figure 5.1) – this is presumably due to the improved directivity resulting from the subreflector "ground plane".

It is observed that the new 90 cm system with thicker dipoles shows only a slight improvement in bandwidth from that of the existing system. The SEFD of the new system in the frequency ranges of 250–300 MHz and 380–470 MHz is only better by less than 1 dB.

The inclusion of 150 MHz–resonant dipoles, on the other hand, offers a useful level of sensitivity over the range of 110–250 MHz; which is currently not covered by the EVLA. An issue of concern with the 150 MHz–resonant dipoles is the additional blockage they might introduce to higher–frequency systems, since, as noted previously, the existing 4 m system reduces sensitivity at L–band by 6% [17]. A new arrangement of dipoles is therefore identified in the next section such that no *additional* blockage is introduced in the process of covering the range 110–250 MHz. This arrangement of dipoles is referred to as "dual length" dipoles.

5.3 Dual Length Dipoles for the EVLA

In this section, a new arrangement for dipole feeds – namely, dual length dipoles – is described. In this arrangement, 74 MHz– and 150 MHz–resonant dipoles are collocated. The previous section demonstrated the utility of 150 MHz–resonant dipoles of 3.37 mm radius, located 0.50 m ($\lambda/4$ at 150 MHz) in front of the subreflector of the EVLA, in covering the range of 110–250 MHz. However, we run the risk of introducing additional blockage to higher–frequency systems when these dipoles are employed at this location. An alternative



Figure 5.2. Estimated SEFD for the 74 MHz–, 150 MHz–, and 327 MHz– resonant dipoles on the EVLA.

option that avoids introducing the additional blockage is to position the dipoles at the image of the focus simultaneously with the existing 74 MHz dipoles.

The question that now arises is how to simultaneously employ the two sets of crossed dipoles at the image of the focus of the EVLA. A method has to be devised such that a single dipole's length can be changed to be equal to the length of a 74 MHz– and a 150 MHz–resonant dipole as needed. One possible approach is introduced here. The concept is shown in Figure 5.3. Crossed dipoles having radius of 2.38 mm and resonant at 74 MHz are employed at the image of the focus. Two switches are incorporated in each of the 74 MHz– resonant dipoles in a manner such that the distance between the switches equals the length of the 150 MHz–resonant dipole. Thus, when the switches are on, the dipole is resonant at 74 MHz–



Figure 5.3. The concept of dual length dipoles

74 MHz, and when the switches are off, the dipole is resonant at 150 MHz. For simplicity, the radius of the 150 MHz resonant dipoles is changed from 3.37 mm to 2.38 mm, to be the same as the 74 MHz-resonant dipoles. In order to find out the effect of this change, the dual length crossed dipoles are employed at the image of the focus and the performance is estimated. In this study, each switch is modeled as a resistive load R_d ; R_d is selected to be 10 M Ω for modeling the switch-off condition.

The results are shown in Figure 5.4. Also shown is the performance of the 150 MHz– resonant crossed dipoles of 3.37 mm radius when they are located 0.50 m ($\lambda/4$ at 150 MHz) in front of the subreflector, and when they are located at the image of the focus. It is observed that the performance of 150 MHz state of the dual length dipoles is worse than the dipoles (of radius 3.37 mm) located 0.50 m in front of the subreflector; specifically, in the frequency ranges of 110–120 MHz and 160–220 MHz, the estimated SEFD for the dual length dipoles is about 1 dB greater. This deterioration of performance is associated mostly with the change in location, as is evident from the comparison (Figure 5.4) of 150 MHz dipoles of same radius (3.37 mm) located at the two different locations –namely, 0.50 m in front of the subreflector, and the image of the focus. Moreover, the reduced radius in the dual length dipoles narrowed the impedance bandwidth. However, "something is better than nothing" – currently there is no observing capability in the range of 110–250 MHz on the



Figure 5.4. SEFD for the dual length dipoles located at the image of the prime focus of the EVLA. Also shown is the performance of 150 MHz–resonant dipoles of radius 3.37 mm when they are located 0.50 m ($\lambda/4$ at 150 MHz) in front of the subreflector, and when they are located at the image of the focus.

EVLA, and the inclusion of the dual length dipoles can cover this range without introducing additional blockage to the pre–existing higher–frequency systems.

The performance of the dual length dipoles along with the existing 90 cm system on the EVLA are summarized in the range 50–470 MHz in Figure 5.5. Comparison to Figure 5.2 shows that not much performance is lost by using the dual length dipole scheme; whereas the convenience of having only one additional set of crossed dipoles is retained.



Figure 5.5. Estimated SEFD for the dual length dipoles (74 MHz– and 150 MHz–resonant dipoles), and the 327 MHz–resonant dipoles on the EVLA. (Compare to Figure 5.2)

5.4 Summary

Contrary to the traditional approach of employing wide impedance-bandwidth feed antennas for covering large tuning ranges, this chapter demonstrated how simple dipole feeds can achieve a usable level of sensitivity over a large tuning range at low frequencies; specifically, only three sets of crossed dipole feeds are observed to offer usable sensitivity in the range 50–470 MHz. Furthermore, the concept of dual length dipoles is introduced for the EVLA which simultaneously employs the 150 MHz–resonant dipoles with the 74 MHz–resonant dipoles at the image of the focus. The results of this chapter are summarized in Figures 5.1 and 5.5.

However, in all schemes considered in this chapter, the dipoles are located on the reflector axis, and thus introduce blockage to higher–frequency systems. With a view to alleviate this situation, we investigate in the next chapter the use of an array of feeds distributed over locations less prone to high frequency blocking.

Chapter 6

Distributed Feed Array

In this chapter, the design of a new class of low frequency feeds – namely, the distributed feed array – is described. Chapter 5 demonstrated that low frequency reflector antenna systems consisting of crossed dipole feeds located at or near the focus can cover the range 50–470 MHz with a useful level of sensitivity. An issue of concern with these focus- or near focus-fed systems is that they introduce blockage to higher-frequency systems, because they are positioned on the reflector axis. However, as mentioned in Chapter 1 (see Figure 1.9), at low frequencies, an appreciable amount of power is available at locations away from the reflector axis, and it might be possible to achieve comparable performance to focus- or near focus-fed systems by properly combining arrays of feeds located at these locations; thereby reducing blockage to higher-frequency systems. This motivation led to the design of the distributed feed array. The concept of a distributed feed array is shown by example in Figure 6.1. In this arrangement, feed elements are distributed in the volume in front of the main reflector, and the signals from these elements are combined to produce the output. In this dissertation, dipoles are used as the feed elements, and the combining coefficients are selected to achieve the maximum SNR. Note that the combining coefficients can also be selected to introduce nulls in the reflector antenna pattern at certain directions in order to reduce interference; e.g., as is commonly proposed with focal plane arrays. However, this is



Figure 6.1. An example of a distributed feed array.

left as a possible area of future work.

The rest of this chapter is organized as follows. Section 6.1 identifies prospective "high frequency safe" areas in front of the reflector for employing distributed feed arrays. Based on the study of Section 6.1, distributed feed arrays are designed for a simple 25 m-diameter reflector antenna in Section 6.2. The objective is to obtain insight about how well the frequency range of 50–470 MHz can be covered with dipole arrays positioned at high frequency safe locations. In order to demonstrate the application of distributed feed arrays, low frequency feeds consisting of arrays of dipoles are employed on an EVLA antenna to cover the range 50–470 MHz; this is described in Section 6.3. Findings are summarized in Section 6.4.

6.1 High Frequency Safe Areas in Front of a Main Reflector

In this section, we identify "high frequency safe" locations in front of a main reflector for employing distributed feed arrays. To serve this purpose, we first consider the path of propagation in the volume in front of the main reflector. The three regions of propagation in a Cassegrain reflector antenna is shown in Figure 6.2. Region 1 corresponds to the plane wave illumination of the reflector. Region 2 is associated with the fields reflected from the main reflector toward the prime focus. Region 3 encapsulates the fields reflected from the subreflector toward the secondary focus. The first two regions can be identified for any reflector antenna, whereas the third region is specific to Cassegrain systems. The power density of the incident field in Region 1 is uniform, and in Region 2 the power density of the incident field increases as we move toward the focus. Also, in Region 3, the power density of the incident field increases as we move toward the secondary focus. Therefore, depending on the region in which the low frequency feeds are employed, the blockage at higher frequencies will vary; blockage increasing as we move from Region 1 to 3. For example, on an EVLA antenna the feeds for the existing 4 m system are located in Region 3, and therefore introduce significant blockage (4 m system reduces sensitivity by 6% at L-band [17]). However, if the feeds can be shifted to Region 1 or 2 the blockage will presumably be reduced.

Based on the above discussion, we narrow down prospective arrangements for distributed feed arrays to the following 2 schemes: 1) arrays of dipoles in the focal plane, corresponding to distributed feed arrays in Region 1; and 2) arrays of dipoles at the border of Regions 1 and 2. The relative merits of these schemes are explored in the following sections.

6.2 Distributed Feed Arrays on a Simple Reflector Antenna

In this section, distributed feed arrays are employed on a simple 25 m-diameter reflector antenna – the same antenna as analyzed in Section 5.1 for focus–fed systems. The objective is to find out how well (from a sensitivity perspective) the desired range 50–470 MHz can



Figure 6.2. Regions of different power densities in the volume in front of a Cassegrain reflector antenna.

be covered with this feeding scheme. Three different arrays consisting of 74 MHz-resonant dipoles, 150 MHz-resonant dipoles, and 327 MHz-resonant dipoles are selected for this purpose. The radii of these dipoles are 2.38 mm, 3.37 mm and 4.76 mm, respectively. It has already been shown in Section 5.1 that these dipoles can cover the range 50–470 MHz with a useful level of sensitivity when they are employed at the focus. In this section, we study the utility of these dipoles when employed as feed elements in a distributed array.

The content of this section is divided into three parts. In Section 6.2.1, arrays of dipoles positioned in the focal plane of a simple reflector antenna are investigated; this corresponds to feeds in Region 1. Section 6.2.2, on the other hand, studies the utility of arrays of dipoles employed at the border of Regions 1 and 2. Findings are summarized in Section 6.2.3.

6.2.1 Dipole Arrays in the Focal Plane of a Simple Reflector Antenna

In this section, the utility of dipole arrays employed in the focal plane of a simple reflector antenna is studied. The arrays consist of 4 dipoles arranged in a boxing ring configuration around the focus, as shown in Figure 6.3. A design parameter to be selected for this scheme is the distance d_f of the dipole centers from the focus. A simple study is performed to help us select reasonable locations in the focal plane.

In this study, we observe the distributions of power density in the focal plane of a simple reflector antenna upon a plane wave illumination at 74 MHz, 150 MHz and 327 MHz. These frequencies are selected because they correspond to frequencies of particular interest in radio astronomy, and are the resonant frequencies of the dipoles that we intend to use in the distributed feed arrays. The distribution of power density at 74 MHz is calculated by radiating the currents on the reflector surface in response to a plane wave illumination, found using MoM. In order to do the same at 150 MHz and 327 MHz, the required MoM models result in too many segments (28,294 and 115,696, respectively) to fit in commonly available computer memory (e.g., 4 GB of RAM), and thus is not feasible. However, at 150 MHz and 327 MHz the diameter of the reflector is large compared to wavelength (14 λ and 28 λ , respectively), and therefore PO might be used to do the analysis. In order to check the validity of PO, the distribution of power density in the focal plane of a 25 m-diameter reflector antenna, in response to a plane wave illumination, is computed at 74 MHz, using both PO and MoM, and the results are compared. The argument here is that if PO works well at 74 MHz, it should work better at higher frequencies. The result of the study is shown in Figure 6.4. It is observed that the results computed using PO and MoM are comparable at 74 MHz. Therefore the use of PO to compute the distributions of power density at 150 MHz and 327 MHz is justified.

Figure 6.5 shows the distributions of power density in the focal plane, with respect to the power density at the focus at the three selected frequencies. At all the three selected frequencies we observe a main peak at the focus, as expected. We presume that the farther



Figure 6.3. Feeds around the prime focus of a simple reflector in a boxing ring configuration.

the dipoles of the distributed feed arrays are from the focus, the better in terms of blockage reduction; however, the amount of available power reduces with increasing distance from the focus. Therefore a balance needs to be struck in selecting prospective locations for the dipole feeds.

Our selection of prospective locations in the focal plane for employing the dipole arrays are shown in Table 6.1. The first choice of location is the point where the first secondary peak occurs in the power distribution. This location is referred to as L_1 . For the next choice of prospective location, we move the dipoles closer to the focus until we reach the point where the power level is doubled from that of the first secondary peak. These locations are identified as L_2 in Table 6.1. The third choice of location is the closest distance to the focus which fits the array. These locations are identified as L_s in Table 6.1. Thus, we have narrowed down the design space to three locations at each frequency. We now estimate the performance (based on SEFD) of the low frequency systems when they are employed at these locations. The analysis for the array of 74 MHz–resonant dipoles is performed using MoM as described in Section 3.6. The systems consisting of 150 MHz– and 327 MHz–resonant dipole arrays are analyzed using the MoM+PO method (described in Section 4.2.2). Note that the arrays of 74 MHz–, 150 MHz–, and 327 MHz–resonant dipoles are separately employed to feed the reflector in this study, i.e., only one array is present at a time. Again, R_L is assumed to be 100 Ω , and T_{int} is assumed to be 250 K for the 74 MHz–resonant dipoles and 100 K for the 150 MHz– and 327 MHz–resonant dipoles. The combining coefficients are selected to maximize SNR (as described in Section 3.4).

The SEFD estimates are shown in Table 6.1. Also shown for comparison in the table is the performance estimated for focus-fed systems described in Section 5.1. Upon comparison with the focus-fed systems, it is observed that there is no obvious optimum choice that balances performance (in terms of sensitivity) and distance from focus (decreasing blockage).

For the sake of further analysis we shall select the systems employed at L_2 locations (i.e., 1.32 m, 0.65 m, and 0.30 m away from the focus at 74 MHz, 150 MHz and 327 MHz, respectively) which are seen to provide sensitivity roughly comparable to that achieved by focus-fed systems. Again, the intended frequency coverage with the individual arrays is 50-88 MHz for the 74 MHz-resonant dipole array, 110-250 MHz for the 150 MHz-resonant dipole array and 250-470 MHz for the 327 MHz-resonant dipole array. The results of our analysis are shown in Figure 6.6. Also shown in the figure is the performance obtained with focus-fed systems in the range of 50-470 MHz as described in Section 5.1. It is observed that the estimated performance of all the three systems follow a common trend; the performance is comparable to focus-fed systems at or below the resonant frequency, and starts to become worse as the frequency is increased. The reason for this behavior is that the electrical separation of the dipoles of the distributed array from the focus is increased as frequency is increased; the focus-fed systems, on the other hand, always have the feeds at the focus.



Figure 6.4. Power density in the focal plane of a simple 25 m-diameter reflector in response to a plane wave illumination at 74 MHz. PO and MoM are separately used to do the analysis.

Nevertheless, the array of 74 MHz–resonant dipoles offer sensitivity within 1 dB of that achieved with prime focus-fed systems at least in the range of 50–80 MHz; while the 150 MHz– and 327 MHz–resonant dipole arrays offer sensitivity within 1 dB of that offered by focus–fed systems in the range of 110–160 MHz and 250–350 MHz, respectively.

6.2.2 Dipole Arrays at the Border of Regions 1 and 2 of a Simple Reflector Antenna

In this section, we study the utility of distributed feed arrays positioned around the conceptual surface which lies at the border of Regions 1 and 2. If we consider only the



Figure 6.5. Power density versus the distance d_f from the focus in the focal plane of a simple reflecting paraboloid (D = 25 m, f/D = 0.36), relative to the power density at the focus, in response to a plane wave illumination. L_1 , L_2 and L_s refer to our choices for prospective locations (see text).

geometrical optics (GO) fields, this surface assumes a conical shape with the vertex of the cone at the focus, and the base along the rim of the reflector. The distributed feed arrays positioned at this border consist of four dipoles arranged in a boxing ring configuration around the axis of the reflector as shown in Figure 6.7.

In this study, we first observe the distribution of power density at the border of Regions 1

Freq.	SEFD for	Fig.	d_f at L_1	d_f at L_2	d_f at L_s
	Focus–Fed Sys.		SEFD	SEFD	SEFD
$74 \mathrm{~MHz}$	32.4 kJy	6.5(a)	2.40 m	$1.32 \mathrm{~m}$	1.00 m
			93.3 kJy	$44.7 \mathrm{~kJy}$	30.2 kJy
$150 \mathrm{~MHz}$	8.3 kJy	6.5(b)	$1.15 \mathrm{~m}$	$0.65 \mathrm{~m}$	0.48 m
			$15.1 \mathrm{~kJy}$	6.2 kJy	4.0 kJy
$327 \mathrm{~MHz}$	3.0 kJy	6.5(c)	$0.53 \mathrm{~m}$	0.30 m	0.22 m
			13.8 kJy	$7.2 \mathrm{~kJy}$	$5.9 \mathrm{~kJy}$

Table 6.1. Prospective locations for employing dipole rings in the focal plane, expressed as the distance of the dipole centers from the focus. Also shown are the estimated performance of dipole rings employed at these locations and the estimated performance of focus–fed systems.

and 2 at 74 MHz, 150 MHz and 327 MHz. This gives us insight about prospective locations. The power distributions are shown in Figure 6.8. The distribution of power density is presented as a function of r_d , where r_d is the radial distance from the focus toward the rim of the reflector. Upon comparison with the distributions of power density in the focal plane (Figure 6.5), it is observed that at a certain distance from the focus, a higher level of power density is available at the border of Regions 1 and 2 than is a in the focal plane.

Our choices of prospective locations are shown in Table 6.2. Again, at each of the three frequencies, our first choice of location for employing the dipoles is the point where the first secondary peak (pointed as L_1) occurs in the power distribution. For the next choice of prospective location we move the dipoles closer to the focus along r_d until we reach the point where the level of power density is twice of that at L_1 . These locations are identified in Table 6.2 as L_2 . The third choice for location is the closest distance to the reflector axis that can fit the dipole rings. These locations are identified as the L_s in Table 6.2.

Thus, we narrowed down the design space at each frequency to three locations. We now estimate the performance (based on SEFD) of the corresponding dipole arrays when they are separately (i.e., only one array at a time) employed at these locations. Again, R_L is assumed to be 100 Ω ; T_{int} is assumed to be 250 K for the 74 MHz–resonant dipoles and 100 K for the 150 MHz– and 327 MHz–resonant dipoles. The combining coefficients are selected to



Figure 6.6. Estimated SEFD for the arrays of 74 MHz–, 150 MHz– and 327 MHz–resonant dipoles in the focal plane of a simple reflector antenna with D = 25 m and f/D = 0.36.

maximize SNR.

The results are shown in Table 6.2. Also shown for comparison in the table is the performance of the focus-fed systems. Upon comparison with focus-fed systems, it is again observed that there is no obvious optimum choice that balances performance and distance from the reflector axis.

For the sake of future analysis, we shall select location L_s for the 74 MHz dipoles (i.e., $r_d = 1.27$ m) and location L_2 for 150 MHz and 327 MHz dipoles, respectively (i.e., $r_d = 0.77$ m and $r_d = 0.35$ m, respectively); these locations are seen to provide sensitivity roughly comparable to that achieved by focus-fed systems.



Figure 6.7. Feeds at the border of Regions 1 and 2 in a boxing ring configuration.

The results are shown in Figure 6.9. Also shown in the figure is the performance obtained with the focus-fed systems described in Section 5.1. It is observed that the estimated performance of all the three systems follow the same trend as observed in the previous section for the dipole rings in the focal plane. The performance of these low frequency systems are within 1 dB of that offered by focus-fed systems in the range of 50–80 MHz for the 74 MHz-resonant dipole array, 120-150 MHz for the 150 MHz-resonant dipole array and 250–330 MHz for the 327 MHz-resonant dipole array. However, as is shown in Figure 6.10, the performance of the selected arrays is somewhat worse than the performance of the corresponding focal-plane arrays; specifically, in the range 110–210 MHz and 250–450 MHz. Thus, the latter



Figure 6.8. Power density versus the distance from the focus along the border of Regions 1 and 2 for a simple reflecting paraboloid (D = 25 m, f/D = 0.36), relative to the power density at the focus, in response to a plane wave illumination. L_1 , L_2 and L_s refer to our choices for prospective locations (see text).

should be considered only if the former is not practical; e.g., due to difficulty of installation. The "dipoles at the border of Regions 1 and 2" scheme, on the other hand, is probably more attractive since existing struts (supporting the existing higher–frequency feeds) are located close to this border, and the dipoles could be strung between these struts. Thus, we are motivated to consider this strategy further.

Freq.	SEFD	Fig.	r_d at L_1	r_d at L_2	r_d at L_s
	for Focus–Fed Sys.		\mathbf{SEFD}	\mathbf{SEFD}	SEFD
$74 \mathrm{~MHz}$	32.4 kJy	6.8(a)	3.20 m	$1.55 \mathrm{~m}$	$1.27 \mathrm{~m}$
			128.8 kJy	$95.5 \mathrm{~kJy}$	$42.7 \mathrm{~kJy}$
$150 \mathrm{~MHz}$	8.3 kJy	6.8(b)	1.60 m	0.77 m	$0.65 \mathrm{~m}$
			47.9 kJy	10.0 kJy	$6.3 \mathrm{~kJy}$
$327 \mathrm{~Mhz}$	3.0 kJy	6.8(c)	0.74 m	$0.35 \mathrm{~m}$	$0.30 \mathrm{~m}$
			13.2 kJy	4.8 kJy	3.0 kJy

Table 6.2. Prospective locations for employing dipole rings at the border of Regions 1 and 2 of a simple reflector antenna, expressed as the distance of the dipole centers from the focus along r_d .

A possible way to improve the performance of the feeds distributed along the border of Regions 1 and 2 is to employ a double ring of dipoles. Based on the distributions of power density shown in Figure 6.8, suitable locations for the second ring of dipoles can be selected. In this study, r_d of 1.55 m, 1.60 m and 0.74 m are selected for the locations of the second rings at 74 MHz, 150 MHz and 327 MHz, respectively. These locations correspond L_2 at 74 MHz and L_1 at 150 MHz and 327 MHz. The broadband performance of these double ring systems is separately analyzed, and the achievable sensitivity is estimated.

The results are shown in Figure 6.11. An improvement in performance with the double ring of dipoles is evident. The double ring of 74 MHz-resonant dipoles perform within $\ll 1$ dB (in terms of SEFD) of the focus-fed system in the range of 50–88 MHz. The performance of the double ring of 150 MHz-resonant dipoles, on the other hand, is observed to be better than the focus-fed system, roughly in the range 110–150 MHz, but then gets worse in the range of 150–230 MHz. For the double ring of 327 MHz-resonant dipoles, performance is better than or comparable to the focus-fed system in the range of 250–410 MHz, but gets worse in the range of 420–460 MHz. As described in the previous section, the performance of these systems becomes worse than focus-fed systems with increasing frequency because the electrical separations of the dipoles from the focus increase with increasing frequency.



Figure 6.9. Estimated SEFD for single rings of 74 MHz–, 150 MHz– and 327 MHz–resonant dipoles at the border of Regions 1 and 2 of a simple reflector antenna with D = 25 m and f/D = 0.36.

6.2.3 Summary

To summarize, this section demonstrated that distributed feed arrays on a simple reflector antenna can effectively cover the frequency range of 50–470 MHz. Dipole arrays in the focal plane of a simple reflector antenna are observed to offer a performance comparable to focus-fed systems within an appreciable bandwidth (about 30–40%) around the resonant frequencies of the dipoles. Single rings of dipoles employed at the border of Regions 1 and 2 of the reflector are observed to be a little worse (about 2-5 dB) than the arrays in the focal



Figure 6.10. Comparison of performance between arrays of dipoles at the border of Regions 1 and 2 (SEFD_{co}), and arrays of dipoles in the focal plane (SEFD_{fc}) of a simple reflector antenna with D = 25 m and f/D = 0.36.

plane, but they presumably have reduced blockage associated with them. Double rings of dipoles employed at the border of Regions 1 and 2 have improved performance.

In the next section, the distributed feed array concept is applied to the EVLA.



Figure 6.11. Estimated SEFD for the double rings of 74 MHz–, 150 MHz– and 327 MHz–resonant dipoles at the border of Regions 1 and 2 of a simple reflector antenna with D = 25 m and f/D = 0.36.

6.3 Application Example: Distributed Feed Arrays on an EVLA Antenna

In this section, we apply the distributed feed array scheme to the EVLA. Since the existing 90 cm system on the EVLA can offer usable sensitivity in the range 250–470 MHz (as shown in Section 4.2.2) without affecting higher–frequency operation, our investigations of distributed feed arrays on the EVLA are limited to covering the range 50–250 MHz. Therefore, arrays consisting of only 74 MHz– and 150 MHz–resonant dipoles are considered. Moreover, we

do not have much freedom in selecting locations for employing feeds in this case, because whatever modifications we propose should be adaptable without requiring major changes to the EVLA's Cassegrain geometry. Considering the EVLA geometry (shown in Figure 2.20) and findings from Section 6.2.2, a reasonable arrangement is identified in this dissertation. In this arrangement, the dipoles are employed between the adjacent struts of the antenna, as shown in Figure 6.1. This scheme is referred to as the "strut-straddling" scheme. In this scheme, the strut-straddling dipoles are positioned either in Region 1 or 2; depending on the precise mounting location. The utility of this scheme is studied in the following sections. In Section 6.3.1, strut-straddling dipoles are employed in the plane that contains the prime focus of the EVLA, i.e., the z = 0 m plane (see Figure 2.20). In Section 6.3.2, dipole arrays are employed at planes parallel to the z = 0 m plane but closer to the reflector. In Section 6.3.3 we describe the use of "dual length dipoles" (described in Section 5.3) in the strut-straddling scheme for the EVLA.

6.3.1 Strut–Straddling Dipoles in the Plane Containing the EVLA Prime Focus

The study in this section considers strut-straddling dipoles in the plane containing the EVLA prime focus, i.e., the z = 0 m plane. Arrays of 74 MHz– and 150 MHz–resonant dipoles, employed between the adjacent struts of an EVLA antenna, are evaluated one at a time, and the performance of these systems is estimated. The radii of these dipoles are 2.38 mm and 3.37 mm, respectively. Again, R_L is assumed to be 100 Ω , and T_{int} is assumed to be 250 K for the 74 MHz–resonant dipoles and 100 K for the 150 MHz–resonant dipoles.

A design consideration in this study is whether to select the outer edges or the inner edges of the struts for employing the dipoles. Dipoles employed between the outer edges are expected to have reduced blockage associated with them, as they are farther away from the axis of the reflector; however, the reduction in sensitivity due to the increased separation from the focus, needs to be calculated to estimate the merit of this arrangement. In order to find the answer, both locations are investigated separately.

Freq.	SEFD for Image	SEFD for Focal	SEFD for Focal	
	Focus–Fed System	Plane Array at	Plane Array at	
		Inner Edges of Strut	Outer Edges of Strut	
74 MHz	60.3 kJy	55.0 kJy	120.2 kJy	
150 MHz	6.2 kJy	38.0 kJy	43.7 kJy	

Table 6.3. Strut-straddling dipole arrays in the focal plane of an EVLA an-tenna.

The results are summarized in Table 6.3. Also shown in the table is the performance achieved when crossed dipoles are located at the image of the prime focus. It is observed that the performance of the array of 74 MHz–resonant dipoles located between the inner edges of the struts is comparable to the focus-fed system. SEFD increases (becomes worse) by about 3.6 dB at 74 MHz as we move the dipoles from the inner edges to the outer edges. On the other hand, the performance at 150 MHz drops (SEFD increases) by 0.8 dB when the dipoles are employed between the outer edges of the struts. It is observed upon comparison with the focus-fed system that the performance of the array of 150 MHz-resonant dipoles at either location (between the inner edges or the outer edges of the struts) is significantly worse, and is also much worse than the performance of the "border of Regions 1 and 2" scheme of Section 6.2.2. However, the disadvantage of reduced sensitivity for the strut-straddling dipoles, may be somewhat negated by the advantage of reduced blockage associated with them combined with the relative ease of installation in this scheme. Based on the above considerations, we now separately study the broadband performance of the array of 74 MHz– and 150 MHz–resonant dipoles located between the inner edges of the struts. The results are shown in Figure 6.12. Also shown in the figure for comparison is the estimated broadband performance for the case when the dipoles are employed at the image of the focus of the EVLA (from Section 5.3). It is observed that the strut-straddling array of 74 MHz-resonant dipoles in the z = 0 m plane of an EVLA antenna performs better than the existing 4 m system in the range of 50–74 MHz; however, as frequency is further increased the performance starts to worsen. Nevertheless, this performance is very encouraging; the strut-straddling array, being away from the axis of the reflector (i.e., in Region 1) is expected to present greatly reduced blockage to higher-frequency systems. The array of 150 MHz-resonant dipoles, on the other hand, performs dramatically worse than the image focus-fed system in the entire region of interest. However, this strut-straddling system offers usable sensitivity over a frequency range (110–250 MHz) which is currently not covered by the EVLA; moreover, higher-frequency operation is presumably not affected by the presence of this system.

6.3.2 Strut-Straddling Arrays in Planes Parallel to the z = 0 m Plane on the EVLA

The second part of our study with strut-straddling arrays on the EVLA investigates dipole rings in planes parallel to the z = 0 m plane, but closer to the reflector. In order to find out prospective locations between adjacent struts, we separately employ single rings of dipoles resonant at 74 MHz and 150 MHz, at different planes parallel to the focal plane, and estimate the sensitivity of the resulting systems at the corresponding resonant frequencies. In this experiment, locations between both the inner and outer edges of the struts are studied.

The results of our analysis with 74 MHz-resonant dipoles are shown in Figure 6.13. Also shown is the estimated performance of the existing EVLA 4 m system. It is observed that single rings located between the inner edges of the struts perform better than when positioned between outer edges as might be expected. The lowest SEFD at 74 MHz is observed when the dipoles are between the inner edges of the struts in the plane z = -1 m. At this location, the SEFD is about 10% better than that for the existing 4 m system. On the other hand, the best performance for rings positioned between the outer edges of the struts occurs at z = -2 m, but the SEFD at this location is about 2 dB worse (i.e., SEFD is higher). Therefore the location between the inner edges of the struts at z = -1 m is selected as the preferred location for mounting a single ring of 74 MHz-resonant dipoles. The second lowest SEFD is observed between the inner edges of the struts at z = -2.25 m,



Figure 6.12. Estimated SEFD for the arrays of 74 MHz– and 150 MHz– resonant strut–straddling dipoles in the focal plane of an EVLA antenna.

and is comparably low. This suggests the possible utility of a second ring of dipoles. This possibility will be considered later in this section.

The broadband performance of the single ring of dipoles at z = -1 m is estimated and compared with the EVLA 4 m system. The results are shown in Figure 6.14. It is observed that this system has comparable performance to the existing (focus-fed) system, and in the range 55 - 75 MHz the proposed system performs better. At 74 MHz the new system yields SEFD ≈ 53 kJy whereas the existing system yields SEFD ≈ 60 kJy.

A possible way to improve the sensitivity obtained with a single ring of strut-straddling dipoles without presumably making blockage worse is to use two rings, as shown in Fig-



Figure 6.13. Location selection for the 74 MHz strut-straddling dipole array.

ure 6.15. As pointed out earlier, z = -1 m and z = -2.25 m appear to be the suitable choices for the locations of the two rings. Figure 6.14 shows the results for this arrangement. Note that the SEFD for the double ring system is better than the existing system at all frequencies considered except for 50 and 80 MHz. The advantage of both the single ring and double ring of dipoles schemes is that the blockage presented to the pre-existing higherfrequency systems is presumably less than that observed for image focus-fed systems. An assessment of the reduction in blockage is presented in Chapter 7.

In a similar way as described for the array of 74 MHz–resonant dipoles, suitable locations for single ring and double ring of 150 MHz–resonant dipoles can be determined. The search



Figure 6.14. Estimated SEFD for strut–straddling arrays of 74 MHz– resonant dipoles. Also shown for comparison is the estimated performance of the existing EVLA 4 m system.

for the best locations between adjacent struts is performed for both the inner and the outer edges of the struts. The results are shown in Figure 6.16. The lowest SEFD at 150 MHz is observed between the inner edges at z = -1 m. However, the SEFD at this location is about 3 dB worse (i.e., higher) than the image focus-fed system described in Section 5.3. The broadband performance of this system is estimated and the results are shown in Figure 6.17. As expected, the single ring of 150 MHz-resonant dipoles performs worse than the image focus-fed system in the entire intended region of operation – namely, 110–250 MHz.

As we did for the 74 MHz dipole array, we now consider using a second ring of dipoles.



Figure 6.15. Double ring of strut-straddling dipoles.

It is observed from Figure 6.16 that a suitable location for the second ring of dipoles is between the inner edges of the struts at z = -1.25 m. However, this will result in four separate rings – two for 74 MHz–resonant dipoles and two for 150 MHz–resonant dipoles – for covering the range 50–250 MHz, which might ruin our objective of coexistence with the pre–existing higher–frequency systems. Therefore an alternative option might be to use dual length dipoles (described in Section 5.3) at z = -1 m and z = -2.25 m to cover the range 50–250 MHz. This possibility is considered in the next section. Note that for the 150 MHz– resonant dipole array, SEFD at z = -2.25 m is worse than that at z = -1.25 by less than 1 dB.

6.3.3 Double Ring of Dual Length Dipoles for the EVLA

In this section we consider a scheme in which the 150 MHz-resonant dipoles are positioned at z = -1 m and z = -2.25 m; i.e., collocated with the double ring of 74 MHz-resonant dipoles. The dual length dipole arrangement, as described in Section 5.3, is proposed for this purpose. Note that the radius of the 150 MHz-resonant dipoles is changed from 3.37 mm to 2.38 mm in this case. The performance of the 74 MHz dipoles in this scheme is the same as shown in Figure 6.14. The broadband performance of the double ring of dual length dipoles is estimated with the switches in the dipoles in the off-mode, i.e., the dipoles are resonant at 150 MHz.

The results are shown in Figure 6.17. It is observed that the performance estimated for this system is better than the single ring of 150 MHz–resonant dipoles in the range 110– 250 MHz, as expected. On the other hand, the performance is still much worse than the image focus–located crossed dipoles. However, the system consisting of the double ring of dual length dipoles offers usable sensitivity in a frequency range (110–250 MHz) at which currently there is no observing capability on the EVLA. Moreover, this enhancement of observing capability does not introduce any *additional blockage* over that of the 74 MHz double ring scheme.

6.4 Summary

To summarize, this chapter demonstrated the utility of distributed feed array schemes for reflector antennas in covering our desired frequency range. Single rings of dipoles in the focal plane and at the border of Regions 1 and 2 of a simple reflector antenna are shown to offer usable sensitivity. The performance is comparable to focus-fed systems within an appreciable



Figure 6.16. Location selection for 150 MHz–resonant strut-straddling dipoles array.

bandwidth (about 30–40%) around the resonant frequencies of the dipoles. The advantage of these systems is that they might be suitable for coexistence with higher–frequency systems due to the reduced blockage expected to be associated with them. As a possible way to improve performance, the employment of double rings of dipoles is identified, and indeed, improvement is noticed.

The distributed feed array scheme is applied to the EVLA to cover the range 50–250 MHz. A double ring of dipoles at z = -1 m and z = -2.25 m is observed to cover the range with a usable level of sensitivity. The "dual length" dipole arrangement is proposed to simultane-



Figure 6.17. Estimated SEFD for strut–straddling arrays of 150 MHz–resonant dipoles on the EVLA. Also shown for comparison is the estimated SEFD for 150 MHz–resonant crossed dipole feeds located at the image of the prime focus.

ously employ dipoles resonant at 74 MHz and 150 MHz at these locations. The performance of this proposed system is shown in Figure 6.18, and compared to the image focus-fed scheme described in Section 5.3. The distributed array of 74 MHz–resonant dipoles performs better than the existing EVLA 4 m system in the range 50–88 MHz. The performance of the array of 150 MHz dipoles, on the other hand, is worse (about 2–5 dB) than the image focus–fed system in the entire frequency range investigated. However, this system is still useful, because the frequency range of 110–250 MHz, which is currently not covered by the



Figure 6.18. Double ring of "dual length" dipoles positioned at z = -1 m and z = -2.25 m on an EVLA antenna. Also shown is the estimated performance when dual length crossed dipoles are employed at the image of the prime focus of an EVLA antenna.

EVLA, is at least covered with a useful level of sensitivity without (presumably) affecting higher–frequency operation.

As noted several times, the utility of any of these schemes depends on the extent to which they interfere with the higher–frequency systems. The blockage associated with this distributed feed array is assessed in the next chapter.

Chapter 7

Assessment of Blockage due to Low Frequency Feeds on the EVLA

In this section, we assess the blockage introduced by various low frequency feeds on the EVLA. Specifically, blockage introduced by the image focus-located crossed dipoles of the existing 4 m system is studied, and a baseline is established. The blockage presented by the double ring of dipoles at z = -1 m and z = -2.25 m, proposed in Chapter 6, is then compared against this baseline. As mentioned in Chapter 6, the motivation behind the distributed feed array is that blockage of pre-existing higher-frequency systems is expected to be reduced compared to that of image focus-located crossed dipole systems. Here, we do a simple analysis to confirm this assumption.

In Section 7.1, we analyze the blockage associated with the dipoles of the existing 4 m system. In Section 7.2, the blockage introduced by the double ring of strut-straddling dipoles at z = -1 m and z = -2.25 m is studied. Findings are summarized in Section 7.3.

7.1 Assessment of Blockage due to the Existing EVLA 4 m System

In this section, we study the blockage presented by the existing EVLA 4 m system to preexisting higher-frequency systems on the EVLA. The objective is to create a baseline against which the blockage associated with the strut-straddling design, proposed in Chapter 6, can be compared, and the claim of reduction in blockage can be verified. In Section 7.1.1 and 7.1.2, the blockage presented to the EVLA L-band system, and the EVLA 90 cm system are investigated, respectively.

7.1.1 Assessment of Blockage at 1.4 GHz due to the EVLA 4 m System

In this section, the blockage introduced at 1.4 GHz by the crossed dipole feeds of the existing EVLA 4 m system is estimated. The frequency of 1.4 GHz is of interest because of the existing L-band system on the EVLA. The dipoles of the EVLA 4 m system are observed in practice to reduce the sensitivity of the L-band system by about 6% [17]. We now seek an independent assessment from simple theoretical electromagnetic considerations.

A direct approach to analyze the effect of the EVLA 4 m system at L-band would be a MoM analysis at 1.4 GHz; however, this is not practical, because the required MoM model will have too many segments to fit in commonly available computer memory. A hybrid method is therefore developed to serve the purpose. The method is as follows.

We first identify three different segments in the path of propagation in a reflector antenna, under Cassegrain operation. For simplicity, we consider a transmit case, as is shown in Figure 7.1, and invoke reciprocity to assume the results to be valid for the receive case. The first segment $\overline{F'S}$ is from the secondary focus to the subreflector. The second segment \overline{SR} is from the subreflector to the main reflector. The third segment \overline{RA} is from the main reflector toward the far field.

In order to estimate blockage in the first two segments – namely, $\overline{F'S}$ and \overline{SR} – at 1.4 GHz,


Figure 7.1. Segments in the propagation path in a Cassegrain reflector antenna in transmit mode.

the reduction in the aperture efficiency of the EVLA antenna due to the image focus-located crossed dipoles is computed. A half-wave dipole feed is employed at the secondary focus. The dipole is excited and the resulting currents on the antenna components present in front of the main reflector (specifically, the subreflector, the struts, the mounts and low frequency feeds being considered) are found using MoM at 1.4 GHz. The models of these components, developed in Section 2.5, are scaled with frequency (to meet NEC2 guidelines at 1.4 GHz) for the MoM analysis. These currents are then radiated to obtain the field incident on the reflector surface. In this method, the feed pattern is not included in computing the field incident on the main reflector (as is the case for the actual L-band feed). Thus, PO currents on the surface of the main reflector are found. These currents are then radiated to find the directivity of the antenna. The study is separately performed for the following two cases: 1) when there are no low frequency dipoles present in the model, and 2) when the crossed dipoles of the existing 4 m system are employed at the image of the focus (the currents on the blocking dipoles are also radiated). The feed antenna employed at the secondary focus for the blockage analysis at 1.4 GHz is a simple dipole feed, as opposed to the horn antenna of the actual L-band system. The horn feed is more directive than the dipole: the horn antenna illuminates the subreflector with an edge taper of -17 dB [52]; on the other hand, the edge illumination for the dipole feed is about -4 dB. Nevertheless, this simplistic approach of using a dipole feed is useful for comparing the *relative* blockage introduced by different low frequency feeds. In other words, the analysis may or may not provide accurate estimation of blockage, but insight about which design is more "high frequency friendly" is obtained.

The results are shown in Figure 7.2. It is observed that when the blocking dipoles are not present, the directivity of the antenna is 35.8 dB. On the other hand, when the crossed dipoles are present at the image of the prime focus, the directivity is 35.6 dB, corresponding to a reduction of about 4.5% in the aperture efficiency.

In the above analysis, blockage in the segment RA is not considered. In order to quantify blockage in this segment at 1.4 GHz, we need to compute the effects of the scattering objects (struts, subreflector, subreflector mount, and blocking dipoles), present in front of the main reflector, on the fields reflected from the main reflector. This requires a full wave analysis at 1.4 GHz, which again is not practical. Instead we adopt another approximate method which is described here.



Figure 7.2. Directivity of an EVLA antenna at 1.4 GHz, in response to a dipole feed located at the secondary focus, for different blocking conditions. The "No low frequency dipoles" and "Double ring of strut–straddling dipoles" results are essentially the same over most of this range.

The electric field $\mathbf{E}_{\mathbf{r}}$ reflected from the surface of the main reflector has a $1/r_f$ ($r_f = f \sec^2(\frac{\theta_f}{2})$ with f being the focal length) magnitude dependence due to the spherical nature of the incident field. For this study, the magnitude of $\mathbf{E}_{\mathbf{r}}$ is assumed to be

$$|\mathbf{E}_{\mathbf{r}}| = \frac{1}{\sec^2\left(\frac{\theta_f}{2}\right)}$$
$$= \cos^2\left(\frac{\theta_f}{2}\right).$$
(7.1)

We can now calculate the total power P_1 in the reflected field that flows through the

aperture, defined by the outline of the reflector, as

$$P_1 = \iint_{aperture} \frac{1}{2\eta} |\mathbf{E}_{\mathbf{r}}(\rho', z)|^2 \rho' d\rho' d\phi' .$$
(7.2)

This is the amount of power that flows through the aperture when no blocking/scattering objects are present in front of the main reflector. However, when the low frequency dipoles are present in front of the reflector some part of P_1 is scattered and the directivity of the antenna is reduced. We now estimate the amount the power that gets scattered when dipoles are positioned at the image of the focus.

It can be observed from Equation 7.1 that the magnitude of the electric field along the main reflector axis (i.e., $\theta_f = \pi$) is 1 V/m. A plane wave of amplitude 1 V/m is therefore used to illuminate the image focus-located dipole and the scattered field \mathbf{E}_s is calculated using MoM. The total power scattered P_2 is then

$$P_{2} = \int_{\theta=0}^{\pi} \int_{\phi=0}^{2\pi} \frac{1}{2\eta} |\mathbf{E}_{\mathbf{s}}(r,\theta'\phi')|^{2} r^{2} \sin\theta' d\theta' d\phi' , \qquad (7.3)$$

where r is the distance to the point of observation.

Results of this study are shown in Table 7.1. It is observed that $P_1 = 0.44$ W and $P_2 = 9.0 \times 10^{-5}$ W; i.e., only 0.02% of the available power is scattered by the 74 MHz–resonant dipoles, present at the image of the focus. Therefore blockage in the segment \overline{RA} by the image focus–located dipoles is negligible.

7.1.2 Effects of the EVLA 4 m System on the Performance of the Existing 90 cm System

In this section, we consider the effect of the EVLA 4 m system on the performance of the 90 cm system. The 90 cm system has already been analyzed in Section 4.2.2 of this dissertation, in the absence of the EVLA 4 m system. The sensitivity was found to be

Source of blockage	Image focus–	Double ring
	located dipoles	of dipoles
	4.5%	< 0.1%
aperture at 1.4 GHz due to blockage		
${\rm in \ segment \ } \overline{{\rm F'SR}}$		
Power through aperture	Scattered power	Scattered power
$P_1 = 0.44 \text{ W}$	$P_2 = 9.0 \times 10^{-5} \text{ W}$	$P_3 = 4.0 \times 10^{-4} \text{ W}$
Percent of total power P_1	0.02%	0.1%
scattered due to		
blockage in segment \overline{RA}		

Table 7.1. Comparison of blockage at 1.4 GHz presented by the image focuslocated dipoles and the double ring of strut–straddling dipoles.

1.39 kJy at 327 MHz. In order to find the effect of the 4 m system, the analysis of sensitivity of the 90 cm system is repeated in the presence of the crossed dipoles, located at the image of the focus. Again, the MoM+PO method is used to do the analysis. It is observed that the sensitivity of the 90 cm system at 327 MHz, in the presence of the image focus–located dipoles, is 1.40 kJy. Thus, there is not any significant (less than 1%) difference in the sensitivity of the 90 cm system, when the 4 m system is present. We assume this assessment to be consistent with that observed in practice, because to the best of our knowledge, no significant degradation of the sensitivity of the 90 cm system by the 4 m system has been reported.

7.2 Assessment of Blockage due to the Proposed Strut–Straddling System

In this section, we assess the blockage presented to pre–existing higher–frequency systems on the EVLA by the double ring of dipoles at z = -1 m and z = -2.25 m. Specifically, we compare the blockage introduced at 1.4 GHz and 327 MHz by this feed to that presented by the existing EVLA 4 m system.

7.2.1 Assessment of Blockage at 1.4 GHz due to the Proposed Strut–Straddling System

In this section, the blockage introduced at 1.4 GHz by the double ring of dipoles, employed at z = -1 m and z = -2.25 m on the EVLA, is assessed. In order to estimate the blockage in the first two segments – namely, $\overline{F'S}$ and \overline{SR} , the directivity of the antenna, with the double ring of dipoles in position, is computed using the method proposed in Section 7.1.1. The results are shown in Figure 7.2. The directivity is found to be 35.8 dB, which is about the same as the unblocked 1.4 GHz system and about 4.5% higher than the directivity observed with the feeds of the 4 m system in position.

The next part of our blockage estimation due to the double ring of dipoles is computing the blockage in the segment \overline{RA} . The power scattered by the double ring of dipoles P_3 is computed in the same manner as described in Section 7.1.1. In this case, the double ring of dipoles are illuminated with a plane wave with amplitude of $\cos^2(6^\circ)$ V/m. This amplitude is used because the location of the centers of the strut-straddling dipoles in the z = -1 m plane correspond to $\theta_f = 12^\circ$, and at this location the magnitude of $\mathbf{E_r}$ is found to be $\cos^2(6^\circ)$ V/m from Equation 7.1. The results are shown in Table 7.1. P_3 is calculated to be 4.0×10^{-4} W which corresponds to a loss of about 0.1% in the total available power (P_1).

It is observed upon comparison with image focus-located dipoles that in the segment \overline{RA} the amount of power scattered is increased (from 0.02% to 0.1%) when the image focuslocated dipoles are replaced by the double ring of dipoles; however, the scattered power constitutes a tiny fraction of the total available power. Therefore blockage in the segment \overline{RA} is not significant for either of the low frequency feed systems considered.

At this point, questions may arise about the support structure for the strut-straddling dipoles and the blockage associated with this support. The dipoles can be easily mounted and supported using non-conducting rope between the struts. Also, the cables carrying the signals from the dipoles can run inside the dipole arms to the struts and then be brought along the struts to the back of the main reflector or subreflector where they can be combined. In order to estimate the effect of the cables, we did a simple study where we model the cables by extending one end of each of the dipoles to the nearest struts, and the blocking in the path $\overline{F'SRA}$ is again computed. In the same manner as described in Section 7.1.1, the directivity of the antenna is calculated in the presence of the extended dipoles and is found to be 35.8 dB. It is observed that there is no significant increase in blockage in the $\overline{F'SR}$ segment for extending the dipoles. Also, there is not any significant increase (less than 0.1%) in the scattered power in the segment \overline{RA} when the dipoles are extended. Therefore the strut-straddling dipoles with their supports are estimated to reduce the blockage at 1.4 GHz relative to the image focus-fed systems.

7.2.2 Effects of the Proposed Strut–Straddling System on the Performance of the Existing 90 cm System

It has been shown that the double ring of dipoles at z = -1 m and z = -2.25 m offers performance comparable to that of the existing EVLA 4 m system, and presents reduced blockage (about 4.5% less than the EVLA 4 m system) at 1.4 GHz. We now consider whether the performance of the EVLA 90 cm system is affected by this distributed feed array. In order to find this out, the performance of the EVLA 90 cm system (in terms of SEFD) is estimated for the case when the double ring of strut-straddling dipoles are present in the antenna geometry, and the result is compared to the case when the existing 4 m system is present. The MoM+PO method described in Section 4.2.2 is used for the analysis.

The SEFD for the 90 cm system at 327 MHz is found to be 1.39 kJy when the double ring of dipoles are employed; i.e., the SEFD is almost the same as that observed for an unblocked 90 cm system, and when the dipoles of the 4 m system are present at the image of the focus.

7.3 Summary

The proposed double ring of dual length dipoles offers a useful level of sensitivity in the frequency ranges of 50–88 MHz and 110–250 MHz. Moreover, the blockage presented by this feed to pre–existing higher–frequency systems is less than that presented by the existing EVLA 4 m system; specifically, there is at least 4.5% improvement in the sensitivity of the L–band system when the image focus–located dipoles are replaced by the double ring of dual length dipoles. Therefore, the double ring of dual length dipoles is a promising candidate for the EVLA LFS.

Chapter 8

Conclusions

This dissertation described modifications of large reflector antennas for incorporating low frequency observing capabilities. The modifications take into account the issue of coexistence with pre–existing higher–frequency systems, specifically, by attempting to minimize blockage. The concepts are applied to an EVLA reflector antenna to demonstrate by example; however, the modifications are applicable to other large reflector antennas as well.

We conclude this dissertation in this chapter, with a summary of our findings and a list of possible future work in Section 8.1 and Section 8.2, respectively.

8.1 Findings

The principal findings of our research can be summarized as follows:

Electromagnetic Analysis of Reflector Antennas at Low Frequencies: High frequency analysis techniques, PO and UTD, should not be assumed valid for reflector antennas smaller than 3.25λ. The results of reflector antenna analysis performed with PO and UTD were compared with MoM in Section 2.4 to find this low frequency limit. Results are summarized in Figures 2.6 through 2.16. The advantage of knowing this limit is that PO and UTD, as alternatives to MoM, can be used at low frequencies, and

thereby computational burden can be reduced.

- MoM model of an EVLA Antenna: A wire grid model of an EVLA antenna was developed (Section 2.5), and the effects of the various components on antenna directivity and pattern were studied; e.g., it was identified (Figure 4.3) that the aperture efficiency of the EVLA 4 m system is reduced by about 18% when a feed dipole of the existing 4 m system is cross-aligned with the support wires as opposed to being co-aligned.
- Analysis of Sensitivity of Reflector Antenna Systems with Array Feeds: In Section 3.4, a methodology for analysis of sensitivity of reflector antenna systems employing array feeds was described. This method takes into account the correlation of external noise amongst the array elements, and also considers mutual coupling. It was shown in Figure 3.6 that the correlation of external noise can desensitize systems at low frequencies, and therefore needs to be considered while estimating sensitivity.
- Focus- or Near Focus-Fed Systems: Three sets of crossed dipole feeds namely, 74 MHz-, 150 MHz-, and 327 MHz-resonant crossed dipoles, located at the focus of a simple reflector, were demonstrated to cover the range 50–470 MHz with a use-ful level of sensitivity in Section 5.1. The estimated performance of these systems is summarized in Figure 5.1. These dipole feeds were then employed (in front of the subreflector) to feed the EVLA, and a baseline of low frequency performance was established in Section 5.2. The concept of dual length dipoles was introduced for collocating 74 MHz- and 150 MHz-resonant dipoles at the image of the focus. The dual length dipoles enhance observing capability in the range 110–250 MHz without introducing additional blockage over that of the 74 MHz-resonant dipoles. The results are summarized in Figure 5.5.
- Distributed Feed Array: A new low frequency feed concept namely, the distributed feed array was introduced in Section 6.2. The advantage of this feeding scheme

is that the blockage presented to pre-existing higher-frequency systems is less than that presented by the focus-located feeds. Single rings of dipoles, employed around the surface of the conceptual cone that encompasses the GO field, reflected from the main reflector toward the focus (i.e., at the border of Regions 1 and 2, as described in Section 6.1) of a reflector, were observed (Figure 6.9) to be a little worse (about 2-5 dB) than the arrays in the focal plane. Double rings of dipoles were observed to have improved performance (Figure 6.11) without making blockage significantly worse.

- LFS for the EVLA: The distributed feed array concept was applied to the EVLA to cover the range 50–250 MHz in Section 6.3. A double ring of strut-straddling "dual length" dipoles (concept shown in Figure 5.3) at z = -1 m and z = -2.25 m (z = 0 m is the plane containing the prime focus) was found (Figure 6.18) to offer improved performance than the image focus–located dipoles in the range 50–88 MHz and reduce blockage presented to the L–band system; moreover, observing capabilities are introduced in the range 110–250 MHz which is currently not available on the EVLA.
- Assessment of Blockage: A method to assess blockage presented to pre-existing higherfrequency systems by low frequency feeds was devised in Chapter 7 based on simple electromagnetic considerations. Using this method, it was demonstrated in Section 7.2.1 that the degradation in sensitivity of the EVLA L-band system is reduced by approximately 4.5% (Table 7.1) when the crossed dipoles of the existing EVLA 4 m system are replaced by the proposed double ring of dipoles. It was also shown that the performance of the existing P-band system is not significantly altered when the double ring of dipoles are employed.

Although the emphasis in this work was on the EVLA, the findings should be applicable to any large reflector antenna.

8.2 Future Work

Possible areas to extend the research presented in this dissertation are identified and described in this section. They are:

- Improvement of the MoM model of an EVLA antenna: A MoM model of the EVLA antenna was developed in Section 2.5 of this dissertation, and used for analyzing the performance of different low frequency systems. It was noted that all components of the model seem to play an important role (Table 2.3). Thus, the model might be improved. For example, the high frequency feed assembly was not modeled in detail; it was modeled as a rectangular box which encapsulates its volume. Also the subreflector mount model consisted of some dense sections in the middle and contained wires connected at acute angles, which is not recommended in NEC2 wire grid models.
- Receiver impedance design: The impedance looking out dipole feed terminals R_L has been assumed to be 100 Ω in estimating performance of reflector antenna systems. However, at low frequencies, higher values of R_L may offer larger usable bandwidth without significantly degrading sensitivity [30]. Investigation of the possibility of using higher-valued R_L 's is identified as future work.
- Improvement of accuracy of MoM+PO method: The MoM+PO method of Section 4.2.2 was devised to estimate sensitivity of reflector antenna systems at frequencies where a complete MoM analysis is not feasible; however, the method consisted of a number of steps which had a few percents of error associated with them. Improving the accuracy achieved with this method as well as further validation of the method might be possible areas of future work.
- Combiner design: Designing of combiners for the distributed feed array is an area of possible future work. In the studies presented in this dissertation the combining co-efficients were recalculated at each frequency, in lieu of an instantaneous broadband

method. In practice, however, the implementation of the coefficients would be narrowband if analog combining is used. An alternative option is to digitally combine the signals. Looking into these possibilities of combiner implementation is identified as future work.

- Introducing nulls in antenna patterns: The beamforming coefficients for combining the distributed feeds can be selected such that nulls are introduced in the reflector antenna pattern at certain locations. This technique is commonly proposed in focal plane arrays in order to suppress interference, and is equally applicable for the distributed feed arrays. Details of designing the coefficients are left as future work.
- Synthesis of dual orthogonal polarizations: The proposed EVLA strut straddling schemes, as described in the dissertation, produce only one polarization. Development of combining schemes to obtain dual orthogonal polarizations is identified as a possible area of future work.

Appendix A

Computation of the UTD Transition Function

In this section we describe the details of the computation of the UTD transition function F [32,53]. A plot of the magnitude and phase of F is shown in Figure A.1. If the argument of F is represented by X,

$$F(X) = 2j|\sqrt{X}|e^{jX}\int_{|\sqrt{X}|}^{\infty} e^{-j\tau^2}d\tau$$
(A.1)

In order to calculate F(X) we used small argument approximation when X < 0.3,

$$F(X) \approx \left[\sqrt{\pi X} - 2Xe^{j\frac{\pi}{4}} - \frac{2}{3}X^2e^{-j\frac{\pi}{4}}\right]e^{j(\frac{\pi}{4} + X)}$$
(A.2)

and large argument approximation when X > 5.5

$$F(X) \approx \left(1 + j\frac{1}{2X} - \frac{3}{4}\frac{1}{X^2} - j\frac{15}{8}\frac{1}{X^3} + \frac{75}{16}\frac{1}{X^4}\right) \tag{A.3}$$



Figure A.1. (a) Magnitude and (b) Phase of the UTD transition function.

For arguments in the range 0.3 < X < 5.5 we used interpolation using Table A.1 as:

$$F(X) \approx X_n + A_n(X - X_n) \tag{A.4}$$

and for arguments X < 0

$$F(X) = F^*(-X)$$
, (A.5)

where * denotes conjugate operation.

X_n	$F(X_n)$	A_n
0.3	0.5729 + j0.2677	0.0000 + j0.0000
0.5	0.6768 + j0.2682	0.5195 + j0.0025
0.7	0.7439 + j0.2549	0.3355 - j0.0665
1.0	0.8095 + j0.2322	0.2187 - j0.0757
1.5	0.8730 + j0.1982	0.1270 - j0.0680
2.3	0.9240 + j0.1577	0.0638 - j0.0506
4.0	0.9658 + j0.1073	0.0246 - j0.0296
5.5	0.9797 + j0.0828	0.0093 - j0.0163

Table A.1. Values used for interpolation for the computation of F(X).

Appendix B

Determination of Wire Grid Model Parameters for Method of Moments Modeling of a Large Reflector Antenna

In this appendix, we describe an experiment for selecting a reasonable wire grid to model a reflector antenna for MoM analysis. Parameters defining the grid are: 1) radius of wire in the grid, r; 2) spacing between the wires in the grid, d; and 3) number of segments per wire, s. This in turn defines the segment length l, where l = d/s. NEC2 has the following guidelines for selecting values for these parameters [35]: a) $l < \lambda/10$, where λ is the wavelength; b) $r < \lambda/20\pi$; and c) l > 2r. A separate issue is that the total number of segments N should be less than about 11,000 so that it can easily fit in a reasonable size (4 GB) RAM.

In this experiment, a wire grid model of a 6.5λ -diameter (26 m at 74 MHz) flat disk is tested as a ground plane, one quarter wavelength beneath a half-wave dipole. The disk is located in the xy plane, centered around the z-axis. The grid parameters d and r are varied, and the patterns are compared to determine leakage (backlobe level). Figure B.1 shows the results. It can be observed that $d = 0.1\lambda$ and $r = \lambda/75$ results in a model for which a very small backlobe of -20 dB is found, and therefore these parameters are judged to suitably represent a solid surface.

In order to confirm the validity of these model parameters, the pattern of a half-wave dipole located quarter wavelength over this wire grid disk at 74 MHz is compared with the pattern obtained using image theory, and also to the pattern obtained by NEC2 assuming a perfect infinite ground. Figure B.2 shows the result. It can be seen that the wire grid result is indeed consistent with the other two results. We infer from this that $d = 0.1\lambda$ and $r = \lambda/75$ is an appropriate wire grid for modeling a 26 m diameter reflector antenna at 74 MHz.



Figure B.1. E-plane pattern for a half-wave dipole $\lambda/4$ above the wire grid ground for different grid parameters. θ is the zenith angle.



Figure B.2. E-plane pattern for a half-wave dipole $\lambda/4$ above different ground models. Grid parameters used for the disk model are $d = 0.1\lambda$, s = 1, $r = \lambda/75$, and N = 7272.

Appendix C

EVLA Subreflector Model

In this appendix we describe how a polynomial equation is derived to define the shape of the Cassegrain subreflector on the EVLA. The Cassegrain subreflector on the EVLA is 2.35 m in diameter. The subreflector is not symmetrical about the axis of the main reflector; instead, its axis is tilted by about 2.4° from the axis of the main reflector in order to point toward specific feeds arranged in a ring in the high–frequency feed assembly. The shape parameters of the subreflector are not known in detail. However, Jim Ruff from NRAO provided a CAD drawing of the subreflector. Using this drawing, data points from one cross section of the subreflector are collected, and used to fit a 5th order polynomial. The resulting polynomial is

$$y = 0.0003x^5 + 0.0379x^4 - 0.0004x^3 + 0.2875x^2 - 0.0001x - 0.0003$$
(C.1)

where x, y are in meters. Figure C.1 shows that the polynomial model follows the data points from the drawing reasonably well.

The order of the polynomial equation is determined such that the residuals have values less than 0.01λ at 74 MHz. "Residual" is the difference between the data point and the predicted value from the model. Table C.1 shows that the residual for our model is well



Figure C.1. Comparison between the data points taken from the subreflector drawing and our polynomial model.

within the limit specified.

	y	
CAD	Model	$\mathrm{Residual}/\lambda$
m	m	at 74 MHz
0.463	0.463	0.0000
0.366	0.357	0.0002
0.253	0.252	0.0005
0.157	0.158	0.0003
0.078	0.078	0.0003
0.018	0.019	0.0002
0.004	0.003	0.0004
0.018	0.019	0.0002
0.078	0.078	0.0002
0.157	0.158	0.0003
0.253	0.252	0.0005
0.366	0.357	0.0002
0.463	0.463	0.0000

Table C.1. The residual from the mathematical model of the subreflectorshape.

Appendix D

Calculation of Antenna Temperatures for Different Pointing Directions

This appendix provides details of the antenna temperature (T_A) calculations used in the study described in Section 3.2. The calculations are for a simple reflector antenna with D = 25 m and f/D = 0.36, fed separately from the focus with several half-wave dipole feeds – namely, dipoles half-wavelength long at 29 MHz, 38 MHz, 74 MHz, and 88 MHz. These results assume $T_{int} = 250$ K and $Z_b = 100 \Omega$. Antenna temperatures are calculated for two different pointing directions – namely, 1) A point (RA: 19 h, DEC: -25° , $T_b = 199 \times 10^3$ K at 38 MHz) in Sagittarius, and 2) a point (RA: 12.5 h, DEC: 13.5° , $T_b = 7000$ K at 38 MHz) in Virgo. For additional details, see Section 3.2.

In Tables D.1 through D.10, we show the resulting antenna temperatures. Also shown in the tables are the IME e_I , and the system temperature T_{sys} , defined as $T_{sys} = e_I T_A + T_{int}$. Note that it is also possible to define an "input-referred" system temperature $T_A + T_{int}/e_I$ which may be more useful for astronomical interpretation; this value is simply T_{sys} as given in the tables, divided by e_I .

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	445580.0 K	938.2 K	1190.0 K
20 MHz	76090.0 K	2670.0 K	2930.0 K
38 MHz	14800.0 K	$2550.0 { m K}$	$2780.0 { m K}$
74 MHz	2710.0 K	371.6 K	621.6 K
88 MHz	1740.0 K	1690.0 K	1940.0 K

Table D.1. Antenna temperature and system temperature data for the reflector—only model, fed with a 29 MHz half—wave dipole, pointing toward Sagittarius.

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	186200.0 K	392.1 K	642.1 K
20 MHz	31800.0 K	1120.0 K	1370.0 K
38 MHz	6190.0 K	1060.0 K	1310.0 K
74 MHz	1130.0 K	155.3 K	405.3 K
88 MHz	727.1 K	704.5 K	954.5 K

Table D.2. Antenna temperature and system temperature data for the reflector–only model, fed with a 29 MHz half–wave dipole, pointing toward Virgo.

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	445780.0 K	$515.5 { m K}$	1190.0 K
20 MHz	76120.0 K	$551.9 { m K}$	2930.0 K
38 MHz	14810.0 K	13800.0 K	2780.0 K
74 MHz	2710.0 K	371.0 K	621.6 K
88 MHz	1740.0 K	206.1 K	1940.0 K

Table D.3. Antenna temperature and system temperature data for the reflector–only model, fed with a 38 MHz half–wave dipole, pointing toward Sagittarius.

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	186600.0 K	$215.8 { m K}$	$465.8~\mathrm{K}$
20 MHz	31900.0 K	231.0 K	481.0 K
38 MHz	6200.0 K	$5780.0 { m K}$	6030.0v
74 MHz	1130.0 K	$155.3~\mathrm{K}$	405.3 K
88 MHz	728.5 K	86.3 K	336.3 K

Table D.4. Antenna temperature and system temperature data for the reflector–only model, fed with a 38 MHz half–wave dipole, pointing toward Virgo.

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	445750.0 K	134.7 K	384.7 K
20 MHz	76110.0 K	198.5 K	448.5 K
38 MHz	14810.0 K	3070.0 K	332780.0 K
74 MHz	2710.0 K	550.2 K	800.2 K
88 MHz	1740.0 K	256.6 K	506.6 K

Table D.5. Antenna temperature and system temperature data for the reflector–only model, fed with a 47 MHz half–wave dipole, pointing toward Sagittarius.

Freq	T_A	$e_I T_A$	T_{sys}
$10 \mathrm{~MHz}$	186600.0 K	$56.4~\mathrm{K}$	$306.4~\mathrm{K}$
20 MHz	31900.0 K	83.1 K	333.1 K
38 MHz	6200.0 K	$6450.0 { m K}$	6030.0 K
74 MHz	1130.0 K	1380.0 K	405.3 K
88 MHz	728.7 K	978.7 K	107.4 K

Table D.6. Antenna temperature and system temperature data for the reflector–only model, fed with a 47 MHz half–wave dipole, pointing toward Virgo.

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	445750.0 K	52.9 K	302.9 K
20 MHz	76110.0 K	68.9 K	318.9 K
38 MHz	14810.0 K	$525.5 { m K}$	775540.0 K
74 MHz	2710.0 K	1220.0 K	1470.0 K
88 MHz	1740.0 K	436.9 K	686.9 K

Table D.7. Antenna temperature and system temperature data for the reflector–only model, fed with a 60 MHz half–wave dipole, pointing toward Sagittarius.

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	186600.0 K	22.2 K	$272.2~{ m K}$
20 MHz	31900.0 K	28.9 K	278.9 K
38 MHz	6200.0 K	220.1 K	470.1 K
74 MHz	1130.0 K	512.2 K	762.2 K
88 MHz	728.7 K	182.9 K	432.9 K

Table D.8. Antenna temperature and system temperature data for the reflector–only model, fed with a 60 MHz half–wave dipole, pointing toward Virgo.

Freq	T_A	$e_I T_A$	T_{sys}
$10 \mathrm{~MHz}$	$445750.0 { m K}$	24.2 K	274.2 K
20 MHz	76110.0 K	29.7 K	279.7 K
38 MHz	14810.0 K	167.1 K	417.1 K
74 MHz	2710.0 K	2620.0 K	2800.0 K
88 MHz	1740.0 K	943.4 K	1120.0 K

Table D.9. Antenna temperature and system temperature data for the reflector–only model, fed with a 74 MHz half–wave dipole, pointing toward Sagittarius.

Freq	T_A	$e_I T_A$	T_{sys}
10 MHz	186600.0 K	10.1 K	$260.1 { m K}$
20 MHz	31900.0 K	12.4 K	$262.4~\mathrm{K}$
38 MHz	6200.0 K	69.9 K	319.9 K
74 MHz	1130.0 K	1100.0 K	1350.0 K
88 MHz	728.7 K	395.1 K	645.1 K

Table D.10. Antenna temperature and system temperature data for the reflector–only model, fed with a 74 MHz half–wave dipole, pointing toward Virgo.

Appendix E

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