UNIVERSIDAD POLITÉCNICA DE MADRID
ESCUELA TÉCNICA SUPERIOR DE INGENIEROS DE TELECOMUNICACIÓN

NEW ADVANCES ON MULTI-FREQUENCY
AND MULTI-BEAM REFLECTARRAYS
WITH APPLICATION TO
SATELLITE ANTENNAS IN KA-BAND

TESIS DOCTORAL

Eduardo María Martínez de Rioja del Nido

Ingeniero de Telecomunicación

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DEPARTAMENTO DE SEÑALES, SISTEMAS
Y RADIOCOMUNICACIONES

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Abstract

Current high throughput satellites (HTS) in Ka-band are required to provide multiple spot beam coverage based on frequency and polarization reuse, both in transmission (Tx, 19.2-20.2 GHz) and reception (Rx, 29-30 GHz). A four colour scheme with two frequencies and two polarizations is normally used, in which adjacent spots must be generated in a different frequency and/or polarization. The design of multi-beam antennas for Ka-band HTS systems must cope with some challenging requirements: generation of a large number of beams (normally between 50 and 100), very small separation between adjacent spots (a typical value is 0.56°), low spillover, etc. To confront these stringent conditions, most of current HTS systems carry four reflector antennas on board the satellite, each reflector being responsible for generating all the beams in the same frequency and polarization (same colour) in a single feed per beam basis. The problem of this configuration has to do with the accommodation of the four reflectors in the satellite. A reduction in the number of apertures required to provide multi-spot coverage would result in significant savings in the cost, weight and volume of the antenna farm in communication satellites that operate in Ka-band.

The motivation of this thesis has been to provide new advances on the design of multi-frequency and multi-beam reflectarray antennas with application to multiple spot beam satellites in Ka-band. In this respect, the thesis can be divided into two main parts: the first part on reflectarrays operating at two different frequencies, and the second for the developing of design techniques to improve the performance of multi-beam antennas.

The first part of the thesis contains the description of a novel reflectarray cell to operate in dual-linear polarization (LP) at two separate frequencies (enabling independent phasing in each polarization and frequency), as well as the design of dual-band reflectarrays to provide independent beams in each polarization and frequency band, including the manufacturing and testing of a 25-cm reflectarray demonstrator to operate in dual polarization (linear or circular) in Ku and Ka bands. The reflectarray element proposed for independent operation in dual-LP at two separate frequencies consists of a two-layer configuration with two orthogonal sets of stacked parallel dipoles. Each set, that adjusts the phase in one polarization, is composed of five parallel

dipoles on the lower layer and three additional parallel dipoles stacked above the previous ones and printed on top of a second dielectric sheet. The geometrical parameters of the cell have been adjusted to operate, first, at Tx frequencies in Ku and Ka bands (12 and 20 GHz), and then, at Tx and Rx frequencies in Ka-band (20 and 30 GHz). The proposed two-layer configuration allows to perform separate design processes for each reflectarray layer: first, the lengths of the lower dipoles are adjusted to match the required phases at the lower frequency, and then, the lengths of the upper dipoles are adjusted to introduce the required phase-shift at the higher frequency. This step-by-step procedure allows for a simpler and computationally faster design process. Moreover, the design is carried out independently for each linear polarization, by adjusting the set of dipoles associated to each polarization.

A Ku/Ka-band reflectarray demonstrator of 25-cm diameter has been designed, manufactured and tested, in order to validate the multi-frequency reflectarray cells and the design technique. The proposed reflectarray permits an independent optimization of the radiation patterns for Ku and Ka bands, as well as a proper accommodation of the feed chains for each frequency band. This concept can be applied to design a satellite transmit antenna which would be able to fulfill independent requirements at each frequency and/or polarization (for example, generation of a contoured beam in Ku-band and multiple spots in Ka-band) by properly designing the elements on each reflectarray layer, using different feed chains for each mission. Moreover, the manufacturing using the technology for multi-layer printed circuits and low profile of the sandwich would lead to significant savings in the costs, weight and volume of the antenna farm for current satellite systems that operate in Ku and Ka bands, thanks to the reuse of the same aperture for two different missions.

The second part of the thesis comprises the development of a bifocal design technique for dual reflectarray and dual transmitarray configurations, and its application to the design of multi-beam antennas in Ka-band. The aim of the bifocal technique is twofold, to improve the multi-beam performance of the antenna and to provide a certain degree of reduction in the angular separation between adjacent beams for a multi-spot coverage from a satellite. Two different approaches have been considered: starting from an axially-symmetrical geometry which allows the rotation of a 2D bifocal design around the symmetry axis, and implementing a general 3D bifocal method that directly provides the required phases on both reflectarrays in the selected antenna configuration.

First, a bifocal design procedure has been developed for both dual reflectarray and dual transmitarray antennas by starting from an axially-symmetrical geometry with the two reflectarrays/transmitarrays placed in parallel planes. A 2D bifocal design performed in the offset plane by means of an iterative ray-tracing routine is rotated around the symmetry axis, and then, both centered and offset configurations are possible by choosing specific portions of the revolution surfaces. In the case of offset dual reflectarray configurations, both reflectarrays can be tilted a certain angle to obtain smoother phase distributions. For this purpose, a novel phase adjustment routine has been implemented to compensate the tilting and maintain the bifocal characteristic of the original design. On the other hand, the design with transmitarrays provides some advantages, such as lower sensitivity to surface deformations, absence of blockage and use of centered geometries with a focal ring. These advantages are achieved at the cost of a larger antenna volume. Hence, different dual transmitarray configurations have been studied to try to reduce the antenna volume, such as placing the feeds close to the first transmitarray (to integrate both elements on the same sub-system), or reducing the distance between the transmitarrays (to hold them with the same supporting structure).

Secondly, a general tridimensional bifocal technique for dual reflectarray antennas has been developed, which makes it possible the direct synthesis of the required phase distributions on each reflectarray without imposing geometrical restrictions in the antenna configuration. The proposed 3D bifocal method is based on an iterative 3D ray-tracing routine that provides a grid of points on the surface of each reflectarray and the values of the partial derivatives of the phase associated to those points. The partial phase derivatives are interpolated, and then, properly integrated to obtain the bifocal phase functions required on each reflectarray.

A preliminary study on the bifocal technique for the design of multi-beam satellite antennas in Ka-band has been carried out, considering three different degrees of reduction in the beam spacing with respect to the equivalent monofocal antenna: high beam spacing reduction (by a factor of 2, in order to provide adjacent beams with 0.56° separation), low beam spacing reduction (by a factor of 1.1 or 1.2), and no beam spacing reduction. The results show that the bifocal technique allows to provide the required 0.56° spacing by using non-overlapping feeds, but at the cost of a lower radiation efficiency of the bifocal antenna (the main reflectarray should be significantly oversized). The most interesting case is that for low beam spacing reduction, which

allows to obtain closer beams with non-overlapping feeds, at the same time as improving the performance of the extreme beams and providing reasonable values of gain and radiation efficiency.

A bifocal dual reflectarray antenna demonstrator with a main reflectarray of 57-cm has been designed, manufactured and tested in order to validate the proposed 3D bifocal technique. The demonstrator has been designed to operate in dual-LP in the 19.2-20.2 GHz band, but the technique can be also used to generate adjacent beams in dual-circular polarization by using a sequential rotation method. The results of the measurements show the capability of the bifocal technique to reduce the beam spacing and provide a better multi-beam performance than the equivalent single-focus antenna (particularly, the gain and side lobes are improved for the most external beams). The first factor will allow to reduce the antenna size with respect to conventional reflectors to provide the same beam spacing. Moreover, the fabrication of the bifocal dual reflectarray antenna involves the same conventional processes used for printed reflectarrays, without any need of custom moulds, allowing a significant reduction of manufacturing time and cost, particularly when compared with bifocal dual reflectors that require expensive custom moulds for the two shaped reflectors.

Finally, a bifocal dual reflectarray antenna with an elliptical main reflectarray has been proposed to provide all the required spots (four colours) for transmission from a geostationary satellite in Ka-band, in order to substitute the four conventional antennas (one for each colour). The bifocal technique has been applied with a high degree of beam spacing reduction to produce adjacent beams with 0.56° separation in the offset plane, while using a monofocal phase condition in the orthogonal plane (beam spacing around 1.1°). The interleaved beams required for providing full multi-spot coverage are generated in the orthogonal polarization. This solution presents some advantages with respect to other configurations that use a single oversized reflector to provide multi-spot coverage, as it requires a smaller aperture size and a lower number of feeds. The use of flat reflectarray panels, which can be fabricated by the same conventional and relatively inexpensive processes used for printed circuits, allows for a more efficient packaging and deployment on the satellite. Moreover, the design of a Tx/Rx multiple spot beam satellite antenna can be addressed by the use of appropriate dual-frequency reflectarray cells that will enable independent phasing at Tx and Rx frequencies in Ka-band.

Resumen

Los actuales satélites de comunicaciones con alta capacidad de datos en banda Ka deben proporcionar una cobertura celular formada por múltiples haces parcialmente imbricados con reutilización de frecuencia y polarización, tanto en la banda de transmisión (Tx, 19.2-20.2 GHz) como en recepción (Rx, 29-30 GHz). Normalmente se emplea un esquema de cuatro colores que combina dos frecuencias y dos polarizaciones, en el que los haces adyacentes se generan a una frecuencia y/o polarización distintas. El diseño de antenas de haces múltiples para los sistemas de satélites en banda Ka debe afrontar una serie de requisitos muy exigentes: generación de un elevado número de haces (habitualmente entre 50 y 100), separación muy pequeña entre haces adyacentes (un valor típico es 0.56°), bajas pérdidas, etc. Para hacer frente a estas especificaciones, la mayoría de los satélites llevan embarcadas cuatro antenas reflectoras, cada una de ellas responsable de generar todos los haces en una misma frecuencia y polarización (mismo color), utilizando un alimentador por haz. El problema de esta configuración viene dado por el hecho de tener que acomodar los cuatro reflectores en el satélite. Una reducción del número de antenas necesarias para proporcionar la cobertura celular traería consigo un importante ahorro en los costes, peso y volumen del conjunto de antenas para los satélites de comunicaciones que operan en banda Ka.

El objetivo de esta tesis es proporcionar nuevos avances en el diseño de antenas reflectarray multi-frecuencia y multi-haz con aplicación a los satélites de comunicaciones para cobertura celular en banda Ka. A este respecto, la tesis puede dividirse principalmente en dos partes: la primera sobre reflectarrays que operan a dos frecuencias distintas, y la segunda sobre técnicas de diseño para mejorar las prestaciones de las antenas de haces múltiples.

La primera parte de la tesis contiene la descripción de una nueva celda reflectarray para operar en doble polarización lineal a dos frecuencias relativamente separadas (permitiendo introducir un desfase independiente en cada polarización y frecuencia), así como el diseño de antenas reflectarray de doble banda que proporcionan haces

independientes en cada polarización y banda de frecuencia, incluyendo la fabricación y medida de un demostrador de antena reflectarray de 25 cm para operar en doble polarización (lineal o circular) en las bandas Ku y Ka.

El elemento reflectarray propuesto para operar de manera independiente en doble polarización lineal a dos frecuencias distintas consiste en una configuración de dos capas con dos conjuntos ortogonales de dipolos paralelos apilados. Cada conjunto, que permite ajustar la fase en una polarización, está compuesto de cinco dipolos paralelos en la capa inferior y tres dipolos paralelos más apilados sobre los anteriores e impresos en la cara superior de una segunda lámina de dieléctrico. Se han ajustado los parámetros geométricos de cada celda para operar, primero, a las frecuencias de Tx en las bandas Ku y Ka (12 y 20 GHz), y después, a las frecuencias de Tx y Rx en banda Ka (20 y 30 GHz). La configuración de dos capas propuesta permite realizar por separado el diseño de cada capa reflectarray: primero se ajustan las longitudes de los dipolos inferiores para proporcionar las fases necesarias a la frecuencia más baja, y después, se ajustan las longitudes de los dipolos superiores para introducir el desfase requerido a la frecuencia superior. Este procedimiento en dos pasos hace posible un proceso de diseño más sencillo y computacionalmente más rápido. Además, el diseño se lleva a cabo de manera independiente para cada polarización lineal, ajustando por separado el conjunto de dipolos asociado a cada polarización.

Para validar tanto las celdas reflectarray multi-frecuencia como la técnica de diseño, se ha diseñado, fabricado y medido una antena reflectarray de 25 cm de diámetro para operar en las bandas Ku y Ka. El reflectarray propuesto permite optimizar de manera independiente los diagramas de radiación en cada una de las bandas de frecuencia, así como una colocación apropiada de las bocinas alimentadoras para cada frecuencia. Este concepto puede aplicarse al diseño de antenas transmisoras para satélite que serían capaces de satisfacer distintas especificaciones en cada banda de frecuencias y/o polarización (por ejemplo, generación de un haz conformado en banda Ku y de múltiples haces en banda Ka) mediante un diseño adecuado de los elementos en cada capa del reflectarray, utilizando alimentadores diferentes para cada misión. Además, la fabricación mediante tecnología de circuitos impresos multicapa y el bajo perfil del sándwich, permitirían un importante ahorro en el coste, peso y volumen del sistema de antenas para los satélites de comunicaciones que operan en las bandas Ku y Ka, gracias a la reutilización de la misma antena para dos misiones diferentes.

La segunda parte de la tesis comprende el desarrollo de una técnica de diseño bifocal para antenas de doble transmitarray o doble reflectarray, y su aplicación al diseño de antenas de haces múltiples en banda Ka. El propósito de utilizar la técnica bifocal es doble: por un lado, mejorar las prestaciones de la antena para la generación de haces múltiples, y por otro, reducir la separación entre haces adyacentes para producir una cobertura celular desde el satélite. Para abordar este problema, se han considerado dos métodos distintos: partir de una configuración con simetría axial que permita rotar un diseño bifocal en 2D alrededor del eje de simetría, e implementar un algoritmo general de diseño bifocal en 3D que proporcione directamente las distribuciones de fase requeridas en los dos reflectarrays en la configuración de antena seleccionada.

En primer lugar, se ha desarrollado un método de diseño bifocal para antenas de doble reflectarray o doble transmitarray a partir de una configuración con simetría axial en la que los dos reflectarrays/transmitarrays están situados en planos paralelos. Un diseño bifocal 2D realizado por medio de una rutina iterativa de trazado de rayos se rota alrededor del eje de simetría, y después es posible diseñar tanto configuraciones centradas como descentradas sin más que seleccionar porciones específicas de las superficies de revolución obtenidas. En el caso de configuraciones de doble reflectarray descentradas, ambos reflectarrays pueden inclinarse un cierto ángulo para obtener distribuciones de fase más suaves. Para ello, se ha implementado una rutina de ajuste de la fase que permite compensar la inclinación de los reflectarrays manteniendo la característica bifocal del diseño original. Por otro lado, el diseño con transmitarrays proporciona algunas ventajas, como menor sensibilidad a las deformaciones de la superficie, ausencia de bloqueo y utilización de geometrías centradas con un anillo focal. Estas ventajas se consiguen a costa de que la antena ocupe un volumen mayor. Por esta razón, se han estudiado diferentes configuraciones de doble transmitarray para tratar de reducir el volumen de la antena, como situar los alimentadores próximos al primer transmitarray (para integrar ambos elementos en el mismo subsistema), o reducir la distancia entre los dos transmitarrays (de manera que compartan la misma estructura de soporte).

En segundo lugar, se ha desarrollado una técnica general de diseño bifocal en tres dimensiones, que permite la síntesis directa de las distribuciones de fase en cada reflectarray sin imponer ningún tipo de restricción geométrica en la configuración de la antena. El método bifocal tridimensional propuesto está basado en un procedimiento

iterativo de trazado de rayos en 3D que proporciona una malla de puntos en la superficie de cada reflectarray, así como los valores de las derivadas parciales de la fase asociados a esos puntos. Las derivadas parciales de la fase se interpolan e integran de forma apropiada para obtener las funciones de fase bifocales requeridas en cada reflectarray.

Se ha llevado a cabo un estudio preliminar sobre la aplicación de la técnica bifocal al diseño de antenas de haces múltiples para satélites en banda Ka, considerando tres grados diferentes de reducción de la separación entre haces con respecto a la antena monofocal equivalente: un grado de reducción elevado (en un factor 2, para proporcionar haces adyacentes con 0.56° de separación), un grado de reducción pequeño (en un factor de 1.1 o 1.2) y sin reducción del espaciado entre haces. Los resultados obtenidos muestran que la técnica bifocal permite proporcionar la separación requerida de 0.56° utilizando alimentadores contiguos, pero a costa de una baja eficiencia de radiación de la antena bifocal (el reflectarray principal debería sobredimensionarse considerablemente). El caso más interesante es el de una reducción pequeña del espaciado entre haces, que permite obtener haces más próximos con alimentadores no superpuestos, al mismo tiempo que se mejoran los resultados de los haces extremos de la cobertura y se alcanzan valores de ganancia y eficiencia de radiación razonables.

Un demostrador de antena bifocal de doble reflectarray con un reflectarray principal de 57 cm ha sido diseñado, fabricado y medido para validar la técnica bifocal 3D propuesta. El demostrador se ha diseñado para operar en doble polarización lineal en la banda comprendida entre 19.2 y 20.2 GHz, pero la técnica de diseño puede ser utilizada igualmente para generar haces adyacentes en doble polarización circular mediante una técnica de rotación secuencial. Los resultados de las medidas demuestran la capacidad de la técnica bifocal para reducir la separación entre haces y para proporcionar unas mejores prestaciones que la antena monofocal equivalente (en concreto, se mejoran la ganancia y el nivel de lóbulos secundarios para los haces extremos de la cobertura). El primer factor permitirá reducir el tamaño requerido para la antena con respecto a los reflectores convencionales que proporcionan la misma separación entre haces. Además, la fabricación de la antena bifocal de doble reflectarray conlleva los mismos procesos convencionales usados para los reflectarray impresos, sin necesidad de emplear moldes específicos para cada caso, posibilitando una importante reducción del tiempo y los costes de fabricación, especialmente si se compara con las antenas bifocales de doble

reflector, que requieren de costosos moldes metálicos para conformar los dos reflectores.

Por último, se ha propuesto una antena bifocal de doble reflectarray con el reflectarray principal elíptico que proporciona todos los haces necesarios (los cuatro colores) para operar en transmisión desde un satélite geoestacionario en banda Ka, con objeto de sustituir a los cuatro reflectores utilizados actualmente (uno para cada color). Se ha aplicado la técnica bifocal con un alto grado de reducción del espaciado entre haces para producir haces adyacentes con 0.56° de separación en el plano de simetría, mientras que en el plano ortogonal se utiliza una condición de fase monofocal (1.1° de separación entre haces). Los haces restantes para formar la cobertura se generan en la polarización ortogonal. La solución propuesta presenta algunas ventajas con respecto a otras configuraciones que emplean un único reflector sobredimensionado para generar la cobertura celular, ya que requiere una apertura de menor tamaño y un menor número de alimentadores. El uso de reflectarrays planos, que pueden ser fabricados mediante los mismos procesos convencionales y de bajo coste que los circuitos impresos, permite implementar mecanismos más eficientes de despliegue en el satélite. El diseño de una antena de haces múltiples para satélites en banda Ka que operen en Tx y Rx puede llevarse a cabo mediante el uso de celdas reflectarray multi-frecuencia que permitan introducir desfases independientes en las frecuencias de transmisión y recepción.

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Chapter 1

Introduction

1.1 Reflectarray antennas and their applications

Reflectarray antennas are composed of an arrangement of radiating elements printed on a flat or curved reflecting surface, with an illuminating feed antenna. The elements reradiate the incident field from the feed with a certain phase-shift, in order to generate collimated or shaped beams. This task is commonly accomplished by varying the dimensions of the printed elements. The reflectarray concept was first introduced in the 1960s, based on variable-length waveguide elements ended in a short circuit [1]. Later on, in the 1990s, the interest in reflectarray antennas was renewed with the development of photo-etching techniques, which allowed the fabrication of reduced size reflectarrays with the same conventional processes used for printed circuits [2], [3].

Reflectarrays borrow several advantages from reflector and array antennas, as wide-angle beam scanning and high radiation efficiency, while avoiding the employ of complex and expensive feeding and beamforming networks [4]. Moreover, reflectarrays are low profile antennas, which require a lower consumption of volume and weight resources than conventional reflectors. Added to this are their ease of integration and relatively low manufacturing costs, thanks to the use of the printed circuit technology. For these reasons, reflectarrays are suitable for most of high-gain applications, such as radar and long distance communications.

The use of plane reflective surfaces instead of shaped curved reflectors represents a reduced cost solution for satellite antennas. Reflectarrays are able to produce contoured beams for continental coverage just by properly adjusting the dimensions of the printed elements [5]-[7], thus using the same manufacturing processes based on photo-etching.

This fact reduces the complexity and costs of their fabrication when compared with conventional shaped reflectors for the same applications, which require expensive custom moulds that are manufactured specifically for each mission. Furthermore, the capability of reflectarrays to implement independent phase-shifts in each polarization [8], [9], or at different frequencies [10], [11], makes it possible that a single aperture can generate independent beams in different frequencies and/or polarizations, when the antenna is illuminated by a single feed. This can be applied to the design of multi-beam antennas, where the reflectarray is used to generate multiple spot beams with a cluster of feeds placed near to the focal area of the antenna [12], [13]. In addition, the reflectarrays' low profile allows for a more efficient packaging and an easier deployment on the satellite. For example, two inflatable reflectarrays were developed by NASA's Jet Propulsion Laboratory to operate at X-band and Ka-band respectively, using a thin-membrane reflectarray surface, as reported in [14]. Thanks to all the previous factors, reflectarrays have become a potential alternative for space antennas in satellite systems working in Ku and Ka bands.

Furthermore, reflectarray antennas have been proposed for their use in dual reflector configurations, in which the sub- [15], the main [16] or both reflectors [17] can be reflectarrays, in order to improve some performances of single reflector antennas. A dual reflectarray antenna provides control of the phase on both reflective surfaces, allowing more degrees of freedom for the design. If both reflectarrays are implemented in flat panels, the antenna can be folded and deployed in a more efficient way than a dual reflector system [18]. In the recent years, several dual reflectarray antennas have been proposed for different applications, such as the provision of shaped beams for mobile service base stations [19], the design of monopulse antennas for radar tracking systems [20], the generation of multiple beams for automotive radar antennas [17].

The integration of PIN diodes, micro-electro-mechanic (MEM) switching devices or varactor diodes at the reflectarray element level [21], [22] has led to the implementation of reconfiguration capabilities in the reflectarrays, making it possible to modify some of their radiation properties (for example, the beam pointing direction). Reconfigurable reflectarrays based on Liquid Cristal (LC) technology [23]-[25] are another alternative for beam scanning or beam switching applications in the millimeter and submillimeter wave ranges, as they are easily reconfigurable by a change in the bias voltage applied to the liquid crystal. In [26], the authors propose the implementation of an LC-based

reflectarray as the sub-reflector in a dual reflector configuration to provide electronic beam steering for radiometer systems operating at 94 GHz.

The main limitation of reflectarray antennas is their narrow operational bandwidth, which is mainly a consequence of two factors [4]: the element bandwidth (which is the most restrictive factor in moderate size reflectarrays) and the variation with frequency of the spatial phase delay (critical for large size reflectarrays and small F/D ratios). However, a significant effort has been done in the last years to improve this constraint. Some of the proposed solutions involve the use of single-layer broadband reflectarray elements, such as concentric cross loops [27], the Phoenix cell [28] and coplanar parallel dipoles [29]. These configurations provide more degrees of freedom to control the phase in the reflectarray cell and increase the operational bandwidth. For example, 27% bandwidth (for 2 dB gain variation) centered at 12.75 GHz has been achieved in both X and Y linear polarizations for a 45-cm reflectarray demonstrator designed with elements based on orthogonal sets of three coplanar parallel dipoles [30]. Reflectarray cells formed by several stacked layers of variable-size patches have been proposed for bandwidth improvement [6], [31], [32], although they have the drawback of higher manufacturing costs and weight for the resulting antenna, due to the larger number of layers. The use of electrically small reflectarray elements with a period less than $\lambda/2$ (sub-wavelength elements, forming an artificial impedance surface) allows to achieve more than 20% bandwidth (1 dB variation) for a single layer reflectarray with rectangular patches [33], [34].

Fig. 1-1 Two different strategies for achieving broadband operation: (a) a single-layer broadband element (the Phoenix cell) [28], (b) two stacked layers of rectangular patches [31].

A different technique consists on the design with true-time delay compensation, using reflectarray elements that provide a wide range of phase variation (several times 360°) to compensate the variations in the differential spatial phase delay [35], [36].

These variations can also be compensated by the use of faceted configurations approximating a paraboloid [37], [38], or by directly printing the reflectarray elements on a parabolic surface [39]. It has been shown that these configurations improve the reflectarray bandwidth, but the cross-polarization is penalized by the inclination of the panels or by the offset parabolic surface. Finally, phase optimization procedures at several frequencies have been applied to reflectarrays designed with multi-resonant elements, either on a single layer [8] or in a stacked configuration [32], in order to compensate the spatial phase delay at different frequencies by a fine adjustment of the elements' dimensions.

Similar to reflectarrays, a transmitarray antenna (or planar lens) is able to collimate the incident waves from a feed into a pencil beam on the output side of the lens. Transmitarrays can be implemented in simple way by using a phase-shifting element to connect the input and output array elements, such as variable-length delay lines [40], [41] or by using apertures to couple the receiving and transmitting arrays [42]. An alternative implementation of transmitarrays is based on a multilayer chain of inductive and capacitive surfaces separated by dielectric layers [43], [44], which can be designed by applying filter theory. When compared to reflectarrays, transmitarrays present two main advantages: they avoid blockage from the feed in centered-fed configurations and present less sensitivity to surface deformations. Moreover, transmitarrays share some significant features of reflectarrays, as their ability to discriminate in polarization [45] or frequency [46], and the possibility of implementing reconfiguration capabilities at the transmitarray element level [47]. In all previous cases, the unit cell should produce any required phase-shift and low reflection from the input side for all possible values of phase delay, thus leading to a more difficult design process.

1.2 State of the art on reflectarray antennas

The review of the state of the art on reflectarray antennas presented in this section includes the most relevant aspects related to this thesis, which are: reflectarrays with independent phase control in each polarization, multi-frequency reflectarrays, reflectarrays in dual reflector configurations (with special emphasis on dual reflectarray antennas), and multi-beam reflectarrays.

1.2.1 Reflectarrays with independent phase control in each polarization

Reflectarray antennas are able to generate independent beams in each polarization when appropriate reflectarray cells are employed. In these cases, a single dual-polarized feed can be used to illuminate the reflectarray, producing two different beams. This task cannot be accomplished with conventional reflectors, as the beams produced by a single feed would radiate in the same direction and with the same characteristics, so that spatially-separate feeds are required to generate beams with a different pointing direction. On the other hand, the capability of reflectarray antennas to introduce independent phase-shifts in each polarization can be used to design a reflectarray to fulfill independent requirements in each polarization (different missions) [8], [9].

Rectangular patches of variable size have been used for implementing different phase-shifts in two orthogonal linear polarizations (LP), either using a single layer [9] or three layers [8] of stacked patches, so that the resulting reflectarrays are able to generate independent beams in each polarization. The drawback of this concept is the narrowband operation of the reflectarray when a single layer of patches is used, while the stacked multi-layer structures provide larger bandwidth, but at the cost of a more complex and expensive fabrication.

In the recent years, several multi-resonant elements printed on a single layer have been proposed for the design of broadband reflectarray antennas, while keeping a simple manufacturing process [27]-[28], [48]-[54]. These elements make it possible to achieve a range of phase-shift variation greater than 360° , thanks to the use of multiple resonances. However, the reflectarrays based on such multi-resonant elements have not been reported to produce independent beams in each polarization, due to the difficulties encountered in achieving an independent control of the two polarizations in a wide frequency band (although some preliminary results have been reported at element level including reconfigurability [48]). The reflectarray elements based on concentric square and cross loops [27], [49] or the Phoenix cell [28] do provide the required phasing for dual-LP, but a 4-fold rotational symmetry of the elements is mostly assumed to maintain low levels of cross-polar radiation, forcing similar phase-shift distributions in both polarizations. When this rotational symmetry is broken, it is not possible to keep an independent phase control of both polarizations, since the variations in the elements' dimensions that control one polarization also modify the phase-shift introduced in the orthogonal polarization [50]. Moreover, the elements based on concentric rings [51] or

fractal patches [52] cannot be used to implement independent phasing in each polarization, as the phasing affects both polarizations simultaneously. Making appropriate gaps in the rings can be a solution for achieving a polarization selective response [53], [54], but in this case the element operates in single polarization.

The implementation of polarization discrimination with broadband reflectarray cells can be carried out by using multi-resonant elements based on coplanar parallel edge-coupled dipoles [55]. Two orthogonally-arranged sets of parallel dipoles, which can be printed on both sides of a dielectric layer [29], shifted in half-a-period on the same layer, or a combination of both [30], have been used for the design of broadband reflectarrays with independent phase control in each linear polarization. The orthogonal polarizations can be separately controlled by varying the lengths of the orthogonal dipoles. An example of a dipole-based cell configuration can be seen in Fig. 1-2(a).

Fig. 1-2 Comparison of reflectarray cells designed to provide independent control of each polarization: (a) in case of working in dual-LP [30], (b) in case of working in dual-CP [56], [57].

The design of reflectarrays to provide independent control of each circular polarization (CP) presents greater difficulty than in the case of linear polarization. A novel reflectarray cell (see Fig. 1-2(b)) has been recently proposed to introduce different phase-shifts in right-handed CP (RHCP) and left-handed CP (LHCP) [56], [57], though the cell exhibits a complex structure with thick metal cavities that lead to a bulky antenna design. Another approach consists on placing a circular-to-linear polarizer in cascade with a dual-LP reflectarray that provides independent control of each linear polarization [58]. In this configuration, the incident waves in orthogonal CP are converted into orthogonal LP by the polarizer, and then the LP are discriminated by the reflectarray and converted back to CP by the polarizer. A more elegant solution to achieve discrimination in dual-CP has been reported in [59], applying sequential

rotation of the reflectarray elements to generate two beams in orthogonal CP in two opposite directions. In contrast with the previous techniques, which require the use of complex and thick cell structures, this method can be implemented by a single layer of variably-rotated reflectarrays elements.

1.2.2 *Multi-frequency reflectarrays*

Several papers have been reported in the recent years for reflectarrays that operate simultaneously at two or more different frequencies. There are two basic strategies of design: either different resonant elements distributed on a single layer are used [53], [60]-[63], or a stacked multi-layer configuration is employed in which each reflectarray layer operates at a different frequency [11], [51], [64]-[68]. An example of each design approach can be seen in Fig. 1-3.

Concerning the first strategy, two single-layer reflectarrays have been designed to operate at 20 and 30 GHz in [60] and [61]. The reflectarrays consist of periodic elements based on a concentric dual split loop [60] and a split loop combined with a Malta cross [61], and they work in single-CP at each frequency. Two separate sets of coplanar open concentric loops have been accommodated in the same unit cell to design a single-layer reflectarray for operation at 12 and 14 GHz in orthogonal LP [53]. Despite its broadband performance, the presence of the two separate sets of concentric rings in each unit cell obliges to increase the size of the cell, which should be limited by the requirement of avoiding grating lobes [4]. A tri-band reflectarray has been shown in [62] to operate at 7, 8.5 and 32 GHz by combining three resonant elements in a single layer with different periods for the lattice. A dual polarized reflectarray providing independent phase control at four closely spaced frequencies (12, 13, 14 and 15.5 GHz) has been demonstrated in [63] by using multi-resonant cells; however, the symmetry of the printed elements does not allow to implement an independent phase control in each polarization.

On the other hand, a reflectarray based on two stacked layers with rectangular patches has been proposed for working at 6.5 and 10.6 GHz in dual-LP [11]. The patches on each layer can be adjusted to control the phase-shift at a different frequency, with the drawback of providing limited bandwidth. In [64], two layers of stacked patches were used to design a 20/30 GHz reflectarray with operation in orthogonal LP at each frequency, using the same concept than in [6] to increase the bandwidth. Two-

layer arrays of concentric split rings [65] and concentric rings with variable diameter [66] have been shown to operate at two closely spaced frequencies in X-band. A two-layer reflectarray with elements based on concentric rings and a circular patch inside a circular ring has been reported in [51] to operate at separate frequencies in Ku and Ka bands (14 and 35 GHz). Moreover, a dual-band transmit/receive reflectarray has been developed by placing a Ka-band reflectarray on top of an X-band reflectarray, and using a frequency selective surface (FSS) as separator [67], [68]. Despite the results of these works are satisfactory, the antenna presents a very thick structure with a high number of layers, and the symmetry of the elements makes them not suitable for independent phase control.

Fig. 1-3 Example of the two strategies for achieving multi-frequency operation: (a) different resonant elements distributed on a single layer [63], (b) stacked multi-layer configuration [51].

Therefore, single-layer multi-frequency reflectarrays present low profile and reduced manufacturing costs, but often at the price of not providing discrimination in polarization. By contrast, the increase in the number of layers, along with the utilization of complex cell structures, can be used to obtain an improved performance for the reflectarray cell, although it also results in higher manufacturing costs, weight and thickness of the final antenna.

1.2.3 Reflectarrays in dual reflector configurations

Dual reflector configurations in which the sub-, the main or both reflectors are reflectarrays have been proposed to obtain a better performance than conventional single and dual reflector antennas. They can be suitable for some applications requiring multiple beams, low cross-polarization or shaped beams [15]-[20], since the use of two

reflecting surfaces allows more degrees of freedom for the design. Also, controllable electronic devices based on MEMS or PIN diodes can be integrated in the elements of the sub-reflectorarray, in order to provide reconfiguration of the beam [18] [26].

The use of a reflectarray sub-reflector with a parabolic main reflector has been proposed for beam scanning applications [15] and the generation of contoured beams [69]. In these cases, the reflectarray sub-reflector provides additional functionalities to the dual reflector system, presenting a relatively simple design and low manufacturing costs. For example, beam scanning in a limited range (up to $\pm 5^\circ$) was demonstrated in [26] by adding a progressive phase on a reflectarray sub-reflector, for a radiometer system operating at 94 GHz.

Dual reflector configurations with a flat reflectarray as main reflector have been proposed in [16] for space applications. The flat reflectarray allows for more efficient packaging and deployment mechanisms than in the case of parabolic reflectors, thanks to the use of a planar surface [14]. On the other hand, the large size of the reflectarray results in narrowband operation, which is mainly caused by the differential spatial phase delay. The implementation of a parabolic main reflectarray [39] would increase the operating bandwidth, providing a smoother phase distribution on the reflectarray. The parabolic surface would focus the beam, while the printed elements would introduce small phase adjustments to shape or point the beam.

An efficient analysis tool has been recently developed and validated for dual-reflector antennas, where the sub-, the main or both reflectors can be substituted by planar or curved reflectarrays [70]. The technique is a modular approach that combines the Method of Moments (MoM) for the analysis of each reflectarray and Physical Optics (PO) when a main parabolic reflector is used. This method provides an accurate prediction of gain, co-polar and cross-polar radiation patterns, and it is numerically efficient, so it can be integrated in optimization routines for practical designs.

A dual reflectarray demonstrator has been designed to produce a collimated beam with very low cross-polarization (30dB of cross-polar discrimination) in a wide frequency band, covering transmit and receive frequencies in Ku-band (12-15 GHz) [71]. The demonstrator, with a 50-cm main reflectarray and 40-cm sub-reflectorarray, has been manufactured and tested (see Fig. 1-4(a)), showing a good agreement with the simulations and compliance with the design requirements.

Furthermore, a transportable dual reflectarray antenna has been shown in [18] to provide a collimated beam with electronic scanning capabilities for Ku-band bidirectional satellite links. The antenna consists of a passive sub-reflectarray and a reconfigurable main reflectarray, and is designed to operate in the 10.7-14.5 GHz band (reception and transmission in Ku-band). The reconfigurable reflectarray element is based on a PIN diode, and provides large bandwidth, dual linear polarization, and 1 bit of relative phase-shift. The main limitation of the prototype is its low radiation efficiency, which is mainly caused by the low performance of the PIN diodes and the 1-bit phase control.

Fig. 1-4 Pictures of manufactured dual reflectarray antennas: (a) compact-range prototype for broadband operation in Ku-band [71], (b) bifocal folded antenna to produce multiple beams at 76 GHz [17].

Another interesting concept is the folded reflectarray antenna, which presents the advantages of being low cost, low profile and low weight [72]. The folded antenna is based on a dual centered configuration, where a linearly-polarized feed illuminates a polarizing grid that is placed in front of a reflectarray. The reflectarray elements introduce the required phase-shift in the incident field (previously reflected by the grid) to collimate the beam, at the same time as providing a 90° twist of polarization, so that the reflected field passes through the grid without blockage. This concept can be used to design a dual reflectarray antenna when the polarizing grid incorporates an array of printed elements. A folded dual reflectarray antenna has been demonstrated in [17] to produce a shaped beam at 25 GHz for point-to-multipoint communications. Moreover, a bifocal folded dual reflectarray has been shown in [19] to generate multiple beams at 76 GHz for automotive radar systems (see Fig. 1-4(b)). The drawback of the folded configuration is that the antenna does not permit operation in dual polarization.

1.2.4 Multi-beam reflectarray antennas

Reflectarrays can be designed to generate multiple beams, either with a single feed [73]-[76], or with one or more feeds per beam [12], [13], [17], [77]-[78]. In the first case, the simplest approach consists on dividing the reflectarray surface into N sub-arrays, so that each sub-array will radiate a beam in a different direction [73]. Note that the power from the feed is divided between the N sub-arrays, each using only $1/N$ of the antenna aperture. A more efficient method, which uses the entire antenna aperture to produce N beams, is based on the superposition of the aperture fields associated to each beam to obtain the required phase-shift distribution on the reflectarray [4] (Ch. 7). A reflectarray with two beams separated 55° was designed by this technique to operate in dual-LP at 11.95 GHz, using two layers of stacked patches, and the fabricated breadboard was tested with satisfactory results [74]. The same concept was used to design a reflectarray with four symmetric beams at 32 GHz [75], and then, particle swarm optimization (PSO) procedures were carried out to produce four asymmetric beams [76]. In the latter case, the phase distribution on the reflectarray shows very sharp variations, which produces some beam distortions and very narrow bandwidth. Other alternatives for single-fed multi-beam reflectarrays consider the use of multi-resonant elements to produce different beams at closely spaced frequencies, as in [63].

On the other hand, reflectarrays can be used to generate multiple beams in a single feed per beam (SFPB) scheme, following a similar design process to the case of multi-fed reflectors. The SFPB configuration is commonly used in multi-beam applications requiring several simultaneous beams that must be generated independently, as in current satellite antennas in Ka-band. In this case, the reflectarray is designed to produce a collimated beam in a given direction for a feed located at the antenna focus. When the reflectarray is illuminated by a cluster of feeds placed near to the primary feed, the antenna produces additional beams in different directions. If the reflectarray is able to discriminate in frequency or polarization, each feed can generate more than one beam [13], [59]. A reflectarray breadboard providing three simultaneous shaped beams in Ka-band (25.5 GHz) in a SFPB basis has been reported in [77], with application to base station sectored antennas for point-to-multipoint communications. In [13], a study on the capability of reflectarrays to generate adjacent beams in Ka-band has been performed, using beam squint effect to point the beams in different directions at two closely spaced frequencies, with the drawback of very narrow bandwidth.

One of the main problems of the SFPB concept in multi-beam reflectarrays is that the beams suffer from aberration effects due to feed defocusing. To minimize the aberration, the positions of the feeds can be adjusted by using the same techniques than in multi-fed reflectors [79]. Moreover, several methods have been proposed in the last years to improve the multi-beam performance of reflectarrays. Some of these methods are based on the optimization of the phase distribution on the reflectarray, using PSO and genetic algorithms optimization (GAO) procedures [12], [78], which have the disadvantage that they may result in very fast variations between the phase-shift introduced by neighbour cells [78]. An interesting alternative to these approaches is the application of the classic bifocal technique [80], [81] to the design of dual reflectarray configurations, in order to obtain a better performance for the extreme beams. A multi-beam bifocal folded reflectarray antenna operating at 76.5 GHz has been demonstrated in [17] with quite good results, providing nine beams with up to $\pm 24.5^\circ$ scanning range.

1.3 State of the art on multi-beam satellite antennas in Ka-band

In the last decade, Ka-band has become a major alternative for satellite systems to satisfy the growing demand for capacity and provide new broadband services. The provision of such services is directly linked to the deployment of a high capacity network, which maximizes the users' throughput by fully exploiting the available bandwidth and power resources. For this purpose, current Ka-band satellites are required to provide a multi-spot coverage based on frequency and polarization reuse, both in transmission (Tx, 19.2-20.2 GHz) and reception (Rx, 29-30 GHz) bands [82].

The use of multiple spot beams instead of providing a large contoured beam (as in most broadcast Ku-band satellite applications), allows for a high degree of frequency and polarization reuse, leading to a significant increase of the system capacity and higher data rates for the users. A four colour reuse scheme with two frequencies and two polarizations is normally used (see Fig. 1-5), in which adjacent spots present different colours, meaning that they must be generated in a different frequency and/or polarization (note that spots of the same colour are spatially isolated from each other). On this basis, a new generation of high throughput satellites (HTS) has been developed to provide around 100 Gb/s for broadband services using Ka-band frequencies, and it is expected that the next HTS generation will reach several hundreds of Gb/s [83].

Fig. 1-5 Example of a four colour scenario for a pan-European multi-spot coverage [82].

The design of multi-beam antennas for Ka-band HTS systems [84] has become a major issue for satellite operators, as it must cope with some challenging requirements. In these cases, the antenna has to generate a large number of high-gain overlapping spot beams, typically between 50 and 100. The angular spacing between adjacent spots is considerably small (a typical value is 0.56°), in order to ensure a high value of gain at the end of the coverage (EOC gain) [85]. This performance cannot be achieved by a single conventional reflector, as the feed size that would provide the required beam spacing is smaller than the feed size that ensures an optimum illumination of the aperture (so that it may result in overlapping feeds). Therefore, either a larger F/D ratio or electrically smaller feeds have to be used, and both solutions lead to higher spillover losses [86]. This first option obliges to use more directive feeds, with the risk of feed overlap, while the second option forces to increase the antenna size, resulting in more weight on board the satellite. Moreover, the large number of spots required leads to further deterioration of the edge beams, which are generated by the feeds with the largest separations from the antenna focus.

In accordance with the above mentioned considerations, two main approaches are possible for the design of multi-beam antennas to provide the four colour coverage in Ka-band. The first approach considers the use of SFPB configurations [87]-[91], in which each beam is generated by a single feed. Despite its hardware simplicity, the drawback of this scheme is that several apertures are required to generate all the beams (typically four reflectors), although using a highly-oversized shaped reflector is also possible. On the other hand, multiple feed per beam (MFPB) systems [92]-[94] use small clusters of feeds to produce each spot. In this case, adjacent spots can be obtained

by means of feed arrays that share some of their elements, so that two apertures can provide multi-spot coverage for Tx and Rx in Ka-band. On the negative side, complex and expensive beamforming networks are required to feed the elements of these arrays. In both design approaches, SFPB and MFPB, the apertures are normally realized by reflector antennas, which provide high gain and bandwidth with limited manufacturing costs [84]; however, other promising alternatives are currently being investigated to produce multi-spot coverage by using a single aperture, such as aperiodic phased arrays [95], [96] and passive [97] and active [98] lenses.

1.3.1 SFPB antenna systems

To confront the stringent requirements of multi-spot applications in Ka-band, most of current HTS systems carry four reflector antennas on board the satellite, each reflector being responsible for generating all the beams in the same frequency and polarization (same color) in a SFPB basis [82], [87]. In early versions, this solution required up to eight reflectors: four reflectors for Tx, and four additional reflectors for Rx [88]. The modern configuration employs only four reflectors operating both in Tx and Rx, which are properly illuminated by Tx/Rx feed chains [82].

The four reflector configuration represents the current state of the art for multiple spot beam antennas in Ka-band. It eliminates the problem of overlapping feeds, since the adjacent beams in different frequencies and/or polarizations are generated by different reflectors, allowing more room to properly accommodate the feeds that illuminate each reflector. Low spillover can be ensured for a reasonable reflector size (between 2 and 2.6 m in diameter), providing a large number of beams with around 50 dBi maximum gain. The four reflectors are then pointed to produce slightly overlapping spots with 0.56° of spacing in a four colour scenario. To illustrate these concepts, Fig. 1-6 shows an example of a current HTS system operating in Ka-band (KA-SAT) [83].

From a mechanical point of view, the problem of this solution is the large number of antennas to be accommodated in the satellite (normally occupying two lateral faces of the satellite, as shown in Fig. 1-6(a)), as well as the large volume and weight of the antenna system. For these reasons, it would be preferable to reduce the number of apertures required to produce the multi-spot coverage, which would also result in lower costs for the antenna farm.

Fig. 1-6 Current state of the art for Ka-band HTS systems: (a) illustration of the KA-SAT with four reflectors [87], and (b) generation of the multi-spot coverage with a four colour scheme [82].

An interesting alternative is the use of three reflectors in a SFPB basis to provide a three colour scenario, as has been reported in [89]. However, when compared to the conventional four reflector solution of similar aperture size, this configuration presents higher spillover losses, which results in a slight degradation of the EOC gain.

Furthermore, the use of a single reflector antenna to produce all the beams (the four colours), leads to a highly oversized reflector, around 4.5 m in diameter, which provides the required 0.56° of spacing by using smaller (less directive) and non-overlapping feeds, at the same time as keeping low spillover [90], [91]. Note that the area of the oversized reflector is comparable to the total area of the previous four reflector solution, so it does not represent a significant improvement. Moreover, the oversized reflector has to be shaped in order to produce wider beams, so that a high EOC gain can be reached for all the beams (a spot diameter of 0.65° is typically considered).

1.3.2 *MFPB antenna systems*

The number of reflectors in the current state-of-the-art HTS antenna systems can be reduced by using a MFPB scheme, in which each beam is produced by a cluster of small feed-horns. Since the adjacent clusters of feeds share some of their elements, they are able to produce overlapping spot beams, as can be seen in Fig. 1-7).

Fig. 1-7 Feed system of a MFPB antenna with shared horns to provide overlapping spots [82].

In [92], two reflectors have been designed in a MFPB basis to provide the multi-spot coverage, with one reflector operating at Tx and the other at Rx. This configuration allows to separately optimize the size of each reflector according to the operating band. An alternative MFPB concept has been recently proposed based on a dual reflector configuration with a dichroic (frequency selective) sub-reflector, providing full Tx/Rx multi-spot coverage by the use of a single main aperture [93]. These solutions represent a significant reduction in weight and volume of the antenna farm, facilitating the accommodation of the antennas on board the satellite. However, the previous advantages are achieved at the expense of implementing complex and expensive beamforming networks [94]. Moreover, the performance of the MFPB solution is slightly inferior to that of the equivalent SFPB configuration with the same aperture size, and they are not suitable for providing non-regular spot lattices with different beam sizes [82]. For these reasons, the use of MFPB antennas for current multiple spot beam satellite applications in Ka-band is not widely spread, and these systems are still under development.

1.4 Motivation and goals of the thesis

As has been shown in the previous section, current multi-spot satellite antennas in Ka-band employ a large number of apertures (typically four) to generate a coverage made of overlapping spot beams in a SFPB basis, with a four colour frequency and polarization reuse scheme. This solution is simpler than the alternative MFPB systems in terms of hardware and ensures low spillover for a reasonable reflector size, but it presents some drawbacks derived from its mechanical complexity.

The ability of reflectarray antennas to provide independent control of the phase in different polarizations and/or frequencies can be of particular interest for this kind of multi-beam applications in Ka-band, leading to a potential reduction in the number of antennas required to provide the full coverage. This would result in significant savings in the cost, weight and volume of the antenna farm in current HTS systems in Ka-band. However, there are some issues that must be previously addressed.

First, the design of such an antenna requires to count on appropriate reflectarray cells that will provide an independent control of the phase at each frequency band (Tx and Rx) and/or polarization. As shown in section 1.2, the dual-frequency reflectarray cells that have been reported so far do not allow independent control of each polarization at both operating frequencies. Moreover, the performance of the reflectarray antenna should be optimized (in terms of gain, beamwidth, side-lobe levels, etc.) to generate a large number of beams in a SFPB basis, and the capability of the antenna to provide adjacent beams with a very small separation (around 0.56°) should be evaluated. With this considerations in mind, the following objectives are set:

1.4.1 Design of dual-frequency and dual-polarization reflectarrays

A dual-polarization reflectarray cell will be proposed for operation at two separate frequencies, providing independent control of the phase-shift introduced in each linear polarization and frequency. The proposed cell is based on previous developments regarding the design of broadband multi-resonant cells that use parallel edge coupled dipoles [29], [30]. A two layer configuration with stacked sets of parallel dipoles will be proposed, in which the calculation of the appropriate dipoles' lengths will be carried out separately for each layer, leading to a simpler and computationally faster design process. The cell dimensions will be adjusted to allow operation first at Tx frequencies in Ku and Ka bands (12 and 20 GHz), and then at Tx and Rx frequencies in Ka-band (20 and 30 GHz). Several dual-band reflectarrays will be designed using the previous cells and the aforementioned step-by-step design process. The reduced number of layers and the simplicity of the printed elements will allow an easier manufacturing and low profile for the resulting antenna.

1.4.2 Experimental validation of the proposed concept for dual-frequency reflectarray antennas

A dual-band reflectarray demonstrator of 25-cm diameter will be designed, manufactured and tested in order to validate both the proposed concept for the reflectarray cell and the design procedure of the antenna. A home-made full-wave electromagnetic code will be used for the analysis and design of the reflectarray. This tool is based on the application of the Method of Moments in the Spectral Domain (SD-MoM) and the local periodicity approach, and has been validated in previous works [30]. The measured radiation patterns of the fabricated demonstrator will be compared with those provided by the SD-MoM electromagnetic code, and the performance of the antenna will be evaluated.

1.4.3 Development of a bifocal design method for dual reflectarray configurations

Concerning the antenna performance for the generation of a large number of beams and its capability to provide closely spaced beams, the application of the classic bifocal technique [80], [81] to dual reflectarray configurations will be proposed in this thesis. The bifocal technique will provide better results for the extreme beams, at the same time as permitting a certain degree of control over the spacing obtained between adjacent beams. The previous works on this topic considered centered and rotationally-symmetrical antenna geometries [17], which have the advantage of a simplified design process. Offset configurations would be more suitable for the intended application (multi-beam satellite antennas), as they will minimize blockage from the sub-reflector and allow operation in dual polarization. Therefore, a bifocal design procedure will be developed for offset dual reflectarray configurations with a large main reflectarray. For this purpose, two different approaches will be considered: starting from an axially-symmetrical configuration which allows the rotation of a 2D design around the antenna symmetry axis [80], and implementing a 3D bifocal method that directly provides the required phases on both reflectarrays in the selected offset configuration [81].

1.4.4 Design of bifocal dual reflectarray configurations for multi-beam satellite antennas in Ka-band

The proposed bifocal method will be applied to different dual reflectarray configurations for the design of multi-beam satellite antennas in Ka-band. The bifocal

technique will allow to reduce the beam spacing with respect to the equivalent single-focus antenna for the same feed separation (so that such beam spacing could not be achieved by the monofocal antenna, as it would produce overlapping feeds). The performance of the bifocal technique will be evaluated for the provision of adjacent beams with 0.56° of spacing, which implies a high degree of beam spacing reduction (by a factor of 2). The use of the bifocal technique to obtain a better performance for the extreme beams providing the same beam spacing than the equivalent monofocal antenna will be also considered.

1.4.5 Experimental validation of the proposed bifocal design method

A bifocal dual reflectarray demonstrator will be designed, manufactured and tested in order to validate the proposed bifocal method and corroborate the advantages of the bifocal antenna over the equivalent single-focus design. With this aim, the bifocal antenna demonstrator will provide a reduced beam spacing and an improved performance for the extreme beams with respect to the equivalent monofocal antenna with the same configuration. A home-made analysis tool implementing the modular technique described in [70] will be used for the analysis of the dual reflectarray antenna. The accuracy of this tool has been validated in previous works, concerning the design, manufacturing and test of a dual reflectarray demonstrator with reduced cross-polar levels [71]. The measured radiation patterns of the bifocal antenna demonstrator will be compared with those computed by this analysis tool.

1.4.6 Application of the bifocal technique to dual transmitarray configurations

The developed bifocal procedure will be applied to the design of dual transmitarray antennas. The use of transmitarrays presents some advantages, as it eliminates blockage from the feeds and the first transmitting structure, allowing for the utilization of centered-fed configurations with rotational symmetry (thus resulting in a simplified design process). In addition, the transmitarray configuration is less sensitive to surface distortions than reflector or reflectarray antennas. The particularities of the bifocal dual transmitarray antenna with respect to the counterpart dual reflectarray will be studied. Finally, the design of bifocal dual transmitarray antennas for multi-beam applications in Ka-band will be addressed in a similar way to the case of dual reflectarrays.

1.5 Thesis organization

This thesis is divided into eight chapters. This chapter contains the current state of the art on reflectarray antennas and multi-beam satellite antennas in Ka-band, and introduces the motivation and objectives of the thesis. The rest of the chapters are organized as follows:

- Chapter 2 proposes a novel reflectarray element for independent operation at two separate frequencies, which provides independent phase control in each polarization at both frequencies. The reflectarray cell is based on a two-layer configuration with stacked sets of coupled parallel dipoles for controlling each linear polarization. The lengths of the dipoles can be separately adjusted for each layer, according to the required phase-shift distributions on the reflectarray at each design frequency. A 25 cm Ku/Ka-band reflectarray demonstrator has been designed, manufactured and tested to validate the proposed reflectarray cell and the antenna design procedure. Then, the element dimensions are modified to allow operation at Tx and Rx frequencies in Ka-band (20 and 30 GHz), and the element is used for the design two reflectarray antennas: a 1.6 m satellite antenna to produce adjacent beams in orthogonal polarizations both at Tx and Rx bands, and a 20 cm terminal antenna that generates a focused beam in dual polarization.
- Chapter 3 presents a bifocal design procedure for dual reflectarray antennas in offset configurations, starting from a rotationally-symmetrical geometry where the two reflectarrays are placed in parallel planes. Then, the reflectarrays can be tilted a certain angle to obtain smoother phase distributions. A preliminary study on the bifocal technique for the design of multi-beam satellite antennas in Ka-band is presented in this chapter, considering two main cases: reduction of beam spacing by a factor of 2 (in order to provide adjacent beams with 0.56° of spacing), and improvement of the multi-beam performance with respect to the equivalent monofocal antenna (in this case, without compressing the beams).
- Chapter 4 contains the application of the bifocal method described in Chapter 3 to the design of bifocal dual transmitarray antennas. In this case, the absence of blockage allows for the use of centered and rotationally symmetrical geometries, which results in a simplified design process. Different configurations regarding

the positioning of the feeds and the two transmitarrays have been studied, in order to facilitate the practical implementation of such an antenna. Then, the bifocal technique is used to design dual transmitarray configurations with a high degree of beam spacing compression.

- Chapter 5 describes a 3D bifocal design method for dual reflectarray antennas, which is a more general procedure than the one shown in Chapter 3. This technique can be applied to obtain the required bifocal phase distributions on each reflectarray, without any restrictions in the antenna geometry. The technique is validated for an axially-symmetrical configuration, and then used to design a multi-beam antenna in an offset compact-range configuration. The phase distributions and radiation patterns of the bifocal antenna are studied for three different design cases (no beam spacing compression, low beam spacing compression and high beam spacing compression), and compared with those provided by the equivalent monofocal antenna.
- Chapter 6 shows the design, manufacturing and testing of a bifocal dual reflectarray antenna demonstrator, in order to validate the bifocal technique presented in Chapter 4. The antenna presents a compact-range configuration, with a main reflectarray of dimensions 57 cm x 42 cm, and has been designed to operate in the 19.2-20.2 GHz band (transmission frequencies from a satellite in Ka-band). The demonstrator generates 10 beams alternating in dual-linear polarization, when the antenna is illuminated by an array of 5 contiguous feeds. The bifocal technique allows to reduce beam spacing by a factor of 1.2 with respect to the equivalent monofocal antenna, at the same time as improving the performance of the extreme beams.
- Chapter 7 presents the design of a bifocal dual reflectarray antenna with an elliptical main reflectarray to provide multi-spot coverage for transmission from a geostationary satellite operating in Ka-band. The bifocal technique is used to produce adjacent beams in the offset plane, combined with a monofocal design in the orthogonal plane. The interleaved beams for obtaining the full coverage are generated in the orthogonal polarization, using the capability of reflectarrays to discriminate in polarization. The multi-beam performance of the bifocal antenna is compared with that of an oversized shaped reflector shown in [90]. The

proposed concept can be also applied to the design of a Tx/Rx antenna if multi-frequency reflectarray cells are used to provide independent phasing in Tx and Rx frequencies.

- Chapter 8 summarizes the main conclusions and original contributions, presents the future research lines, and enumerates the publications and research projects related to the work carried out in this thesis.

Chapter 2

Design of reflectarrays for operation in dual polarization at two separate frequencies

2.1 Introduction

The reflectarray element that will be proposed in this chapter is based on a two layer configuration with two orthogonal sets of coupled parallel dipoles. This type of multi-resonant cell was intended to take advantage of the lateral coupling between the dipoles for reaching a single-layer broadband element [55], e. g., the unit cell based on three coplanar parallel dipoles provides a similar performance in terms of bandwidth and phase range to that of three stacked rectangular patches, with the additional benefits of being cheaper and easier to manufacture [99]. The idea of placing two orthogonally-arranged sets of coupled dipoles printed on the same layer, but shifted half-a-period, has been used for the design of broadband reflectarrays in Ku-band with independent phase control in each linear polarization [29], [30]. A new modification in this type of cell, which consists of adding a group of parallel dipoles on a second level of metallization, will be introduced to demonstrate the capability of reflectarrays to provide independent phase control in each linear polarization at two separate frequencies.

The proposed element can be used to design a reflectarray antenna which is able to fulfill independent requirements at each frequency and/or polarization. For example, in the case of a satellite transmit antenna, a contoured beam can be generated in Ku-band by optimizing the elements on the lower layer [7], [8], and at the same time, multiple spot beams can be obtained in Ka-band by properly designing the elements on the top layer [13], considering different feed chains for each mission. The implementation of

different missions on the same reflectarray antenna would result in significant savings in the costs, weight and volume of the antenna farm, especially in the case of telecommunication satellites that operate in Ku and Ka bands.

In the next sections, several dual polarized reflectarray antennas will be designed by using the proposed element to operate independently in the transmit frequencies (from the satellite) in Ku (11-13 GHz) and Ka (19-20 GHz) bands, and in the Ka-band transmit (19-20 GHz) and receive (29-30 GHz) frequencies. Furthermore, a 25-cm Ku/Ka-band reflectarray antenna demonstrator that generates a focused beam in dual-polarization using different feeds for each band has been designed, manufactured and tested, in order to validate the concept.

2.2 Dual polarized reflectarray to operate in Ku and Ka bands

A preliminary design of a dual polarized reflectarray antenna to operate at two separate frequencies in Ku and Ka bands (11.95 and 20 GHz) is presented in this section. First, the structure of the reflectarray cell and its operating principle are described in detail, and two variations of the same baseline configuration are proposed for the design of the cell. Then, the element dimensions are adjusted in both cell configurations to allow simultaneous operation in dual polarization at 11.95 and 20 GHz. Finally, a 33 cm circular reflectarray is designed by using the previous elements, and its performance is evaluated by means of full-wave electromagnetic simulation.

2.2.1 *Design of the reflectarray cell*

Two reflectarray cells based on a stacked two-layer structure and orthogonal sets of parallel dipoles are studied. The difference between the two alternative configurations for the cell relies on the number of coupled dipoles on the lower layer. These configurations have been selected with the aim of providing independent phases in dual polarization at two different design frequencies, by means of a simple reflectarray element (only two layers) that will lead to a low-cost antenna.

2.2.1.1 *Two layer reflectarray cell with six dipoles for each polarization*

The basic periodic element, shown in Fig. 2-1, consists of two orthogonal sets of dipoles. Each set comprises three coplanar parallel dipoles printed on a dielectric layer (named layer A), and three additional parallel dipoles, which are stacked above the

previous ones and are printed on the top of a second dielectric sheet (layer B). The lateral dipoles of each arrangement are symmetrical with respect to the central dipole, in order to keep low levels of cross-polarization [30]. The toll to be paid with this constraint is that some degrees of freedom are lost when adjusting the phase for each frequency and polarization.

Fig. 2-1 View of the reflectarray periodic structure, including four unit-cells for X polarization and one unit-cell for Y polarization.

Phase shifting can be implemented by adjusting the lengths of all dipoles in the way that follows. The phase introduced in X polarization will be controlled by the lengths of the dipoles in the direction of the x -axis, whereas the phase introduced in Y polarization will be achieved with the appropriate lengths of the dipoles in the direction of the y -axis. Similarly, the dipoles on the lower layer will adjust the phases in Ku-band, while the dipoles on the higher layer will provide the phasing in Ka-band.

The dual-frequency operating principle works because of the difference in the dimensions of the elements between the two layers. Since the dipoles on the top layer will be shorter than those on the bottom layer, the phase response of the unit-cell at the lower design frequency (in this case, 11.95 GHz) will not be affected by the length of the upper dipoles. Although it is not reciprocal, because the phase response at the higher design frequency (20 GHz) will slightly depend on the length of the lower dipoles, in a certain way the lower dipoles will behave as a ground plane for higher layer elements. In practice, the longer dipoles on the bottom layer will add a reactance (phase-shift) at the higher frequency that must be compensated in the design process.

A period of 10 mm, which is $2\cdot\lambda/3$ at 20 GHz, has been fixed for the horizontal (P_X) and vertical (P_Y) dimensions of the cell. This period has been chosen to provide both, enough room for the dipoles and enough range of phase-shift in the lower band, and at the same time, to avoid grating lobes as much as possible in the higher band, considering a maximum angle $\theta = 30^\circ$ (angle of incidence from the feed on the reflectarray elements or angle of the radiated beam), according to the expression provided in [4] (p. 84):

$$P \leq \frac{\lambda}{1 + \sin \theta} \quad (2-1)$$

Commercially available materials have been used to implement both dielectric layers: the lower layer is an AD255C sheet, while the upper substrate is a Diclad 880B sheet. The electrical properties and thickness of the two dielectric layers are: relative permittivity $\epsilon_{rA} = 2.55$, $\epsilon_{rB} = 2.17$; loss tangent $\tan\delta_A = \tan\delta_B = 0.0009$; and thickness $h_A = 2.363$ mm, $h_B = 1.524$ mm.

A parametric study has been carried out by changing the different geometrical parameters (width, separation and ratio of lengths for the dipoles) and finally, the following parameters have been chosen to provide a smooth variation in phase response and cover a wide range of phase ($\geq 360^\circ$) in both frequency bands: dipole width $w = 0.5$ mm; separations between laterally coupled dipoles $S_{XA} = S_{YA} = 1.5$ mm, $S_{XB} = S_{YB} = 1.5$ mm; and relative sizes of the lateral dipoles $l_{A1} = 0.63\cdot l_{A2}$, $l_{A3} = 0.63\cdot l_{A4}$ (where l_{A2} and l_{A4} correspond to the lengths of the central dipoles in the lower layer), $l_{B1} = 0.78\cdot l_{B2}$, $l_{B3} = 0.7\cdot l_{B4}$ (where l_{B2} and l_{B4} correspond to the lengths of the central dipoles in the upper layer).

2.2.1.2 Two layer reflectarray cell with eight dipoles for each polarization

An alternative reflectarray cell has been proposed (see Fig. 2-2) following the same operating principle than the element presented in the previous section. It comprises two sets of five coplanar parallel dipoles printed on the lower dielectric layer (layer A), and three parallel dipoles stacked above the first sets and printed on the top of a second dielectric sheet (layer B). The reason for introducing this modification on the cell (two additional dipoles on the lower layer for each linear polarization) is to obtain greater stability in the phase response at the higher frequency and to provide an additional length variable for implementing the design process in each polarization.

The cell period is chosen again as $P_X = P_Y = 10$ mm. After a parametric study, the following parameters have been selected to obtain a smooth phase variation and a wide range of phase-shift in both frequency bands: dipole width $w = 0.5$ mm; edge-to-edge separations between dipoles $S_{XA} = S_{YA} = 0.5$ mm, $S_{XB} = S_{YB} = 1.5$ mm; and relative sizes of lateral dipoles $l_{A1} = 0.59 \cdot l_{A3}$, $l_{A2} = 0.75 \cdot l_{A3}$, $l_{A4} = 0.59 \cdot l_{A6}$, $l_{A5} = 0.75 \cdot l_{A6}$ (where l_{A3} and l_{A6} correspond now to the central dipoles' lengths in the lower layer), $l_{B1} = 0.8 \cdot l_{B2}$, $l_{B3} = 0.8 \cdot l_{B4}$.

Fig. 2-2 View of the reflectarray periodic structure, including four unit-cells for Horizontal polarization and one unit-cell for Vertical polarization.

2.2.1.3 Comparison of phase and amplitude curves

The simulated phase and amplitude curves of the co-polar reflection coefficient are shown in Fig. 2-3 for both proposed reflectarray cells, considering X-polarization and oblique incidence, $\theta_i = 20^\circ$ and $\varphi_i = 0^\circ$ (which is related to the elements located on the central part of the reflectarray that will be designed in the next section). Note that the response in both polarizations is identical for normal incidence. In this case ($\theta_i = 20^\circ$), a similar response is obtained for Y-polarization, although the results are not shown here.

The results in Fig. 2-3 are presented as a function of the lengths of the central dipoles in the direction of x -axis: the lower central dipole in Fig. 2-3(a), and the upper central dipole in Fig. 2-3(b). Note that the lengths of the lateral dipoles are also varied, while keeping the relations given above between their lengths and those of the central dipoles. As can be seen, the phase and amplitude curves are similar for both proposed reflectarray cells, although the second configuration (8+8 dipoles) provides a wider phase variation range in the lower frequency band. Smooth variations in phase are

achieved in both configurations, especially at Ku-band, and the encompassed phase range is wide enough (more than 360°) to perform the dual-band design process. Moreover, the orthogonal dipoles that control each linear polarization are practically uncoupled, as shown in [30].

Fig. 2-3 Phase and amplitude of the co-polar reflection coefficient for X-polarization: (a) at 11.95 GHz, (b) at 20 GHz.

The phase-shift introduced at each frequency band can be mostly controlled by adjusting the lengths of the dipoles in each layer, as shown in Fig. 2-4 for the (6+6) dipole element (note that a similar behaviour is obtained for the second cell configuration, although the graphics are not included here). This figure presents the variation in the phase of the co-polar reflection coefficient with respect to the lengths l_{A2} and l_{B2} (lengths of the lower and upper central dipoles in the direction of x -axis) at 11.95 and 20 GHz, considering X-polarization and oblique incidence ($\theta_i = 20^\circ$, $\varphi_i = 0^\circ$). The phase-shift introduced at 11.95 GHz can be adjusted by the lengths of the dipoles on the bottom layer, as shown in Fig. 2-4(a). However, the dipoles on layer A have a certain influence on the phase response at 20GHz, see Fig. 2-4(b). This effect must be corrected in the design process, obtaining the appropriate lengths for the dipoles on layer B once the dimensions of the lower layer elements have been fixed.

Fig. 2-4 Phase of the cell reflection coefficient (in degrees) with respect to the lengths of the central dipoles in both layers, considering X-polarization and oblique incidence ($\theta = 20^\circ$): (a) at 11.95 GHz and (b) at 20 GHz.

Finally, it should be remarked that the curves shown in Fig. 2-3 are only used for the periodic element characterization. During the subsequent design process, the real angles of incidence on each reflectarray element (θ_i , φ_i) will be taken into account to calculate the phase of the reflection coefficient and adjust the lengths of the dipoles. This variation on the angles may slightly modify the trace of the phase curves in Ka-band, while having almost no effect in Ku-band. A more detailed characterization of the element (operating at slightly different frequencies) will be provided in section 2.3.1, regarding the design of a limited size dual-band reflectarray demonstrator.

2.2.2 Design of a Ku/Ka-band dual polarized reflectarray antenna

A circular reflectarray with 33 cm diameter has been designed to operate in dual-linear polarization (dual-LP) at 11.95 and 20 GHz. The reflectarray consists of 861 elements placed in a 33 x 33 circular grid, with cell size 10 x 10 mm. The antenna is illuminated by an ideal feed-horn operating in dual-polarization and dual-frequency, placed with a certain offset in x -axis from the broadside radiation position, according to the absolute reference coordinate system shown in Fig. 2-5 (where the origin of the coordinate system is located at the geometrical center of the reflectarray). The coordinates of the horn phase center are $(x_F, y_F, z_F) = (-192, 0, 627.5)$ mm.

Fig. 2-5 Reflectarray antenna, with feed-horn position and reference coordinate systems.

The electromagnetic field radiated by the feed-horn has been modeled using a simple $\cos^q(\theta)$ distribution (more details can be found in [100], pp. 28-29), with a q -factor equal to 26 for Ku-band and 50 for Ka-band. Under these conditions, the illumination levels on the reflectarray edges are close to -10.2 dB at 11.95 GHz and -11.8 dB at 20 GHz, looking for a compromise between the illumination and spillover efficiencies which maximizes the antenna gain.

The reflectarray has been designed to produce two focused beams in the directions $\theta_{bX} = 13^\circ$, $\phi_{bX} = 0^\circ$ for X-polarization and $\theta_{bY} = 20^\circ$, $\phi_{bY} = 0^\circ$ for Y-polarization both at Ku and Ka bands, using the two types of cell described in section 2.2.1. The required phase-shift to be introduced by each reflectarray element for a single polarization will be achieved by adjusting the dimensions of the two stacked sets of parallel dipoles placed on each layer, according to the phase-shift distributions shown in Fig. 2-6. These

distributions have been obtained applying the expression provided in [4] (p.34), which relates the required phase for the cell reflection coefficient (φ_R) with the phase of the incident field (radiated by the feed) on that cell and the specified direction for the radiated beam (θ_b, φ_b):

$$\varphi_R(x_i, y_i) = k_0(d_i - \sin \theta_b (x_i \cos \varphi_b + y_i \sin \varphi_b)) \quad (2-2)$$

where k_0 is the propagation constant in vacuum, d_i is the distance from the phase center of the feed to the cell, and (x_i, y_i) are the coordinates of the i -element.

Fig. 2-6 Phase-shift distributions (in degrees) to be introduced by the reflectarray in: (a) X-polarization at 11.95 GHz, (b) Y-polarization at 11.95 GHz, (c) X-polarization at 20 GHz, (d) Y-polarization at 20 GHz.

The dipoles with the largest lengths will be located at those elements with phase-shift values close to -360° , according to the phase curves showed in Fig. 2-3. The majority of these elements are placed near to the geometrical center of the antenna, which will be the most illuminated part, with incidence angles between $\theta_i = 10^\circ$ and $\theta_i = 20^\circ$ (see Fig. 2-7). An appropriate election of phase constants for the simulations will be performed in

order to minimize the errors in this zone of the reflectarray, which will be decisive in shaping the antenna radiation pattern.

Fig. 2-7 Angles of incidence (in degrees) from the feed on each reflectarray cell: (a) theta, (b) phi.

2.2.2.1 Two-step design procedure for dual-band reflectarrays

As stated earlier in this chapter, the length of the dipoles in layer A will be larger than the length of dipoles in layer B, so that upper dipoles will not disturb the reflectarray phase response at 11.95 GHz, and lower dipoles will behave in a certain way as a ground plane at 20 GHz for higher layer elements. Due to this operation mode, the dipole lengths needed to match the required phase distributions on the reflectarray can be calculated separately for each frequency band, considering only the dipoles in one of the two layers.

First, the dipoles on the bottom layer are adjusted to provide the required phase-shift at 11.95 GHz. Since the dimensions of the bottom layer influence the phase response at 20 GHz, the lengths of the dipoles on the top layer are adjusted to produce the required phase at 20 GHz at each reflectarray cell, taking into account the real dimensions of the lower dipoles previously calculated at 11.95 GHz. In this way, the effect of the bottom dipoles on the phase response at 20 GHz (see Fig. 2-4(b)) is taken into account in the design of the antenna. Moreover, the dipoles in the direction of the x and y axes can be independently adjusted due to the uncoupling of the phases for the two orthogonal polarizations, as was demonstrated in [30]. All these factors allow for performing an easier and computationally faster design process.

To obtain the appropriate lengths for each arrangement of coplanar parallel dipoles, it is employed a zero-finding procedure which iteratively calls a home-made software analysis routine that applies the Method of Moments in the Spectral Domain (SD-MoM) [29]. First, the lengths of the dipoles are gradually increased (keeping the relations given above between the side and central dipoles' lengths), until the phase-shift exceeds the required value. This provides a short interval which contains the target length for the central dipole. Then, the accuracy of this approximation is improved by applying the Newton-Raphson method, which results in fast convergence to the target value after only a few iterations of the algorithm. For this purpose, it is important that the reflectarray elements present a smooth variation in the phase response at both frequency bands (see Fig. 2-3).

The SD-MoM analysis tool used for the design of the dipoles has been developed at the Applied Electromagnetics Group of Universidad Politécnica de Madrid (UPM), in collaboration with the Faculty of Physics of Universidad de Sevilla. This full-wave electromagnetic code is based on the use of multilayered Green's functions and assumes an infinite periodic array model for the analysis of each cell. The analysis technique is an extension of the one presented in [101] for a reflectarray cell made of three parallel dipoles. It computes the reflection matrix of each cell, Γ , formed by the co-polar (Γ_{XX} and Γ_{YY}) and cross-polar (Γ_{XY} and Γ_{YX}) reflection coefficients that relate the tangential components of the incident electric field in the direction of dipoles with the tangential components of the reflected field, as shown in eq. (2-3). For this purpose, the real angles of incidence (θ_i, ϕ_i) on each reflectarray cell are taken into account.

$$\begin{pmatrix} E_X^{Ref} \\ E_Y^{Ref} \end{pmatrix} = \begin{pmatrix} \Gamma_{XX} & \Gamma_{YX} \\ \Gamma_{XY} & \Gamma_{YY} \end{pmatrix} \cdot \begin{pmatrix} E_X^{Inc} \\ E_Y^{Inc} \end{pmatrix} \quad (2-3)$$

The SD-MoM code provides a fast and accurate calculation tool; further details about the validation and performance of this code can be found in [30] and [102]. This includes a comparison between the simulated and measured radiation patterns of a 40-cm reflectarray demonstrator with cells also made of coupled parallel dipoles, and the significant savings in computational times that this numerical code offers over other commercial tools, such as CST [103]. For example, the simulation time of a 24-cm sided reflectarray at 11.95 GHz has been reduced from 8 hours (using CST without local periodicity) to only half a minute (SD-MoM plus local periodicity) [102].

Once the design of the dipoles is completed, independently for each frequency, a further optimization is run, in which the dimensions of all dipoles are optimized element-by-element to simultaneously match the phases at the central and extreme frequencies both in Ku and Ka bands. In this case, a 2 GHz bandwidth is considered in Ku-band (10.95-12.95 GHz), while 1 GHz bandwidth is enforced in Ka-band (19.5-20.5 GHz). The optimization involves a fine adjustment of the dipoles' lengths, following a procedure very similar to that shown in [32]. As in the design process, the optimization can be separately run for the dipoles in the direction of the x and y axes. In the end, the optimization provides an improved performance for the antenna at the extreme frequencies of the prescribed bands, at the cost of a slightly worse performance at the central frequency of each band.

2.2.3 *Results of the simulations*

The simulated radiation patterns in gain (dBi) of the 33-cm reflectarray antenna are shown in Fig. 2-8 for the design performed with the (6+6) dipole element, and in Fig. 2-9 for the design with the (8+8) dipole element, in both cases before conducting the multi-frequency optimization. The results are presented at the central frequency of each band (11.95 and 20 GHz), in the elevation and azimuth orthogonal planes. Note that the elevation plane coincides with the xz -plane for both polarizations (the beams are radiated at $\varphi = 0^\circ$), while the azimuth planes are orthogonal to the xz and form 13° and 20° with respect to z -axis for X and Y polarizations, respectively (although the two cuts are superimposed in the same figure). The patterns include both co-polar and cross-polar components of the radiated field for each polarization. The electric field on the aperture of the feed-horn has been used to compute the incident field on the reflectarray. Then, the radiation patterns have been obtained from the tangential reflected field at each reflectarray cell, using the aforementioned SD-MoM code.

As can be seen in Fig. 2-8 and Fig. 2-9, the patterns at the lower frequency are very similar for both designs. The beam is focused in the direction $\theta_{bX} = 13^\circ$, $\varphi_{bX} = 0^\circ$ for X-polarization and $\theta_{bY} = 20^\circ$, $\varphi_{bY} = 0^\circ$ for Y-polarization, as it was intended. A gain around 30.7 dBi is reached at 11.95 GHz for the beam in X polarization, while 30.1 dBi is achieved for the beam in Y-polarization. The side-lobe levels (SLL) are close to -21 dB with respect to the co-polar maximum for both polarizations, and the cross-polar discrimination (XPD) is around 34 dB, measured within a 3-dB beamwidth.

(a)

(b)

(c)

(d)

Fig. 2-8 Simulated radiation patterns of the (6+6) dipole antenna: (a) xz -plane at 11.95 GHz, (b) superposition of azimuth cuts at 11.95 GHz, (c) xz -plane at 20 GHz, and (d) superposition of azimuth cuts at 20 GHz.

Fig. 2-9 Simulated radiation patterns of the (8+8) dipole antenna: (a) xz -plane at 11.95 GHz, (b) superposition of cuts in the azimuth plane at 11.95 GHz, (c) xz -plane at 20 GHz, (d) superposition of cuts in the azimuth plane at 20 GHz.

On the other hand, the use of the second cell structure, with eight dipoles for each polarization, results in a more accurate beam shaping at 20 GHz (see Fig. 2-9(c) and Fig. 2-9(d)), with an important reduction in the SLL (from -20 to -23 dB) with respect to the design performed with the (6+6) dipole element. Moreover, the (8+8) configuration contributes to leveling the results obtained for the beams in X and Y polarizations. In the patterns shown in Fig. 2-8 for the (6+6) dipole antenna, the gain of the beam produced at 20 GHz in X-polarization is 34.1 dBi, which is around 1.6 dB higher than the gain of the beam in Y-polarization (32.5 dBi). However, in the case of the (8+8) dipole antenna, this difference is reduced to 0.6 dB, since 34.1 and 33.5 dBi gain is attained for the beams in X and Y polarizations, respectively. Finally, the (8+8) dipole antenna presents a slight improvement in XPD, which is around 22.4 dB instead of 19.2 dB, as in the case of the (6+6) dipole antenna.

It has been checked that the radiation patterns at the extreme frequencies of each band present better results for the beams in X-polarization than for those in Y-polarization, which show a further deterioration with respect to the patterns at the central design frequencies, particularly in the case of Ka-band (see Fig. 2-10 to Fig. 2-15 for the (8+8) dipole antenna). After carrying out the multi-frequency optimization, the antenna performance is improved for both polarizations at the extreme frequencies in Ku and Ka bands (higher gain, lower SLL, better beam shaping, etc.), at the expense of obtaining a slightly worse performance at the central frequencies. The results of the optimization have been included in Fig. 2-10 to Fig. 2-15 for the (8+8) dipole reflectarray, since the design made with the (6+6) configuration showed a worse response against frequency shifts respect to the central frequencies.

In the case of X-polarization, the available bandwidth in Ku-band is around 2 GHz centered at 11.95 GHz (16%), with a gain variation of 1 dB, while 1 GHz bandwidth is reached at 20 GHz (5%) with the same gain variation. The optimization provides improvements of up to 2 dB in the gain and up to 7.5 dB in the SLL for the patterns at the extreme frequencies of both operating bands with respect to the results obtained before the optimization. Regarding the beams in Y-polarization, the antenna presents reasonably good results in Ku-band, where 16% bandwidth (2 GHz centered at 11.95 GHz) is achieved with a 1.3 dB gain variation and similar enhancements in the gain and SLL of the patterns than in the case of X-polarization. However, the improvement provided by the optimization is not sufficient to reach an acceptable performance at the

extreme frequencies in Ka-band (19.5 and 20 GHz), where the patterns present around 2.8 dB lower gain and 6 dB higher SLL than those obtained at 20 GHz.

Fig. 2-10 Simulated radiation patterns of the (8+8) dipole antenna at 10.95 GHz before and after multi-frequency optimization: (a) in the xz -plane, and (b) superposition of cuts in the azimuth plane.

Fig. 2-11 Simulated radiation patterns for the (8+8) dipole antenna at 11.95 GHz before and after multi-frequency optimization: (a) in the xz -plane, and (b) superposition of cuts in the azimuth plane.

Fig. 2-12 Simulated radiation patterns for the (8+8) dipole antenna at 12.95 GHz before and after multi-frequency optimization: (a) in the xz -plane, and (b) superposition of cuts in the azimuth plane.

Fig. 2-13 Simulated radiation patterns of the (8+8) dipole antenna at 19.5 GHz before and after multi-frequency optimization: (a) in the xz -plane, and (b) superposition of cuts in the azimuth plane.

Fig. 2-14 Simulated radiation patterns of the (8+8) dipole antenna at 20 GHz before and after multi-frequency optimization: (a) in the xz -plane, and (b) superposition of cuts in the azimuth plane.

Fig. 2-15 Simulated radiation patterns of the (8+8) dipole antenna at 20.5 GHz before and after multi-frequency optimization: (a) in the xz -plane, and (b) superposition of cuts in the azimuth plane.

There are several reasons for the degradation of the antenna performance at Ka-band frequencies: the large period of the cells, which is required to provide enough range of length variation for the lower dipoles (allowing 360° of phase-shift variation margin in Ku-band), the presence of phase errors in the design of the upper dipoles, and the appearance of grating lobes due to the excitation of higher order Floquet modes for certain angles of incidence (associated to a few cells near the reflectarray edge). This degradation of the radiation patterns in Ka-band will be studied in more detail in section 2.3.4, concerning the measurement of a Ku/Ka-band reflectarray demonstrator.

Finally, the antenna radiation efficiency is estimated at 68% in Ku-band and 54% in Ka-band, which are typical values for conventional parabolic reflectors. Note that the conducted simulations take into account most of the losses, such as spillover, illumination and dielectric losses. The previous efficiencies have been calculated as the ratio between the simulated gain and the maximum directivity at each frequency. The latter magnitude can be approximated using eq. (2-4), which relates the gain of the antenna with the size of the aperture (S_{ap}), the wavelength at the operation frequency and the radiation efficiency (ϵ_{rad}). According to eq. (2-4), the maximum directivity is obtained for $\epsilon_{rad} = 1$ (100% efficiency).

$$G = \frac{4\pi}{\lambda^2} \cdot S_{ap} \cdot \epsilon_{rad} \quad (2-4)$$

2.2.4 Conclusions

A 33-cm diameter circular reflectarray antenna has been designed to operate in Ku and Ka bands (11.95 and 20 GHz) with independent beams in each linear polarization. The proposed cell structure for the reflectarray element consists of two stacked layers, each of them comprising two orthogonal sets of coplanar parallel dipoles. The element enables independent phasing in each linear polarization at both design frequencies. Two variations of this structure have been studied, providing four and five degrees of freedom for each polarization.

The lengths of the dipoles have been optimized to meet the required phase-shift distributions at each frequency band and polarization. For this purpose, a two-step design procedure has been implemented, in which the appropriate lengths of the dipoles are computed first for the elements on the lower layer (according to the phases at 11.95

GHz), and then for those at the upper layer (according to the phases at 20 GHz). This staged procedure allows for a simpler and computationally faster design process. The conducted simulations show a good behaviour of the antenna in both frequency bands, with gain values greater than 30 dBi and SLL close to -20 dB. A better performance at the higher frequency band is obtained for the reflectarray designed using the second cell structure, with eight dipoles for controlling the phase in each linear polarization.

These results show the potential of reflectarray antennas for working at different bands, with independent beam shaping in each polarization and band. Furthermore, the ease of integration and the reduction in cost, with only two layers of printed elements, are clearly appreciated. This may be a key factor in communication satellites that operate in Ku and Ka bands, where volume and weight are determining issues. The next step will involve the design, manufacturing and measurement of a reflectarray prototype to corroborate the results of these simulations and validate the design concept.

2.3 Design, manufacturing and test of a dual polarized reflectarray demonstrator to operate in Ku and Ka bands

Telecommunication satellites require an increasingly larger number of antennas to provide fixed services in Ku-band and broadband access in Ka-band [86], [87]. A major trend is to accommodate multiple payloads for different functions in the satellite. Dichroic sub-reflectors have been proposed to enable the reuse of the same main reflector for Ku- and Ka-band missions [104], or for both Tx and Rx operation in Ka-band [84], [93]. This will provide a significant saving of space in the satellite, at the same time as allowing to separate the feed chains for each frequency band. However, the use of such frequency selective surfaces (FSS) in telecommunication antennas is not straightforward, because the reflector has to be optimized in a different way for each mission. For example, whereas the reflector is typically shaped in Ku-band to provide a contoured coverage, it should provide multiple spots in Ka-band, and in both cases stringent requirements in gain, cross-polar and co-polar isolation are needed [87].

The reflectarray element proposed in the previous section can be applied to design a satellite transmit antenna, which is optimized to fulfill the requirements for two simultaneous missions in Ku and Ka bands, considering different feed chains for each mission. For example, a contoured beam can be generated in Ku-band by optimizing the

dipole lengths in the lower layer as in [30], and at the same time, multiple spots can be obtained in Ka-band by optimizing the dipoles on the upper layer [13], having more room to properly accommodate the feed chains of each mission. This kind of space applications will require a large reflectarray antenna, around 1.5 m in diameter. As a proof of concept, a limited sized demonstrator that generates a focused beam in dual-polarization using different feeds for Ku and Ka bands has been designed, manufactured and tested.

The proposed reflectarray is able to operate independently in the transmit frequencies in Ku-band (11-13 GHz) and Ka-band (19-20 GHz), but similar reflectarray cells can be designed to operate in the receive frequencies (13.75-14.80 GHz in Ku-band and 29-30 GHz in Ka-band). This technology permits an independent optimization of the radiation patterns and position of feed-chains for spacecraft antennas in Ku and Ka bands. Although it obliges to separate transmit and receive antennas, the reuse of the same aperture for Ku and Ka missions would lead to significant savings in the costs, weight and volume of the antenna farm in telecommunication satellites that operate at those frequency bands.

2.3.1 *Design of the reflectarray cell*

The unit-cell used to provide the phasing in the reflectarray presents the same structure than the one described in section 2.2.1.2 (see Fig. 2-2). It is composed of two orthogonal arrangements of eight parallel dipoles, which are distributed on a two-layer configuration: there are five parallel dipoles on the lower layer (layer A), and three stacked dipoles on the higher layer (layer B). This configuration has been selected since it provides a better performance at the higher frequencies than the alternative cell with six dipoles for each polarization (see section 2.2.3). The geometrical parameters of the cell (period, dipole width, separation between adjacent dipoles, ratio of lengths for the side and central dipoles) are the same than were indicated in section 2.2.1.2. Moreover, the two dielectric layers are implemented by the same materials, AD255C (layer A) and Diclad 880B (layer B).

In this case, the central design frequencies in Ku- and Ka-band will be 12 and 19.5 GHz, respectively. Note that the higher frequency has been slightly reduced with respect to the previous design (20 GHz), in order to obtain a better performance of the antenna in Ka-band. Fig. 2-16 shows the results for the module and phase of Γ_{xx} (co-polar

reflection coefficient for X-polarization) with respect to the lengths of the dipoles in the direction of x -axis, under oblique incidence conditions ($\theta_i = 20^\circ$, $\phi_i = 0^\circ$). Similar curves are achieved for the orthogonal polarization, considering the lengths of the dipoles in the direction of y -axis. As can be seen, the phase shows a fairly smooth variation (almost linear at the lower frequencies) in a 400° interval, which comprises the required 360° margin for the design of the reflectarray.

Fig. 2-16 Magnitude and phase of the cell reflection coefficient, considering X-polarization and $\theta_i = 20^\circ$ incidence: (a) at Ku-band frequencies, (b) at Ka-band frequencies.

The element response has been studied for different angles of incidence (see Fig. 2-17), showing slight variations of phases in Ka-band and almost no effect on the phases in Ku-band. To avoid any possible errors in Ka-band, the calculation of the reflection coefficients will take into account the real angles of incidence from the feed in the design and analysis of the antenna.

Fig. 2-17 Phase of the cell reflection coefficient for X-polarization under different angles of incidence: (a) at 12 GHz, (b) at 19.5 GHz.

Moreover, Fig. 2-18 shows the variation in phase of Γ_{xx} and Γ_{yy} co-polar reflection coefficients with respect to the lengths of upper and lower dipoles in the corresponding directions, at 12 and 19.5 GHz for $(\theta_i = 20^\circ, \phi_i = 0^\circ)$ angles of incidence. The element phase response at 12 GHz can be completely controlled by the lengths of the lower dipoles, which means that the design of the bottom layer at 12 GHz can be carried out without considering the dipoles on the higher layer. However, this is not exactly true for the design of the higher layer at 19.5 GHz, since there is certain dependence between the lengths of upper dipoles and the lengths previously calculated for the lower ones, which may slightly affect the phase response of the element. Despite this fact, once the lower layer dimensions have been fixed, there is still enough phase range to implement the required phases at 19.5 GHz with the lengths of upper dipoles.

Fig. 2-18 Phase (in degrees) of the cell reflection coefficient with respect to the lengths of the dipoles in both layers, at 12 GHz (a) for X-polarization and (b) for Y-polarization; and at 19.5 GHz (c) for X-polarization and (d) for Y-polarization.

Therefore, almost independent design processes can be performed for each layer, as was indicated in section 2.2.2. For example, the dipoles on the lower layer can be designed to generate a contoured beam in Ku-band by fitting the appropriate phase distributions in that band. Then, keeping fixed the dipoles in the lower layer, the dipoles on the upper layer can be adjusted to match different phase-shifts, for example to fulfill the requirements for multi-beam operation in Ka-band. The implementation of this step-by-step method simplifies the design of the antenna, and causes an important reduction in computational times. Possible phase errors can be corrected by means of an additional optimization that will be run after the design process.

2.3.2 *Design of the demonstrator*

A Ku/Ka-band reflectarray antenna demonstrator, comprising 625 elements disposed in a 25 x 25 grid (250 mm sided antenna), has been designed, manufactured and tested, in order to show the capability of reflectarrays to provide independent phase control at

two frequencies and two polarizations. The reflectarray will produce a collimated beam in dual polarization at 12 and 19.5 GHz. The beam will radiate in the direction $\theta_b = 20^\circ$, $\phi_b = 0^\circ$ at both design frequencies.

Two pyramidal feed-horns have been employed to illuminate the antenna in an offset configuration. The phase centers of the horns are located at the following coordinates, according to the reference system shown in Fig. 2-19: $(x_{F1}, y_{F1}, z_{F1}) = (-114, 0, 223)$ mm for Ku-band horn; and $(x_{F2}, y_{F2}, z_{F2}) = (-62, 0, 247)$ mm for Ka-band horn. The angle of radiation (20°) has been chosen to avoid blockage from the feeds, and also to maintain an intermediate value between the angles subtended by the two horns and the z -axis from the reflectarray center, which are 14° and 27° (see Fig. 2-19).

Fig. 2-19 Schematic view of the reflectarray and the two feed-horns in the symmetry plane ($y = 0$).

Conventional $\cos^q(\theta)$ functions, based on electromagnetic simulations of both horns, have been used to model their radiation patterns. The Ku-band horn (model VT140SGAH15SK from Vector Telecom), with a gain at 12 GHz of 14.47 dBi and a 3-dB beamwidth of 30° , is modelled with $q = 10$. On the other hand, the Ka-band horn (model NARDA 638) presents 15.6 dBi gain at 20 GHz, the 3-dB beamwidth is 26.7° , and $q = 13$ has been used for the simulations. Under these conditions, the following illumination levels are reached on the reflectarray edges: -9.8 dB at 12 GHz and -13.8 dB at 19.5 GHz.

The phase-shift distributions required on the reflectarray to generate a collimated beam in the specified direction at 12 and 19.5 GHz are shown in Fig. 2-20 for both polarizations (with the electric field in the direction of the dipoles). Since the antenna is designed to provide the same radiation pattern for the two orthogonal components of the incident field, it will operate in dual-circular polarization (CP) when illuminated by a dual-CP polarized feed. This flexibility may be particularly useful for a multi-mission space antenna, as most of Ku-band transponders operate in linear polarization, while circular polarization is commonly employed in Ka-band transponders.

Fig. 2-20 Required phases (in degrees) to be implemented on the reflectarray surface in X and Y polarizations: (a) at 12 GHz, (b) at 19.5GHz.

A home-made full-wave electromagnetic code that applies the Method of Moments in the Spectral Domain and assumes an infinite periodic array model for the analysis of each cell is used for the design of the antenna (further details about this analysis tool have been provided in section 2.2.2). The SD-MoM code is iteratively called by a zero-finding routine, which is used to obtain the appropriate lengths for each arrangement of dipoles in the way that follows. First, the required phases at 12 GHz (see Fig. 2-20(a)) are implemented by considering only the elements on layer A, i.e., the lengths l_{A1} to l_{A3} are varied to control X-polarization, and the lengths l_{A4} to l_{A6} are varied to control Y-polarization. Then, once the lower layer dimensions have been fixed, the required phases at 19.5 GHz (see Fig. 2-20(b)) are accomplished by adjusting the elements on layer B in a similar way: the lengths l_{B1} and l_{B2} are varied to control X-polarization, and the lengths l_{B3} and l_{B4} are varied to control Y-polarization.

The single-frequency design provides acceptable results for the antenna radiation patterns at both design frequencies (as shown in section 2.2.3); however, an additional

optimization has been run in order to correct residual phase errors and ensure the required bandwidth in Ku and Ka bands. This process involves the fine tuning of the lengths previously calculated, with the aim of fitting not only the required phases at the central design frequencies, but also the phases at the extremes of both operating bands: 2 GHz bandwidth is enforced in Ku-band (11-13 GHz, 16%), and 1 GHz bandwidth in Ka-band (19-20 GHz, 5.2%). The dipoles in the direction of the x and y axes can be independently optimized due to the uncoupling of the phases for the two orthogonal polarizations, thus simplifying the whole optimization process [30].

2.3.3 *Manufacturing of the demonstrator*

The designed prototype for this reflectarray antenna has been manufactured and measured. The two levels of printed dipoles are produced by conventional chemical photo-etching process. The dipoles are printed on both sides of a DiClad 880B substrate cladded with 17- μm copper. Then, the dielectric layer with the printed dipoles has been bonded to an AD255C sheet backed by the ground plane by using a thermoplastic film (CuClad 6250). Fig. 2-21 shows the lateral view of the sandwich configuration for the reflectarray prototype.

Fig. 2-21 Sandwich configuration of the reflectarray (lateral view).

The layouts with the dimensions of the dipoles on each layer are generated in AutoCAD, using the dipoles' lengths that were computed in the design of the demonstrator. In the case of the lower layer, which contains the longest dipoles, a minimum distance of 110 μm has been enforced between orthogonal dipoles in neighbour cells. The reason for this provision is to avoid potential contacts between the dipoles that control each linear polarization, which may be produced because of the

tolerances in the photo-etching of the dipoles. The masks corresponding to the two layers of the reflectarray, the bottom layer and the upper one, are shown in Fig. 2-22 and Fig. 2-23, respectively.

Fig. 2-22 Photo-etching mask for the bottom layer of the reflectarray demonstrator and detail of the dipoles.

Fig. 2-23 Photo-etching mask for the upper layer of the reflectarray demonstrator.

An aluminum structure has been built at the facilities of the Applied Electromagnetics Group of Universidad Politécnica de Madrid to properly sustain the reflectarray (an AutoCAD scheme of this structure is shown in Fig. 2-24). The structure includes an arm with methacrylate supports to hold the two feed-horns. The antenna has been fixed to the aluminum supporting plate by means of nylon screws which are placed near the corners of the reflectarray surface. These parts of the antenna will be less illuminated than the center, so the distortion produced by the screws will be almost imperceptible. The resulting breadboard can be seen in Fig. 2-25.

Fig. 2-24 AutoCAD scheme with the structure of the demonstrator.

Fig. 2-25 Manufactured reflectarray demonstrator at UPM facilities.

2.3.4 Measurement of the demonstrator and comparison with simulations

The manufactured reflectarray antenna has been tested at the anechoic chamber of Universidad de Sevilla, in a spherical near-field measurement system (see Fig. 2-26). The radiated field has been measured for the two orthogonal polarizations X and Y in the angular range $-80^\circ < \theta < 80^\circ$, $-90^\circ < \phi < 90^\circ$. The dimensions of the printed dipoles were checked before the measurement of the antenna, and this test showed that there had been an average error of 50 μm in excess in all the lengths and widths of the dipoles of both layers, plus a random error of $\pm 10 \mu\text{m}$. This tolerance error was taken into account in the calculation of numerical radiation patterns, showing that it can lead to a reduction in gain of about 1 dB at Ka-band frequencies, while the effects in Ku-band (where the tolerance error is smaller in terms of the wavelength) were almost negligible.

Fig. 2-26 Reflectarray prototype and measurement setup.

A comparison between the measured and simulated radiation patterns in the principal planes has been performed for different frequencies in Ku and Ka bands (see Fig. 2-27 to Fig. 2-32, and Fig. 2-34 to Fig. 2-37). The patterns include the co- and cross-polar components of the radiated field for the two linear polarizations in the principal planes: the one tilted 20° with respect to z -axis (azimuth) and the xz -plane (elevation). The SD-MoM code has been used to calculate the tangential components of the reflected electric and magnetic fields on the reflectarray surface. Then, the numerical radiation patterns have been derived from these components.

Figures 2-27 and 2-28 show a quite good agreement between experimental and simulated radiation patterns at 12 GHz (the central design frequency in Ku-band) for X

and Y polarizations, respectively, although there is a gain loss of 1.2 dB with respect to the simulations performed using the nominal values of ϵ_r and $\tan\delta$ at 10 GHz provided by the manufacturer. The patterns at the extreme frequencies in Ku-band (11 and 13 GHz) are shown in Fig. 2-29 to Fig. 2-32 for both X and Y polarizations. As can be seen, the measurements fit quite acceptably the simulations and present the same discrepancies as the patterns obtained at 12 GHz.

Fig. 2-27 Measured and simulated radiation patterns at 12 GHz for X-polarization in (a) azimuth and (b) elevation planes.

(a)

(b)

Fig. 2-28 Measured and simulated radiation patterns at 12 GHz for Y-polarization in (a) azimuth and (b) elevation planes.

(a)

(b)

Fig. 2-29 Measured and simulated radiation patterns at 11 GHz for X-polarization in (a) azimuth and (b) elevation planes.

Fig. 2-30 Measured and simulated radiation patterns at 11 GHz for Y-polarization in (a) azimuth and (b) elevation planes.

(a)

(b)

Fig. 2-31 Measured and simulated radiation patterns at 13 GHz for X-polarization in (a) azimuth and (b) elevation planes.

(a)

(b)

Fig. 2-32 Measured and simulated radiation patterns at 13 GHz for Y-polarization in (a) azimuth and (b) elevation planes.

On the other hand, the comparison between simulated and measured radiation patterns in Ka-band showed stronger disagreements than in Ku-band. It was checked that the tolerance errors in the lengths and widths of the dipoles (an average increment of 50 μm) were not large enough to justify the differences. The bonding film (CuClad 6250, with $\epsilon_r = 2.32$ and $\tan\delta = 0.0013$), which had been originally neglected in the design of the antenna because of its small thickness (38 μm), was also considered in the simulations by the inclusion of a new dielectric layer between layers A and B. As a result, slight differences were detected in the radiation patterns, but they were not sufficient to justify the difference between measured and simulated gain in Ka-band.

Afterwards, the dielectric constant and loss tangent of both substrates were measured by using the technique described in [105]. The results showed that the dielectric constant in Ka-band is always larger than the nominal values provided by the manufacturer at 10 GHz. For AD255C, $\epsilon_{rA} = 2.7$ was measured, instead of $\epsilon_{rA} = 2.55$, and for Diclad 880B, $\epsilon_{rB} = 2.3$ was measured, instead of $\epsilon_{rB} = 2.17$, being in both cases around 6% higher than nominal values (1% tolerance is specified by the manufacturer at 10 GHz). The measured loss tangent is also larger than the nominal values: $\tan\delta$ close to 0.005 was measured for both substrates, instead of $\tan\delta = 0.0009$. Therefore, additional simulations were carried out by taking into account the measured values of permittivity and loss tangent of layers A and B. As a consequence, the corrected radiation patterns in Ka-band showed a better correspondence with experimental results (as will be shown later, see Fig. 2-34), while the patterns in Ku-band presented a slight reduction in gain (see Fig. 2-27 to Fig. 2-32 for the simulated patterns in Ku-band with the corrected values of ϵ_r and $\tan\delta$).

The amplitude and phase response of the reflectarray cell was also analyzed considering the corrected values of ϵ_r and $\tan\delta$, and the results for X-polarization are presented in Fig. 2-33, superimposed to the curves obtained with the nominal values of ϵ_r and $\tan\delta$ that were shown in Fig. 2-16. As can be seen, the worst effects occur in Ka-band, where the losses can be several times higher than originally expected. Moreover, significant phase errors appear in those elements whose upper central dipoles present length values larger than 5.5 mm (for example, these errors can be up to 80° at 20 GHz in some cells of the reflectarray), being responsible for the deterioration of the antenna radiation patterns.

Fig. 2-33 Magnitude and phase of the cell reflection coefficient, considering X-polarization and $\theta_i = 20^\circ$ incidence: (a) at Ku-band, (b) at Ka-band.

Finally, it has been found that there are a few cells near the edge of the reflectarray, for which the first higher order Floquet mode starts propagating at the higher frequencies, which is the condition for grating lobe appearance. The analysis tool has been modified to take into account the contribution of higher order Floquet harmonics in the radiated field and the antenna has been analyzed again.

A slightly better agreement between simulations and measurements was found after inclusion of higher order Floquet modes; however, the main source of discrepancies between the original simulations and measurements is not the excitation of grating lobes, but the difference between nominal and measured values of ϵ_r and $\tan\delta$ in the materials, as can be checked in Fig. 2-34. This figure shows the comparison of the experimental and numerical radiation patterns for X-polarization at 19.5 GHz (central design frequency in Ka-band), considering the effect of all the aforementioned factors in the simulations: the correction performed in the values of ϵ_r and $\tan\delta$, the inclusion of higher order Floquet modes and the sum of the two previous factors. As can be seen, the differences between the patterns in green line (corrected values of ϵ_r and $\tan\delta$) and red line (corrected values of ϵ_r and $\tan\delta$ and higher order Floquet modes) are very small, providing a reasonably good agreement with the measured patterns at 19.5 GHz.

The bad initial estimation of the dielectric constant of the substrate layers shifted the antenna operation in the upper band to lower frequencies (from 19-20 GHz to around 18.5-19.5 GHz), while the higher measured loss tangent of the substrate layers also had an effect on reducing the expected gain of the antenna, especially in Ka-band. The simulated corrected patterns in Ka-band for X and Y polarizations, considering both the measured values of ϵ_r and $\tan\delta$ and the effect of higher order Floquet modes, are shown in Fig. 2-34 and Fig. 2-35 for operation at 19.5 GHz, and in Fig. 2-36 and Fig. 2-37 for operation at 19 GHz. As can be seen, a gain of 27.6 and 26.3 dBi is attained at 19 GHz for the beams in X and Y polarizations, respectively, which is around 3 dB lower than the originally expected, according to the simulations performed with the nominal values of ϵ_r and $\tan\delta$ of the materials. Due to the shift in the frequency response, the patterns present further deterioration at 19.5 GHz, where the measured gain is about 4 dB lower than the expected for X-polarization and 6 dB lower for Y-polarization. The inclusion of the aforementioned correction factors in the simulations contributes to obtain a better agreement with the measurements, causing a reduction in gain and an increase of side lobes, as can be observed in the patterns.

(a)

(b)

Fig. 2-34 Measured and simulated radiation patterns at 19.5 GHz for X-polarization in (a) azimuth and (b) elevation planes.

(a)

(b)

Fig. 2-35 Measured and simulated radiation patterns at 19.5 GHz for Y-polarization in (a) azimuth and (b) elevation planes.

(a)

(b)

Fig. 2-36 Measured and simulated radiation patterns at 19 GHz for X-polarization in (a) azimuth and (b) elevation planes.

(a)

(b)

Fig. 2-37 Measured and simulated radiation patterns at 19 GHz for Y-polarization in (a) azimuth and (b) elevation planes.

Figure 2-38 shows the evolution of gain with frequency in each polarization for both measurements and simulations, considering first the nominal values of ϵ_r and $\tan\delta$ of the two dielectric materials, and then, the corrected values. The radiation patterns were also measured at 18 GHz in order to obtain a more complete characterization of the antenna operation in the upper band. As can be seen, there is a shift in the frequency response in Ka-band (the maximum gain is measured at 19 GHz instead of at 19.5 GHz), as a result of the increased value of the measured dielectric constant. It can be observed that the manufactured demonstrator is wideband in the lower band (roughly 20% bandwidth for a gain variation smaller than 2 dB) and narrowband in the upper band (around 5% bandwidth for a gain variation smaller than 2 dB).

(a)

(b)

Fig. 2-38 Measured vs simulated gain graphs in Ku and Ka bands: (a) for X-polarization, (b) for Y-polarization.

The resulting radiation efficiency for the fabricated antenna, calculated as the ratio between the measured (or simulated) gain and the maximum directivity at each frequency, varies between 34% and 43% in Ku-band (between 47% and 56% for the simulations performed with nominal values), and is very low in Ka-band (the highest value is 19% at 19 GHz). The other parameters of the antenna radiation patterns (side-lobe level, SLL, and cross-polar discrimination, XPD) are summarized in Table 2-1 for both simulations and measurements. The radiation patterns and antenna gain should be improved by using the real properties of the materials at the operation frequencies (extracted from measurements) in the design process.

TABLE 2-1
COMPARISON OF ANTENNA PARAMETERS FOR THE KU/KA-BAND DEMONSTRATOR

Freq. (GHz)	SLL meas. (dB)	SLL corrected (dB)	SLL nominal (dB)	XPD meas. (dB)	XPD corrected (dB)	XPD nominal (dB)
Pol. X						
11	13.40	13.71	13.60	21.48	22.31	24.62
12	13.65	17.75	16.51	23.82	22.07	24.79
13	14.15	17.18	18.12	24.03	22.19	26.73
19	16.50	16.34	17.58	21.66	21.36	24.51
19.5	8.47	10.47	15.40	26.18	14.92	19.54
20	2.41	5.34	10.63	18.78	14.43	15.53
Pol. Y						
11	16.02	14.53	13.92	28.85	27.87	27.45
12	15.59	16.84	19.23	24.25	25.52	24.31
13	12.90	17.42	18.59	23.47	25.82	25.58
19	12.03	17.76	14.55	18.67	21.12	22.40
19.5	7.48	7.71	20.25	18.98	16.54	19.49
20	0.12	0.39	10.55	14.02	11.24	18.59

2.3.5 Conclusions

A reflectarray antenna capable of operating independently in the transmit frequencies from the satellite in Ku-band (11-13 GHz) and Ka-band (19-20 GHz) has been proposed and demonstrated. The reflectarray cell is composed of two orthogonally-arranged sets of coupled parallel dipoles, which are distributed in a two-level configuration. A 25-cm reflectarray prototype has been designed, manufactured and tested. The demonstrator produces a collimated beam in dual polarization which radiates in the direction $\theta_b = 20^\circ$. The results of the measurements show a quite good agreement with the simulations in

Ku-band, and some discrepancies at Ka-band mainly due to a variation in the electrical properties of the materials (ϵ_r and $\tan\delta$), which produces a more severe impact at higher frequencies. To avoid this problem, the materials should be accurately characterized by measuring the dielectric constant and loss tangent at the operation frequencies, and using then these values in the design of the antenna.

The dual-frequency reflectarray demonstrator is a proof-of-concept to show that the proposed reflectarray can operate independently in two frequencies (Ku and Ka bands) and two polarizations with separate feeds. This concept can be applied to transmit antennas in Telecom satellite systems to reuse the same aperture for the generation of a prescribed contoured beam in Ku-band and multiple spots in Ka-band with different feed chains for each mission, which would result in significant savings in the cost, volume and weight of the antenna farm.

2.4 Design of dual polarized reflectarrays to operate at transmit and receive frequencies in Ka-band

As explained in Chapter 1, Ka-band currently represents the main alternative for satellite systems to satisfy the growing demand for capacity. Modern satellite antennas in Ka-band are required to generate a large number of high-gain overlapping spot beams, with a very small separation between adjacent spots [82]. This factor, combined with the deployment of frequency and polarization reuse schemes, leads to an increase in the users' data rates and the overall capacity of the network, enabling the provision of high speed broadband services in Ka-band. From the users' side, conventional reflectors [106] and phased arrays [107] have been proposed for transmit and receive terminal antennas; however, the different frequencies for uplink (30 GHz) and downlink (20 GHz) lead to a more complex antenna design.

Reflectarray antennas are able to generate independent beams in each polarization, provide high values of gain and radiation efficiency, and operate simultaneously at different frequencies. These capabilities can be of particular interest for the design of both terminal and satellite antennas in Ka-band, but first, it requires to count on appropriate reflectarray cells that will provide an independent control of the phase at each frequency band (Tx and Rx) and/or polarization. For this purpose, the cell configuration proposed in the previous sections has been adapted to allow independent

operation in dual-linear polarization at Tx and Rx frequencies in Ka-band. Afterwards, the element has been used to design first a 20-cm reflectarray VSAT (Very Small Aperture Terminal) antenna to generate a focused beam in dual polarization (linear or circular) at 19.7 GHz and at 29.5 GHz, and then, a 1.6-m satellite antenna to produce two closely spaced beams in orthogonal linear polarizations at both Tx and Rx frequencies in Ka-band.

2.4.1 Design of the reflectarray cell

The unit-cell presents the same characteristics than the one used in the previous section. It consists of two orthogonal sets of five parallel dipoles printed on a dielectric layer, and two additional sets of three parallel dipoles stacked above the first sets and printed on the top of a second dielectric sheet (see Fig. 2-39). The period, $P_X = P_Y = 6.5$ mm, is chosen as $2\lambda/3$ at the higher design frequency (29.5 GHz) to avoid the appearance of grating lobes up to 30° incidence. Both dielectric layers have been implemented by Diclad 880 sheets; their electrical properties are $\epsilon_{rA} = \epsilon_{rB} = 2.17$, $\tan\delta_A = \tan\delta_B = 0.0009$, and their thickness is $h_A = 1.5$ mm, $h_B = 1$ mm.

Fig. 2-39 View of the reflectarray periodic structure, including four unit-cells for X polarization and one unit-cell for Y polarization.

The geometrical parameters of the cell have been fixed after a careful parametric study, in order to provide a smooth variation in the phase response in a range larger than 360° in both frequency bands (19.2-20.2 GHz and 29-30 GHz). The dipole width is set to 0.25 mm, separations between laterally coupled dipoles are $S_{XA} = S_{YA} = 0.5$ mm, $S_{XB} = S_{YB} = 1$ mm, and relative sizes of lateral dipoles are $l_{A1} = 0.65 \cdot l_{A3}$, $l_{A2} = 0.8 \cdot l_{A3}$, $l_{A4} = 0.65 \cdot l_{A6}$, $l_{A5} = 0.8 \cdot l_{A6}$ (where l_{A3} and l_{A6} correspond to the lengths of the central dipoles

on the lower layer), $l_{B1} = 0.8 \cdot l_{B2}$, $l_{B3} = 0.8 \cdot l_{B4}$ (where l_{B2} and l_{B4} correspond to the central dipoles' lengths on the upper layer). The amplitude and phase curves of the cell reflection coefficient as a function of the dipole lengths are shown in Fig. 2-40 at the central and extreme frequencies of each band, considering X-polarization and normal incidence conditions.

Fig. 2-40 Phase and amplitude of the cell reflection coefficient for X-polarization under normal incidence: (a) at Tx band and (b) at Rx band.

Furthermore, the phase curves of the cell reflection coefficient exhibit a robust response against variations in the angle of incidence (see Fig. 2-41), which have almost no effect at the lower frequency band, while producing slight variations in the phase-shift at the higher frequencies.

Fig. 2-41 Variation with the angle of incidence in the phase of the cell reflection coefficient for X-polarization: (a) at 19.7 GHz, and (b) at 29.5 GHz.

A practically independent phase control based on the lengths of upper and lower dipoles can be observed in Fig. 2-42, which shows the variation in the phase of the cell reflection coefficient at 19.7 and 29.5 GHz, for X-polarization under normal incidence (note that the same response is obtained for Y-polarization, although it is not shown here). As can be seen, the upper dipoles will not disturb the phase response at 19.7 GHz, as they are shorter than the ones in the bottom layer, while lower dipoles will have a certain influence on the phase at 29.5 GHz. Therefore, the dipole lengths can be obtained separately for each band, as in the case of the Ku/K-band reflectarray designed in the previous section: first, the dipoles on the bottom layer are adjusted to provide the required phase-shift at 19.7 GHz, and then those on the top layer are adjusted to provide the required phase at 29.5 GHz.

Fig. 2-42 Phase (in degrees) of the cell reflection coefficient with respect to the lengths of the dipoles in both layers, considering X-polarization (a) at 19.7 GHz and (b) at 29.5 GHz.

2.4.2 Design of a Tx/Rx terminal SatCom antenna in Ka-band

A 20-cm sided reflectarray, consisting of 900 elements arranged in a 30 x 30 grid, has been designed to generate a focused beam in the direction $\theta_b = 13^\circ$, $\phi_b = 0^\circ$ for the two orthogonal polarizations (with the electric field in the direction of the dipoles) at 19.7 and 29.5 GHz. Since the antenna is designed to provide the same radiation pattern for the two orthogonal components of the incident field, it will operate in dual-circular polarization when it is illuminated by a dual-circularly polarized feed-horn.

The phase center of the feed is placed at coordinates $(x_F, y_F, z_F) = (-40, 0, 195)$ mm relative to the geometrical center of the reflectarray (origin of the coordinates system). The electromagnetic field radiated by the feed-horn is modeled using a $\cos^q(\theta)$ distribution, with $q = 10.5$ for 20 GHz band and $q = 10.7$ for 30 GHz band.

The SD-MoM electromagnetic code has been employed to adjust the lengths of the dipoles in both layers, in order to provide the phase-shift distributions shown in Fig. 2-43 for both polarizations, at 19.7 GHz and 29.5 GHz, considering the real angles of incidence on each reflectarray element (see Fig. 2-44). Then, the dimensions of all dipoles are optimized element-by-element to simultaneously match the phases at the central and extreme frequencies in the lower (19.2-20.2 GHz) and higher (29-30 GHz) frequency bands, following a procedure similar to the one described in section 2.3.2 for the design of the Ku/Ka-band reflectarray demonstrator.

Fig. 2-43 Required phase-shift distributions (in degrees) to be implemented on the reflectarray for both polarizations: (a) at 19.7 GHz, (b) at 29.5 GHz.

Fig. 2-44 Angles of incidence (in degrees) from the feed on each reflectarray cell: (a) theta, (b) phi.

2.4.2.1 Results of the simulations

The simulated radiation patterns in gain (dBi) have been obtained from the tangential electric and magnetic reflected fields at each reflectarray cell, using the aforementioned SD-MoM software. Fig. 2-45 shows the simulated radiation patterns of the reflectarray VSAT antenna at 19.7 GHz in the elevation and azimuth orthogonal planes, including the co- and cross-polar components of each linear polarization. A gain of 31.4 dBi and 31 dBi is achieved for the beams in X and Y polarizations, respectively, with side-lobe levels (SLL) close to -22 dB with respect to the co-polar maximum. The cross-polar discrimination (XPD), measured within a 3-dB beamwidth (4.72°) of the main lobe, is around 33 dB.

(a)

(b)

Fig. 2-45 Simulated radiation patterns in gain (dBi) at 19.7 GHz for X and Y polarizations: (a) xz -plane (elevation), (b) orthogonal plane in the direction of the beam (azimuth).

On the other hand, Fig. 2-46 shows the simulated radiation patterns of the reflectarray VSAT antenna at the central design frequency in the higher band (29.5 GHz). In this case, a gain of 34.2 dBi and 33.8 dBi is achieved for the beams in X and Y polarizations, respectively, with XPD around 31 dB and SLL close to -20 dB with respect to the maximum. The 3-dB beamwidth is now 3.15°.

(a)

(b)

Fig. 2-46 Simulated radiation patterns in gain (dBi) at 29.5 GHz for X and Y polarizations: (a) xz -plane (elevation), (b) orthogonal plane in the direction of the beam (azimuth).

After the multi-frequency optimization, an 8% bandwidth can be achieved in the lower frequency band with a gain variation of 1 dB for X-polarization and 1.5 dB for Y-polarization. The simulated radiation patterns in the principal planes at the extreme frequencies of this band, 18.9 GHz and 20.5 GHz, are shown in Fig. 2-47 and Fig. 2-48, respectively. In the case of the higher frequency band, a 5% bandwidth can be achieved with the same gain variation (1 dB for X-polarization and 1.5 dB for Y-polarization). The simulated radiation patterns in the principal planes at the extreme frequencies, 28.8 GHz and 30.2 GHz, are presented in Fig. 2-49 and Fig. 2-50, respectively.

Fig. 2-47 Simulated radiation patterns at 18.9 GHz for the VSAT reflectarray antenna, for X and Y polarizations: (a) xz-plane, (b) orthogonal plane in the direction of the beam (azimuth).

Fig. 2-48 Simulated radiation patterns at 20.5 GHz for the VSAT reflectarray antenna, for X and Y polarizations: (a) xz-plane, (b) orthogonal plane in the direction of the beam (azimuth).

Fig. 2-49 Simulated radiation patterns at 28.8 GHz for the VSAT reflectarray antenna, for X and Y polarizations: (a) xz-plane, (b) orthogonal plane in the direction of the beam (azimuth).

Fig. 2-50 Simulated radiation patterns at 30.2 GHz for the VSAT reflectarray antenna, for X and Y polarizations: (a) in the azimuth plane, (b) in the elevation plane.

The antenna efficiency has been estimated at 67% in the lower band and 57% in higher band, considering illumination, spillover and dielectric losses. These results show the potential of reflectarrays for dual-frequency and dual-polarization operation as a low-cost alternative for Ka-band terminal antennas.

2.4.3 *Design of a Tx/Rx satellite antenna in Ka-band*

A circular reflectarray antenna, consisting of 49,080 elements arranged in a 250 x 250 grid (162.5 cm diameter), has been designed to generate a focused beam in the direction $\theta_b = 10^\circ$, $\varphi_b = 0^\circ$ for X-polarization and a closely spaced beam at $\theta_b = 10.5^\circ$, $\varphi_b = 0^\circ$ for Y-polarization at 19.7 GHz and 29.5 GHz, which are transmit and receive frequencies for multi-spot satellite antennas in Ka-band.

The reflectarray is illuminated by a corrugated feed-horn, whose phase center is placed at coordinates $(x_F, y_F, z_F) = (-300, 0, 1000)$ mm with respect to the reflectarray center. The field radiated by the horn is modeled using a $\cos^q(\theta)$ function, with $q = 6$ for the Tx band and $q = 8$ for the Rx band. The edge illumination levels are close to -16 dB at 19.7 GHz and -18 dB at 29.5 GHz.

The required phase-shift distributions to be introduced by the reflectarray elements in each frequency (19.7 and 29.5 GHz) and polarization (X and Y) are shown in Fig. 2-51. The lengths of the dipoles on both layers are adjusted by using the SD-MoM software tool, which takes into account the real angles of incidence (θ_i , φ_i) on each reflectarray cell (see Fig. 2-52) to calculate the amplitude and phase of the reflection coefficient, considering the element in a periodic environment.

Fig. 2-51 Phase-shift distributions (in degrees) to be introduced by the reflectarray: at 19.7 GHz (a) in X-polarization and (b) in Y-polarization, and at 29.5 GHz (c) in X-polarization and (b) in Y-polarization.

Fig. 2-52 Angles of incidence (in degrees) from the feed on each reflectarray cell: (a) theta, (b) phi.

In this case, the design of the reflectarray elements has been performed only at the central frequencies of each band. First, the dipoles in the lower layer are adjusted to match the required phases at 19.7 GHz, and then, the dipoles on the top layer are adjusted to do the same with the phases at 29.5 GHz. The execution of a multi-

frequency optimization routine could improve the antenna performance within each operating band by a fine tuning of the dipole lengths, but it would be a very time-consuming process due to the extremely large number of elements (it would require several days for optimizing the whole antenna). On the other hand, the design of the dipoles at the central frequency of each band takes only a couple of hours.

2.4.3.1 Results of the simulations

The simulated radiation patterns of the 1.6-m reflectarray antenna in the elevation and azimuth planes are shown in Fig. 2-53 (at 19.7 GHz) and Fig. 2-54 (at 29.5 GHz).

Fig. 2-53 Simulated radiation patterns in gain (dBi) at 19.7 GHz for X and Y polarizations: (a) xz -plane (elevation), (b) orthogonal plane in the direction of the beam (azimuth).

(a)

(b)

Fig. 2-54 Simulated radiation patterns in gain (dBi) at 29.5 GHz for X and Y polarizations: (a) xz -plane (elevation), (b) orthogonal plane in the direction of the beam (azimuth).

As can be seen, the proposed reflectarray is able to generate two closely spaced beams (only 0.5° of spacing) in orthogonal polarizations both at Tx and Rx frequencies, when the antenna is illuminated by a single feed operating in dual-linear polarization. A gain of 48.3 dBi is attained at 19.7 GHz, and 50.7 dBi gain is reached at 29.5 GHz, with low levels for cross-polar components (around 30 dB below the co-polar maximum). The 3-dB beamwidth is between 0.6° and 0.7° , and SLL is close to -30 and -25 dB with respect to the co-polar maximum at 19.7 and 29.5 GHz, respectively. Finally, the radiation efficiency of the 1.6-m reflectarray can be estimated as 66% at the Tx band and 48% at the Rx band, which are typical values for conventional reflectors.

2.4.4 Conclusions

These results show the potential of reflectarrays for working at transmit and receive frequencies in Ka-band, with independent beam-shaping in each polarization and frequency band. The dimensions of the reflectarray element used in sections 2.2 and 2.3, based on two stacked layers with orthogonal sets of parallel dipoles, have been modified to allow simultaneous phase adjustment in both linear polarizations at 19.7 and 29.5 GHz, providing a linear phase variation within a 360° range and a robust behaviour with respect to variations in the angle of incidence.

The simulated radiation patterns are very promising for terminal SatCom antennas and multi-spot satellite antennas that operate in Ka-band. The proposed 20-cm reflectarray VSAT antenna is able to generate a focused beam in dual-polarization (linear or circular) at both transmit and receive frequencies in Ka-band. The reduced number of layers and the simplicity of the printed elements make it a low-cost alternative to reflectors and phased arrays. On the other hand, the 1.6-m satellite reflectarray antenna produces two adjacent beams (0.5° of spacing) in orthogonal linear polarizations at both Tx and Rx bands, when the antenna is illuminated by a single dual-polarized feed. In this case, the beams can also be generated in dual-circular polarization if the orthogonal polarizations are discriminated by means of a sequential rotation technique applied to the reflectarray elements, as proposed in [59].

2.5 Conclusions

In this chapter, a novel reflectarray cell has been proposed and demonstrated to operate at two separate frequencies in dual polarization. The element consists of two orthogonal sets of parallel dipoles, arranged in a two-layer configuration. Each set is composed of five parallel dipoles on the lower layer, and three additional parallel dipoles which are stacked above the previous ones and are printed on the top of a second dielectric sheet. The geometrical parameters of the cell have been adjusted to operate, first, in transmission in Ku and Ka bands (12 and 20 GHz), and then, in transmission and reception in Ka-band (20 and 30 GHz), providing a smooth variation in the phase of the co-polar reflection coefficient with respect to the dipoles' lengths and covering a range of phase greater than 360° at both design frequencies.

The phase-shift introduced in each linear polarization is controlled by the orthogonal sets of dipoles, which are practically uncoupled. Moreover, the dipoles on the lower layer will provide the required phases at the lower frequency, while the dipoles on the upper layer will do the same with the phases at the higher frequency. This operating principle is made possible as the lengths of the lower dipoles are larger than those of the upper dipoles, providing an independent control of the phase at each design frequency. This allows for performing separate design processes for each reflectarray layer: first, the lengths of the lower dipoles are adjusted to match the phases at the lower frequency, and then, the lengths of the upper dipoles are adjusted to introduce the required phase-shift at the higher frequency, while correcting the effect of the bottom dipoles at the same time.

The proposed element has been used to design several reflectarray antennas, first to operate at Tx frequencies in Ku and Ka bands, and then at Tx and Rx frequencies in Ka band. These designs show the capability of reflectarrays to generate independent beams in each polarization at both frequency bands. Also, a 25-cm demonstrator that operates in Ku (11-13 GHz) and Ka (19-20 GHz) bands in dual polarization (linear or circular) has been designed, manufactured and tested in order to validate the concept. The results of the measurements show a quite good agreement with the simulations in Ku-band, and some discrepancies in Ka-band due to a variation in the electrical properties of the dielectric sheets, which produces a more severe impact at higher frequencies. This problem can be avoided by an accurate characterization of the materials, before performing the design of the antenna.

The proposed concept can be applied to design a reflectarray antenna which is able to fulfill independent requirements at each frequency and/or polarization (different missions). For example, a contoured beam can be generated in Ku-band and multiple spots can be obtained in Ka-band by properly designing the elements on each reflectarray layer, considering different feed chains for each mission. The reuse of the same aperture for both missions would result in significant savings in the costs, weight and volume of the antenna farm, especially in the case of telecommunication satellites which operate in Ku and Ka bands.

Chapter 3

Application of the bifocal technique to dual reflectarray configurations

3.1 Introduction

The bifocal technique has been considered for many years as one of the main alternatives for the design of multiple beam and wide-angle beam scanning antennas. The bifocal concept was introduced in the 1950s, concerning the design of two-surface axially-symmetrical dielectric lenses which are able to collimate the rays from a point source, placed at either of two conjugate off-axis locations, into a plane wave forming an angle of $\pm\theta_0$ with respect to the symmetry axis [108], [109]. In these conditions, the bifocal design provides an improved performance for the off-axis beams (in terms of gain, side-lobe levels and beam shaping) with respect to the equivalent single focused case, at the cost of a slightly worse performance for the central beam.

On the basis of dielectric lenses, the bifocal technique was adapted to allow the simultaneous design of dual reflector systems for two focal points [80], [81], so that two feeds placed at the foci will produce two beams emerging at directions θ_1 and θ_2 by properly shaping both reflectors. Note that at least two reflectors are required to have enough degrees of freedom for the design. The appropriate shape of each reflector can be obtained by means of a geometrical optics (GO) procedure, based on the application of Snell's law of reflection and the equal path-length condition, which involves the interpolation of two sets of points (one set per reflector) and their associated normal vectors. In the case of an axially-symmetrical configuration [80], a simple solution to the 3D design problem can be obtained by rotation of a 2D bifocal design performed in the offset plane (the antenna symmetry plane, which contains the foci and the beams).

This technique allows to design not only centered, but also offset configurations by properly selecting specific parts of the revolution surfaces that are obtained [110].

The extension of the 2D GO procedure to 3D for dual offset configurations without axial symmetry was carried out by Rappaport [81]. However, the design of these offset configurations (such as Cassegrain or Gregorian) has proved to be very challenging. Particularly, it requires to define a set of even-degree polynomials that will determine the shape of each reflector along the antenna cross section. The difficulty in obtaining smooth profiles for both reflectors has led to new design approaches, as the one based on the optimization of the effective area of the bifocal antenna [111]. This method tries to achieve the most efficient illumination of the main reflector by illuminating only a half of the sub-reflector for each of the two beams generated from the foci.

Thanks to their large field of view, several bifocal dual reflectors have been proposed in the last years for wide-angle beam scanning applications, such as THz imaging [112], [113] and satellite ground stations [114]. Moreover, a few papers have been reported on the bifocal design of dual reflectarray antennas (DRA) [17], [115], [116], where the main issue relies on the calculation of the required phase-shift distributions for each reflectarray. The initial works [17], [115] focused on the design of small-size DRAs for automotive radar applications (the main reflectarray diameter is lower than 30 cm), considering centered and rotationally-symmetrical geometries where blockage from the sub-reflectarray is avoided by using a gridded sub-reflector and a 90° twist of polarization on the main reflectarray. The limitation of this configuration is that the antenna can only operate in single linear-polarization. The application of the bifocal technique to offset DRA configurations of larger size was first studied in [116], where the phase distributions on both reflectarrays were approximated from an equivalent offset bifocal dual reflector obtained by applying Rappaport's technique [81], but the results were not satisfactory.

In this chapter, a novel bifocal procedure is proposed for the design of offset dual reflectarray configurations of large size, starting from an axially-symmetrical geometry with parallel reflectarrays in which a 2D GO algorithm is applied, followed by the rotation of the resulting phase curves around the symmetry axis. An offset DRA system can be formed by selected parts of the revolution surfaces, and then, both reflectarrays can be tilted to obtain smoother phase distributions. The proposed technique has been applied to the design of multi-beam satellite antennas in Ka-band, in order to provide a

reduced beam spacing and an improved performance for the extreme beams with respect to the equivalent single-focus antenna.

3.2 Bifocal design procedure for dual reflectarray antennas

The simplest way of implementing the bifocal concept on a dual reflectarray antenna considers the design of an axially-symmetrical configuration (see Fig. 3-1), following a similar approach to the one presented in [80] for the design of dual reflectors. Originally, both reflectarrays are placed in parallel planes, in order to subsequently exploit the symmetry of the configuration by rotation of a 2D bifocal design.

Fig. 3-1 Geometry and main parameters of the bifocal dual reflectarray antenna with parallel reflectarrays, including the first step of the bifocal ray-tracing routine in the xz -plane.

The initial parameters that characterize the geometry of the DRA system and must be fixed before starting the bifocal procedure are: distance between foci (d), distance between the foci and the sub-reflectarray (L_1), distance between the two reflectarrays (L_2), and directions of the radiated beams (θ_{b1} and θ_{b2}). As can be observed in Fig. 3-1, a symmetrical arrangement of the focal points along the x -axis is considered, so that their

coordinates are $(x_{F1}, y_{F1}, z_{F1}) = (-d/2, 0, L_2-L_1)$ and $(x_{F2}, y_{F2}, z_{F2}) = (d/2, 0, L_2-L_1)$. Note that the beam directions are also symmetric with respect to the horizontal axis, so they fulfill $\theta_{b1} = -\theta_{b2}$ (being $\theta_{b1} > 0^\circ$ in the case shown in Fig. 3-1).

The derivative of the phase distribution with respect to the length variable along the reflectarray profile, $\partial\Phi/\partial x$ (also named Φ'_x), will be determined for a discrete set of points (S_i and M_i) on the vertical axis of each reflectarray by means of an iterative 2D ray-tracing routine implemented in the xz -plane (see Fig. 3-1). The samples of Φ'_x will be interpolated by polynomials and then integrated to obtain the bifocal phase functions for each reflectarray in the xz -plane. These functions will be rotated around z -axis, and the resulting phase distributions will enable the design of both centered and offset DRA configurations. A block diagram with the steps of the proposed bifocal algorithm is presented in Fig. 3-2.

Fig. 3-2 Steps of the developed bifocal design procedure which starts by considering an axially-symmetrical DRA configuration.

3.2.1 Iterative ray-tracing routine in 2D

The 2D ray-tracing routine maintains the same philosophy than in the bifocal design of dual reflectors, alternating transmitted and received rays in the same way that is described in [80]. However, a specific phase condition has to be considered on each reflectarray cell, instead of applying the Snell's law of reflection. As shown in [17], the following expression relates the phase derivative on the reflectarray with the angles of the incident (θ_i) and reflected (θ_o) rays:

$$\Phi'_x = \frac{\partial \Phi}{\partial x} = \frac{2\pi}{\lambda} \cdot (\sin \theta_i - \sin \theta_o) \quad (3-1)$$

The first iteration of the ray-tracing algorithm requires an starting point on the sub-reflectarray axis, $S_1 = (x_{S1}, 0, L_2)$, and the value of the phase derivative at that point, $\Phi'_x(S_1)$. Note that the value of $\Phi'_x(S_1)$ cannot be arbitrarily set: for every point on the sub-reflectarray we can obtain the incidence angle θ_{i1} from focus F_1 (being $\theta_{i1} > 0$ in the case that $x_{S1} > -d/2$), and then, eq. (3-1) will fix an upper and a lower limit for the phase derivative at S_1 , as the reflection angles must be between -90° and 90° :

$$\frac{2\pi}{\lambda} \cdot (\sin \theta_{i1} - 1) < \frac{\partial \Phi}{\partial x} < \frac{2\pi}{\lambda} \cdot (\sin \theta_{i1} + 1) \quad (3-2)$$

In the case of Fig. 3-1, where a symmetrical arrangement of the foci and the radiated beams with respect to z -axis is considered, it can be deduced that the phase distributions of both reflectarrays in the xz -plane will present even symmetry with respect to z -axis. The phase will have a maximum or a minimum at $x = 0$, which means that the value of Φ'_x at $x = 0$ will be equal to zero. Hence, the point $S_1 = (0, 0, L_2)$ on the sub-reflectarray with $\Phi'_x(S_1) = 0$ can be selected to start the first iteration of the algorithm.

The bifocal ray-tracing algorithm works in the way that follows. First, a transmitted ray from focus F_1 that impinges on S_1 is used to obtain a new point on the main reflectarray, $M_1 = (x_{M1}, 0, 0)$, and its associated phase derivative, $\Phi'_x(M_1)$, by applying (3-1) first on the sub-reflectarray (where the incidence angle and the phase derivative are known), and then, on the main reflectarray (where both the incidence and output angles are known), see Fig. 3-1. Similarly, if we consider a received ray in the direction θ_{b2} that impinges on M_1 , a new point on the sub-reflectarray, $S_2 = (x_{S2}, 0, L_2)$, and the value of its phase derivative, $\Phi'_x(S_2)$, are obtained by enforcing the ray to reach focus F_2 , following the same procedure than before. Then, a transmitted ray from F_1 that

impinges on S_2 can be used to start a new iteration of the process. After N iterations, two sets of points are originated in the xz -plane, one for the main reflectarray and one for the sub-reflectarray, with their corresponding phase derivatives. A flow chart with the steps of the 2D bifocal ray-tracing algorithm is shown in Fig. 3-3.

Fig. 3-3 Flow chart with the steps of the iterative 2D ray-tracing procedure.

This procedure can be applied in a similar way starting at the point $M_1' = (0, 0, 0)$ on the main reflectarray axis (see Fig. 3-4), also with $\Phi'_x(M_1) = 0$ due to the symmetry of the antenna configuration. A received ray in the direction θ_{b2} that impinges on M_1' will provide a new point on the sub-reflectorarray, S_1' , and the value of its phase derivative, $\Phi'_x(S_1')$, by applying (3-1) on both reflectarrays and enforcing the ray to reach focus F_2 . Then, a transmitted ray from F_1 that impinges on S_1' can be used to obtain a second point on the main reflectarray, M_2' , and its associated phase derivative, $\Phi'_x(M_2')$, thus continuing with the previously described ray-tracing technique. In the end, this procedure allows to double the number of points that characterize each reflectarray, which will improve the accuracy in the calculation of the bifocal phase curves.

Fig. 3-4 Second execution of the iterative ray-tracing routine, starting on the main reflectarray.

3.2.2 Integration of the phase derivatives

The phase derivative samples obtained in the xz -plane are interpolated by means of even polynomials depending on x variable, providing two functions named $\Phi'_{x \text{ SUB}}(x)$ and $\Phi'_{x \text{ MAIN}}(x)$. Then, these functions are integrated to obtain the required bifocal phases on each reflectarray, $\Phi_{\text{SUB}}(x)$ and $\Phi_{\text{MAIN}}(x)$. For simplicity, the integration constant is assumed to be zero, although a different value can be considered in order to

adjust the phase distributions on both reflectarrays (note that the addition of a constant to the phase distributions will have no effect on the antenna radiation patterns). An example of the results that can be achieved after performing two independent executions of the ray-tracing routine, one starting at $S_1 = (0, 0, L_2)$ and the other at $M_1 = (0, 0, 0)$, is shown in Fig. 3-5 and Fig. 3-6, for a DRA system designed at 20 GHz with the following parameters: $d = 20$ cm, $L_1 = 1$ m, $L_2 = 1.5$ m, $\theta_{b1} = 1.5^\circ$ and $\theta_{b2} = -1.5^\circ$.

Fig. 3-5 Interpolation of the phase derivative samples on the: (a) sub-reflectorarray, (b) main reflectarray.

Fig. 3-6 Phase curves obtained after the integration of the phase derivatives on the: (a) sub-reflectorarray, (b) main reflectarray.

3.2.3 Rotation of the phase curves

The previous phase distributions only allow to collimate the beams in the xz -plane. Thus, the results of the bifocal synthesis have to be extended from 2D to 3D, so as to obtain a surface phase distribution for each reflectarray. In this case, taking advantage of the symmetry of the antenna configuration with respect to z -axis and the placement of the two reflectarrays in parallel planes, both phase curves can be rotated in the xy -plane

around z -axis. The rotation implies a change of variable in the phase polynomials obtained in the xz -plane:

$$x \rightarrow \sqrt{x^2 + y^2} \quad (3-3)$$

This process is analogous to the one performed with reflectors [112], and results in a focal ring which contains F_1 and F_2 . This fact improves the antenna performance for the generation of multiple beams, as it extends the focal region out of the xz -plane. The resulting phase distributions allow to design either centered or offset DRA configurations just by selecting specific portions of the phase distributions of both reflectarrays, as can be seen in Fig. 3-7. The offset configuration has the advantage of reducing blockage from the sub-reflectarray, and will be the selected option for the upcoming designs. The appropriate size of each reflectarray for obtaining a proper illumination (which contributes to achieve good radiation efficiency) can be estimated based on the separation of the points obtained in the xz -plane for both reflectarrays after the ray-tracing. Note that the illumination on each reflectarray will be concentrated in the regions where the points are very close one to each other, and therefore these regions should be included in the antenna to avoid a high spillover. On the other hand, the regions with few and disperse points will have a low illumination and can be removed. The size of the main reflectarray, as well as the illumination level at the edges, will determine the gain of the bifocal antenna.

Fig. 3-7 Schematic representation of the DRA system obtained after rotation.

The main drawback for the design of offset configurations with parallel reflectarrays is that the resulting phase-shift distributions present a high number of 360° cycles, especially in the case of the main reflectarray. This reduces the potential bandwidth of the reflectarrays and makes difficult their practical implementation in a real antenna, as the local periodicity approach, commonly used in the electromagnetic analysis of the reflectarray, assumes a smooth variation in phase between adjacent cells. As an example, Fig. 3-8 shows the results obtained for the phase distributions in the previous DRA design performed at 20 GHz with parallel reflectarrays, whose bifocal phase curves were shown in Fig. 3-6. An offset configuration has been selected to minimize blockage: the points from $x = 0.2$ m to $x = 0.8$ m have been chosen for constituting the sub-reflectarray (0.6 m diameter), and the associated points from $x = 0.8$ m to $x = 2.6$ m, for the main reflectarray (1.8 m diameter). The cell period in both reflectarrays is 10 mm. The large size of the main reflectarray and the initial conditions of the bifocal synthesis (parallel reflectarrays) are responsible for the high number of 360° cycles that can be seen in Fig. 3-8.

Fig. 3-8 Bifocal phase distributions (in degrees) for: (a) the sub-reflectarray and (b) the main reflectarray.

3.2.4 *Reflectarray tilting and correction of the phase distributions*

The use of an axially-symmetrical geometry with parallel reflectarrays has been considered due to the possibility of applying rotation to the bifocal phase curves in the xz -plane, which provides a simple 3D solution for the design of the bifocal DRA. However, a more natural configuration would present both reflectarrays tilted a certain angle with respect to x -axis, trying to assimilate as much as possible to an equivalent Cassegrain reflector. Consequently, the geometry of the DRA is modified in the way

that is shown in Fig. 3-9, where θ_S and θ_M are the tilt angles of the sub- and main reflectarrays, respectively.

Fig. 3-9 Geometry of the dual reflectarray antenna: (a) initially, (b) after tilting both reflectarrays.

Note that both reflectarrays are tilted about their geometrical centers and that the foci are rotated together with the sub-reflector, in order to maintain the same angles of incidence on the sub-reflector. Furthermore, the beams are pointed in the same directions with respect to z -axis (in the absolute reference system) than in the original design, although the relative directions with respect to the normal vector to the main reflectarray surface ($\hat{z}_M$) are now $\theta_M + \theta_{b1}$ and $\theta_M + \theta_{b2}$.

The values of the angles θ_S and θ_M that will provide the smoothest variation in the final phase distributions of both reflectarrays can be estimated from the unwrapped bifocal phase curves obtained in the xz -plane, or phase delay not limited to 360° (called Φ^d), as follows:

$$\theta_S \approx \sin^{-1} \left[\frac{\max(\Delta\Phi_S^d) \cdot \lambda/2\pi}{D_S/2} \right] \quad (3-4)$$

$$\theta_M \approx 0.5 \cdot \sin^{-1} \left[\frac{\max(\Delta\Phi_M^d) \cdot \lambda/2\pi}{D_M/2} \right] \quad (3-5)$$

where D_S and D_M are the reflectarray diameters, and $\max(\Delta\Phi_S^d)$ and $\max(\Delta\Phi_M^d)$ are the maximum variations in the phase delay (unwrapped) along each reflectarray in the xz -plane with respect to the central element (which typically will be associated to one of the extremes of the reflectarray). Note that a higher number of 360° cycles in the phase-shift distributions will require a larger tilt angle to compensate them. In the case of the main reflectarray, the inclusion of a 0.5 multiplying factor in the estimation of θ_M is due to the decision of keeping the original beam directions in the absolute reference system, which compensates half of the required inclination for the main reflectarray. Otherwise, if the original beam directions with respect to $\hat{z}_M$ were maintained, the 0.5 multiplying factor should be eliminated and the main reflectarray should be tilted an angle which is double the previous value. This would produce greater impact on the antenna geometry, degrading the performance of the phase distributions obtained by rotation in the cross section to the xz -plane, as well as increasing the chances of having blockage from the sub-reflectarray (see Fig. 3-9). For these reasons, it is preferable to keep the original beam directions and work with smaller tilt angles.

The tilt of both reflectarrays must be compensated in their phase-shift distributions, ensuring that the bifocal characteristic of the original design remains. A novel phase adjustment technique has been implemented in order to compensate the effect of the changes that have been introduced in the antenna geometry, following a similar procedure to the bifocal ray-tracing routine described in section 3.2.1. The objective is to obtain a set of points for each reflectarray in the xz -plane with the appropriate phases that compensate the variations in path length (ΔPL) from the original configuration with parallel reflectarrays to the new tilted antenna, considering $\Delta PL_{i,j}$ as associated to a ray impinging first on S_i (on the sub-reflectarray) and then on M_j (on the main reflectarray). Again, the procedure alternates transmitted and received rays. The phase constants $\Delta\Phi(S_i)$ and $\Delta\Phi(M_j)$ that will be added to the phase-shift introduced by the reflectarray cells must fulfill:

$$(2\pi/\lambda) \cdot \Delta PL_{i,j} = \Delta\Phi(S_i) + \Delta\Phi(M_j) \quad (3-6)$$

The algorithm starts with a transmitted ray from F_1 that impinges first on the sub-reflectarray at point S_1 , and then on the main reflectarray at point M_1 (see Fig. 3-10(a)). After tilting both reflectarrays, the ray has to be enforced to reach the same point M_1 on the main reflectarray (note that S_1 will remain the same, as the foci are rotated together with the sub-reflectarray), and then, the ray will be reflected with an angle $\theta_M + \theta_{b1}$

respect to $\hat{z}_M$ (instead of θ_{b1}), so that there will be a variation in the path length equal to $\Delta PL_{1,1}$ with respect to the case with parallel reflectarrays. It is required to set an initial value for one of the phase constants associated to S_1 and M_1 , e. g., $\Delta\Phi(S_1) = \Delta\Phi_0$. Then, the other constant can be obtained as: $\Delta\Phi(M_1) = (2\pi/\lambda) \cdot \Delta PL_{1,1} - \Delta\Phi(S_1)$. Similarly, a received ray in the direction θ_{b2} that impinges first on M_1 and then on S_2 (a new point on the surface of the sub-reflectarray, see Fig. 3-10(b)) can be used to obtain the value of the phase adjustment associated to S_2 : $\Delta\Phi(S_2) = (2\pi/\lambda) \cdot \Delta PL_{2,1} - \Delta\Phi(M_1)$.

Fig. 3-10 Example of performance of the phase adjustment routine in the xz -plane: (a) transmitted ray from F_1 , and (b) received ray that goes to F_2 .

The same process can be repeated by starting with a transmitted ray from F_1 that impinges on S_2 . After N iterations, a set of points is obtained for each reflectarray with their associated phase constants. Then, the samples of the phase adjustment required on each reflectarray can be interpolated by polynomials, and the result added to the initial phase distributions obtained by rotation.

In the previous DRA design, whose phase distributions are shown in Fig. 3-8, the tilt angles that have been estimated for each reflectarray are $\theta_s = 10^\circ$ and $\theta_M = 15^\circ$. The required phase adjustment in the xz -plane related to those tilt angles is presented in Fig. 3-11. The final phase-shift distributions resulting from the application of this method are shown in Fig. 3-12. As can be seen, these distributions present a rather smooth

variation, especially in the case of the sub-reflectorarray. The most important result of this technique is that the antenna radiation patterns will present the same bifocal characteristic in the xz -plane, while maintaining quite good results for the orthogonal plane, as a consequence of the previous rotation of the phase curves.

Fig. 3-11 Phase adjustment required in the xz -plane for: (a) sub-reflectorarray and (b) main reflectarray.

Fig. 3-12 Adjusted bifocal phase-shift distributions for: (a) sub-reflectorarray and (b) main reflectarray.

3.2.5 Radiation patterns of the bifocal antenna

The results of the proposed bifocal method have been evaluated for the DRA system designed at 20 GHz to radiate two beams at $\theta_{bi} = \pm 1.5^\circ$, whose final geometry after tilting both reflectarrays is shown in Fig. 3-9(b). The DRA is composed of a 1.8-m diameter main reflectarray and a 60-cm sub-reflectorarray; the adjusted phase-shift distributions of both reflectarrays are shown in Fig. 3-12. The analysis of the DRA system has been carried out by a specific homemade software tool, which applies the modular technique described in [70]. The accuracy of this tool has been validated in previous works involving the design, manufacturing and test of a dual reflectarray

demonstrator [71]. In this case, ideal reflectarray cells that provide the required phase-shift at 20 GHz are assumed, which simplifies the analysis of the antenna. The diameter of the feed-horns to provide -12 dB illumination on the edges of the sub-reflectarray for a subtended angle of 26.8° is estimated at 65 mm. The electromagnetic field radiated by the horns is modelled by a $\cos^q(\theta)$ function, with $q = 50$.

The simulated radiation patterns for the initial DRA configuration with the two reflectarrays in parallel planes (see Fig. 3-9(a)) are shown in Fig. 3-13. The results are presented at 20 GHz (Tx frequency in Ka-band), in the elevation (xz) and azimuth orthogonal planes. Note that the azimuth plane forms 1.5° with respect to z -axis for the beam generated by F_1 , and -1.5° for the beam generated by F_2 , although they are superimposed in the same figure. As the design has been performed considering ideal phases, there are no cross-polar components of the radiated field. A gain of 49.6 dB is reached for the beam at 1.5° , and a 49.2 dB gain is attained for the beam at -1.5° . The SLL is lower than -20 dB respect to the maximum.

Fig. 3-13 Simulated radiation patterns at 20 GHz for the initial DRA system with parallel reflectarrays: (a) in the elevation plane, and (b) in the azimuth plane.

On the other hand, Fig. 3-14 shows the radiation patterns of the tilted bifocal antenna (see Fig. 3-9(b)). A gain of 48.2 dBi is reached for the beam at 16.5° , and a 48 dBi gain is attained for the beam at 13.5° . There is a small loss in gain (about 1.3 dB) with respect to the radiation patterns of the bifocal antenna with parallel reflectarrays, mainly due to the tilt of the main reflectarray, which causes a slight reduction in the effective aperture of the antenna, as well as a beam displacement from the initial directions at $\pm 1.5^\circ$ to the final pointing at $15^\circ \pm 1.5^\circ$ respect to the $\hat{z}_M$ axis of the tilted main reflectarray. Despite this fact, the beam shaping and SLL are conserved.

Fig. 3-14 Simulated radiation patterns at 20 GHz for the final DRA system, after tilting both reflectarrays (a) in the elevation plane, and (b) in the azimuth plane.

The amplitude of the incident field on both reflectarrays produced by the feeds placed at F_1 and F_2 is presented in Fig. 3-15. The illumination levels are close to -12 dB on the edges, in order to maximize the gain and radiation efficiency of the bifocal antenna. In the case of the main reflectarray illumination, it can be seen that it moves from its upper part (when the antenna is illuminated from F_1) to its bottom edge (when illumination from F_2 is considered). These results can be improved in a more detailed

design of the antenna by pointing to different zones of the sub-reflectarray with each feed, so as to obtain a more centered illumination on the main reflectarray. In a Cassegrain system (like the current one), the lower feed (F_1) should point to the lower part of the sub-reflectarray, while the upper feed (F_2) should point to its upper part. An oversized sub-reflectarray could be used for this purpose, but always keeping in mind the problem of the blockage caused by the sub-reflector.

Fig. 3-15 Amplitude (dB) of the incident field on the sub-reflectarray for (a) F_1 and (b) F_2 , and on the main reflectarray for (c) F_1 and (d) F_2 .

After the calculation of the radiation patterns for the two feeds placed at F_1 and F_2 , additional feeds are considered in order to evaluate the multi-beam performance of the antenna. Since the distance between the foci is $d = 20$ cm and the diameter of the horns is estimated at 65 mm, only two additional feeds can be placed between those at F_1 and F_2 , with a separation of 66.7 mm between the phase centers of consecutive horns. Then, two more feeds have been added in the extremes of the feed array, placed on the xz -plane, so as to generate a total of six beams. The simulated radiation patterns in the

elevation plane for the six beams are shown in Fig. 3-16 (solid lines). As can be seen, the separation between adjacent beams is around 1° , the gain varies from 47.7 dB to 48.2 dB, and SLL is lower than -21 dB for all the beams. Moreover, the beams generated by a reference reflectarray antenna in a single reflector configuration, designed to provide the same gain by using adjacent feeds with 66.7 mm separation, have been obtained and included in Fig. 3-16 (dashed lines). As can be seen, the monofocal beams present around 1.5° - 1.6° separation, and the main lobe of the extreme beams is considerably broadened with respect to the main lobe of the bifocal beams. As can be seen, the proposed bifocal method allows to obtain a similar performance in terms of gain and SLL for all the beams, and also to generate closer beams with non-overlapping feeds than in the single-focus design.

Fig. 3-16 Simulated radiation patterns at 20 GHz in the xz -plane for the bifocal dual reflectarray antenna (solid lines) and the single-focus reference antenna (dashed lines).

The performance of the bifocal antenna has been also evaluated for the beams out of the plane where the bifocal synthesis was performed (xz -plane). The pattern contours of 38 dBi, 45 dBi (which approximately corresponds to -3 dB with respect to the maximum gain) and 47.5 dBi are shown in Fig. 3-17 for the beam produced from focus F_1 ($\theta = 16.5^\circ$) and a ring of five adjacent beams with 1° separation with respect to the central one. Note that the three beams in the plane $\nu = 0$ are produced by the horns no. 1, 2 and 3, according to Fig. 3-16. As can be seen, the minimum separation between adjacent beams that can be achieved in the current design is limited by the diameter of the feed-horns. To obtain less than 1° separation, the size of the horns must be reduced (less directive horns), leading to the necessity of oversized reflectors.

Fig. 3-17 Radiation pattern contours of 38 dBi, 45 dBi and 47.5 dBi for the beams produced from focus F_1 of the bifocal antenna and a ring of five beams.

Although the results in this section are presented only at 20 GHz (Tx frequency from a satellite in Ka-band), the combination of the proposed bifocal method with the capabilities of polarization and frequency discrimination of reflectarrays, will allow to perform independent bifocal synthesis for each frequency (Tx and Rx) and/or polarization. This will require the use of appropriate reflectarray cells that will provide independent phase control at different frequencies and/or polarizations.

3.3 Considerations on the design of bifocal dual reflectarray antennas

The initial conditions of the synthesis have an important effect on the performance of the bifocal algorithm, and particularly, on the results obtained for the phase distributions and relative dimensions of the two reflectarrays. The setting of very extreme or improper initial conditions may lead to convergence problems of the algorithm, not being possible to reach a valid solution for the bifocal design of the antenna with such characteristics. For this reason, it may be useful to count on a single-focus design (reference reflector or dual reflectarray antenna) that serves to estimate the main geometrical parameters of the bifocal antenna, at least as a first approximation. Some additional aspects to be taken into account concerning the performance of the bifocal algorithm are set out below.

3.3.1 Setting of the beam spacing

The minimum separation that can be achieved between adjacent beams in a SFPB multi-beam antenna system is limited by the size of the feeds, which is fixed to provide an optimum illumination according to the F/D ratio of the antenna. To achieve a smaller beam spacing, there are two possible solutions: to reduce the size of the feeds (less directive feeds), which forces to increase the antenna size in order to maintain low spillover, or to keep using the same feeds and increase both the focal distance and the antenna diameter (thus preserving the F/D ratio of the antenna). Note that both solutions end up with an oversized antenna, in order to avoid the problem of overlapping feeds.

On the other hand, the bifocal technique allows a certain degree of control over the separation between adjacent beams for a given feed spacing by properly setting the initial parameters of the design algorithm. Regarding the performance of the bifocal antenna, it can be designed to obtain a larger beam spacing, a smaller beam spacing or a similar beam spacing to the one provided by the equivalent single-focus design. Bifocal antennas with wider beam spacing are suitable for those applications in which a large beam scanning angle is required [17]. In the case of this thesis, we are more interested in the design of bifocal antennas with a reduced separation between adjacent beams, which can be a potential alternative for current multi-beam satellite antennas in Ka-band.

To illustrate the performance of the bifocal technique for each of the three possible design cases, the bifocal algorithm has been executed maintaining some of the initial parameters of the DRA system designed in the previous sections and changing only the directions of the beams. The initial parameters of the bifocal synthesis for the three designs (Design 0, Design 1 and Design 2) are shown in Table 3-1. The samples and the interpolated curves of the phase derivative on each reflectarray can be seen in Fig. 3-18 for the three design cases, while the resulting bifocal phase functions in the xz -plane are presented in Fig. 3-19.

As can be inferred from the relation of the phase derivative samples, the design with the largest beam spacing (Design 1) leads to similar sizes for both reflectarrays, as well as to the smoothest phase variation on the main reflectarray. The first factor obliges to design centered or slightly offset geometries, where the antenna will present blockage from the sub-reflectarray (in [17], this blockage is avoided by a 90° twist of polarization on the main reflectarray).

TABLE 3-1
INITIAL PARAMETERS OF THE BIFOCAL SYNTHESIS

Parameter	Design 0	Design 1	Design 2
L_1	1 m	1 m	1 m
L_2	1.5 m	1.5 m	1.5 m
d	0.2 m	0.2 m	0.2 m
θ_{b1}	+1.5°	+3°	+0.75°
θ_{b2}	-1.5°	-3°	-0.75°

(a)

(b)

Fig. 3-18 Interpolation of the phase derivative samples obtained on the: (a) sub-reflectorarray and (b) main reflectarray.

(a)

(b)

Fig. 3-19 Bifocal phase curves obtained after the integration of the interpolated phase derivatives on the: (a) sub-reflector array and (b) main reflectarray.

The design with the smallest beam spacing (Design 2) produces the fastest phase variation on the main reflectarray, and the largest size of the main reflectarray for a fixed size of the sub-reflector array. Meanwhile, the original design (Design 0) provides an intermediate solution for the phase variation on both reflectarrays. Both designs, the one with $\theta_b = \pm 1.5^\circ$ and the one with $\theta_b = \pm 0.75^\circ$, are suitable for the deployment of dual offset antenna geometries that will minimize blockage from the sub-reflector array, allowing the antenna for operating in dual-polarization. Also, note that the shape of the bifocal phase curve obtained on the sub-reflector array changes from a concave response (Design 1, with $\theta_b = \pm 3^\circ$) to a convex one (Design 2, with $\theta_b = \pm 0.75^\circ$).

The previous effects on the phases can be associated to a shifting of the virtual focal position in the equivalent monofocal system: when the bifocal technique is applied to increase the beam spacing, the virtual focus moves away from the sub-reflectarray (in the direction of positive z -axis), which results in similar sizes for both reflectarrays and a concave phase response on the sub-reflectarray. Conversely, when the bifocal technique is used to reduce the beam spacing, the virtual focus moves towards the sub-reflectarray, producing highly offset geometries with a very large main reflectarray and a convex phase response on the sub-reflectarray.

3.3.2 Design of a Gregorian system

Regarding the bifocal designs that have been performed so far, the beam generated from F_1 has been always considered to radiate at $\theta_{b1} > 0^\circ$ (according to the reference system shown in Fig. 3-1), while the beam produced from F_2 presents $\theta_{b2} < 0^\circ$ and fulfills $\theta_{b2} = -\theta_{b1}$. However, it is also possible to swap the directions of the beams associated to the foci, so that F_1 will generate the beam radiating at $\theta_{b1} < 0^\circ$ and F_2 will produce the beam at $\theta_{b2} > 0^\circ$. In that case, the samples obtained on the sub-reflectarray after applying the bifocal ray-tracing process will grow in the direction of the negative x -axis, leading to a Gregorian instead of a Cassegrain system (see Fig. 3-20).

Fig. 3-20 Performance of the bifocal ray-tracing in the case of designing for a Gregorian system.

As in the case of Cassegrain systems, the bifocal phase curves obtained in the xz -plane can be rotated around z -axis, and then both reflectarrays can be tilted a certain angle by applying the previously described phase adjustment routine, in order to achieve smoother phase distributions. The use of a Gregorian configuration provides similar results than those obtained with a Cassegrain one, but the latter has been the selected option for performing the previous designs because it allows a larger equivalent F/D ratio, resulting in a main reflectarray with a lower number of 360° cycles.

3.3.3 Conclusions on the application of the bifocal method to offset configurations

The bifocal ray-tracing algorithm described in section 3.2.1 can be applied to a DRA system by considering already tilted reflectarrays in the xz -plane. In this case, the problem of setting the initial value of the phase derivative at the point S_1 on the sub-reflectarray may not have a unique solution, since the symmetry conditions existing in the geometry with parallel reflectarrays may not be applicable here. Several options are then possible for $\Phi'_x(S_1)$, provided that they fulfill the conditions given in eq. (3-2). Moreover, the extension of the 2D design performed in the xz -plane to a 3D solution is not trivial (this aspect will be addressed in Chapter 5). This is the main reason for starting from an axially-symmetrical geometry with parallel reflectarrays, in which the phases obtained in the xz -plane can be rotated around z -axis. However, the proposed bifocal method presents some geometrical constraints that limit its use for the design of any bifocal DRA configurations.

First, the initial positions of the foci must be symmetrical with respect to z -axis, in order to provide a focal ring after the rotation of the bifocal phase curves. Note that the relative positions of the foci with respect to the sub-reflectarray are conserved in the subsequent process of tilting the reflectarrays. Furthermore, if the reflectarrays are tilted a large angle with respect to their initial positions in parallel planes, or if the tilting angle of one reflectarray is much greater than that of the other, the resulting bifocal antenna presents an important degradation of its radiation patterns with respect to those obtained with the reflectarrays in parallel planes.

Consequently, the proposed bifocal method can be suitable for the design of both centered and offset DRA configurations with parallel reflectarrays, and also Cassegrain configurations which require a small adjustment in the tilting of both reflectarrays, as in

the case shown in Fig. 3-9. On the other hand, this method is not recommended for highly offset geometries, such as the compact-range DRA configuration in [71], which demand the implementation of a more general bifocal algorithm (as will be shown in Chapter 5).

3.4 Preliminary design of bifocal dual reflectarray configurations for multi-beam satellite antennas in Ka-band

The separation between adjacent beams required for the current multi-spot satellite applications in Ka-band is so small (typically 0.56°) that it would not be possible to produce all the beams with a single reflector antenna, as it would require overlapping feeds (unless using a highly-oversized shaped reflector, as in [90]). Therefore, four reflector antennas are commonly used to generate all the beams of the four colour coverage scenario (one reflector per colour), both in transmission (Tx) and reception (Rx), using a SFPB architecture [82].

Reflectarray antennas are a potential alternative for this kind of multi-spot applications with frequency and polarization reuse, but firstly it is necessary to evaluate their performance for the generation of multiple beams and their capability to provide closely-spaced beams. For this purpose, the bifocal technique has been applied to the design of multi-beam dual reflectarray antennas in Ka-band, considering two different approaches: generation of adjacent beams with a smaller separation than in a single-focus antenna, and improvement of the antenna performance for the edge beams. Although the results presented in this section are only preliminary, they are useful to get an idea about the performance of the bifocal technique in each of these two cases.

3.4.1 Generation of adjacent beams

The bifocal technique has been applied to a Cassegrain dual reflectarray configuration in order to obtain beam spacing compression, meaning that the separation between beams generated with adjacent feeds is so small that it could not be achieved with a single-focus antenna, as it would require overlapping feeds. The beam compression ratio (BCR) can be defined as the beam spacing achieved applying the bifocal technique divided by the beam spacing achieved without the bifocal technique, considering the same feed spacing. In this case, the bifocal antenna will provide 0.56° of

beam spacing, which is the required value for current multi-beam applications in Ka-band. This separation would traditionally require 50% feed overlap in a single-focus antenna, so the BCR is close to 2.

The geometry of the designed Cassegrain dual reflectarray antenna is shown in Fig. 3-21. The diameters of the sub- and main reflectarrays are $D_S = 80$ cm and $D_M = 1.9$ m, respectively, in order to provide around 50 dBi of gain at 20 GHz. A linear array of seven non-overlapping 54 mm feeds has been considered, where the phase centers of the second and the sixth feeds are placed at F_1 and F_2 (the foci of the bifocal antenna), respectively. Therefore, the distance between foci is $d = 21.6$ cm. The beam directions associated to the foci are $\theta_{b1} = 1.12^\circ$ and $\theta_{b2} = -1.12^\circ$, in order to obtain 0.56° of final separation between the beams generated by adjacent feeds. The other parameters of the bifocal synthesis are: $L_1 = 0.9$ m and $L_2 = 2.4$ m.

Fig. 3-21 Geometry of the bifocal dual reflectarray antenna to provide 0.56° of beam spacing.

A realistic model of a Ka-band feed-horn that has been characterized by Astrium [82] is considered for this study (see Fig. 3-22). The feed-horn presents a diameter of 54 mm and produces a taper level of -12 dB at 20 GHz on the sub-reflectorarray edges when illuminating with 36° subtended angle. For simulation purposes, the electromagnetic field radiated by the feed-horn at 20 GHz has been modeled by a $\cos^q(\theta)$ function with a q -factor equal to 28.

Fig. 3-22 Engineering model of a user/gateway feed chain [82].

The high value of the BCR (≈ 2) affects significantly the performance of the bifocal algorithm. First, a lower number of points is obtained in the xz -plane after executing the bifocal ray-tracing routine, which difficulties the subsequent interpolation of the phase derivative samples. Second, the bifocal phase curves of both reflectarrays present a steeper slope (although it has not been applied to the current design, both reflectarrays can be tilted to reduce the number of 360° cycles in their phase distributions, as explained in section 3.2.4). Finally, there is a different spatial relationship between the points obtained on both reflectarrays in the same iteration of the algorithm.

The simulated radiation patterns at 20 GHz in the principal planes for the two beams generated by the foci of the bifocal antenna have been computed assuming ideal reflectarray elements and are shown in Fig. 3-23. A gain close to 46.2 dBi is achieved for both beams, while SLL is around -16 dB with respect to the maximum. The two beams point in the specified directions ($\pm 1.12^\circ$), so a final separation of 0.56° is obtained between the beams generated by contiguous feeds placed in the xz -plane, as can be seen in Fig. 3-24.

The main characteristics of the beams (gain, SLL and beamwidth) produced by the bifocal DRA system are presented in Table 3-2. Despite achieving the required degree of beam spacing compression, the bifocal antenna presents a serious efficiency problem that can be noticed in the reduced gain of the beams (between 45.5 and 46.8 dBi), which is lower than the gain expected for a 1.9 m main aperture (it should be between 50 and 51 dBi).

Fig. 3-23 Simulated radiation patterns at 20 GHz for the two beams generated by the foci of the bifocal antenna: (a) in the xz -plane (elevation), (b) in the orthogonal plane in the direction of the beam (azimuth).

Fig. 3-24 Simulated radiation patterns at 20 GHz in the xz -plane for the bifocal antenna to provide 0.56° of beam spacing.

TABLE 3-2
MAIN CHARACTERISTICS OF THE BEAMS ($\text{BCR} = 2$)

No. beam	Beam direction (°)	Gain (dBi)	Beamwidth at 43 dBi (°)	SLL
1	-1.68	45.63	0.613	17.19
2	-1.12	46.12	0.612	17.39
3	-0.56	46.65	0.609	17.94
4	0	46.78	0.606	18.15
5	0.56	46.62	0.594	17.07
6	1.12	46.14	0.565	15.79
7	1.68	45.51	0.537	14.88

To better understand the cause of the efficiency problem, the amplitude distributions of the incident field over the symmetry axis of both reflectarrays produced by the feeds placed at the foci of the bifocal antenna, are shown in Fig. 3-25. The illumination levels on the edges of the sub-reflectarray are equal or lower than -12 dB, since a proper illumination has been ensured by using the 54-mm feeds. However, the illumination on the edges of the main reflectarray is very high, thus leading to higher spillover and lower radiation efficiency for the bifocal antenna.

Fig. 3-25 Amplitude (dB) of the incident field on the two reflectarrays produced by F_1 and F_2 .

Introducing some modifications in the initial conditions of the bifocal synthesis or changing the antenna configuration (for example, using a Gregorian system) does not mitigate this problem, which is caused by the high degree of beam spacing compression ($\text{BCR} = 2$) imposed in the design of the bifocal antenna. The only possible solution to reduce spillover and reach around 50 dBi gain would be to increase the size of the main

reflectarray, but this implies using a very large reflectarray, around 3.8 m in diameter (which is twice the size that was initially considered).

3.4.2 *Improvement of the extreme beams*

The bifocal technique has been used to correct beam aberration and improve the multi-beam performance of the dual reflectarray antenna, without reducing the beam spacing. In this case, the bifocal antenna will provide a separation of 1.12° between adjacent beams, which is twice the separation provided in the previous section. This separation can be also achieved with a single-focus antenna without using overlapping feeds, so the BCR value is equal to 1.

The bifocal algorithm has been applied to the compact-range DRA geometry shown in Fig. 3-26. The diameters of the two reflectarrays are $D_M = 1.9$ m and $D_S = 1.4$ m. An array of eleven non-overlapping feeds has been used to illuminate the antenna, so that the phase centers of the fourth and the eighth feeds are placed at F_1 and F_2 , respectively. The feed-horns present the same characteristics that those used in the previous section (54 mm in diameter and simulated with $q = 28$), so distance between foci is $d = 21.6$ cm. The beam directions associated to the foci, $\theta_{b1} = 28.24^\circ$ and $\theta_{b2} = 23.76^\circ$, have been selected to avoid blockage from the sub-reflectarray and provide 1.12° of spacing between beams generated by adjacent feeds. Due to the large offset of the antenna configuration, the design has been performed only in the xz -plane, according to what was exposed in section 3.3.3.

Fig. 3-26 Geometry of the bifocal dual reflectarray antenna to provide 1.12° of beam spacing.

The simulated radiation patterns at 20 GHz in the xz -plane for the two beams produced by the foci of the bifocal antenna are shown in Fig. 3-27. The gain of the beams has been estimated based on the beamwidth obtained at -3 dB in the xz -plane, and assuming that the design of the bifocal antenna can be extended from 2D to 3D in order to achieve the same beamwidth in the two orthogonal planes. Under these conditions, a gain close to 51 dBi is obtained for both beams. Moreover, SLL is around -22 dB with respect to the maximum gain and the 3-dB beamwidth is 0.58° . The new requirement of beam spacing (associated to $\text{BCR} = 1$) produces reasonable illumination tapers on both reflectarrays, as can be seen in Fig. 3-28, and the radiation efficiency is increased with respect to the case with $\text{BCR} = 2$. The compact range geometry forces to use a larger sub-reflectarray, although proper illumination is obtained for all the feeds.

Fig. 3-27 Simulated radiation patterns at 20 GHz in the xz -plane for the two beams generated by the foci of the bifocal antenna.

Fig. 3-28 Amplitude (dB) of the incident field on the two reflectarrays produced by F_1 and F_2 .

The bifocal dual reflectarray antenna has been compared with a reference antenna to analyze if the multi-beam performance has been improved by the application of the bifocal technique. An offset single-focus parabolic reflector with a 1.9 m diameter and $F/D = 1.34$, equivalent to the compact-range geometry shown in Fig. 3-26, has been designed and used as reference configuration. The comparison of the beams generated at 20 GHz by the bifocal antenna (solid lines) and those produced by the reference reflector (dashed lines) is shown in Fig. 3-29, considering a linear array of eleven feed-horns, each with 54 mm diameter, to generate eleven beams separated 1.12° .

Fig. 3-29 Performance of the bifocal dual reflectarray antenna (solid lines) in comparison with the single focused reference reflector (dashed lines). The directions of the beams are indicated as the variation in theta ($\Delta\theta$) respect to the direction of the central beam ($\theta = 26^\circ$).

TABLE 3-3
MAIN CHARACTERISTICS OF THE BEAMS (BCR = 1)

No. beam	Gain (dBi)	Beamwidth at 47 dBi ($^\circ$)	C/I (dB)	$\Delta C/I$ (dB)
1	51.28	0.645	21.45	-
2	51.35	0.657	18.05	+5.05
3	51.24	0.653	19.50	+5.50
4	51.19	0.656	20.36	+3.79
5	51.01	0.654	20.31	+3.01
6	50.92	0.655	20.17	-0.15
7	50.81	0.656	20.29	+1.97
8	50.60	0.653	20.62	+3.87
9	50.48	0.654	20.30	+5.22
10	50.25	0.648	19.83	+6.16
11	50.05	0.642	19.61	-

The main characteristics of the beams (gain, beamwidth, single-entry C/I and improvement in the single-entry C/I) produced by the bifocal antenna are summarized in Table 3-3. The single-entry C/I (where the interference is produced by the radiation of the adjacent beams) has been calculated within a 0.65° beamwidth, which approximately corresponds to a 47 dBi gain, and varies between 18.05 dB and 20.62 dB for the eleven beams. The lowest C/I values are associated to the interferences produced by the extreme beams, which are the most broadened beams. As can be seen in Fig. 3-29, the bifocal technique provides a better shaping of the main lobe and lower SLL for the extreme beams, which allows for a significant improvement in C/I with respect to the reference single-focused antenna (more than 5 dB). For the rest of the beams, the bifocal antenna presents around 2-3 dB of improvement in C/I, except for the central beam, which presents a slightly worse performance than in the single-focus case. Note that the improvement in C/I for the beams no. 1 and no. 11 cannot be calculated in a precise way, since the higher interference would be produced by the adjacent beams radiating at -6.72° and 6.72° , which have not being considered in the simulations.

3.4.3 Conclusions

The bifocal technique has been applied to different dual reflectarray configurations for the design of multi-beam satellite antennas in Ka-band, in order to evaluate its performance for the generation of adjacent beams with 0.56° of spacing and improving the results obtained for the extreme beams. With respect to the first objective, the bifocal antenna is able to provide a high degree of beam spacing compression (by a factor of 2) with respect to conventional single-focus antennas, so as to generate adjacent beams with 0.56° separation by using contiguous and non-overlapping feeds. However, the radiation efficiency of the bifocal antenna is not as high as it should be, due to the high spillover on the main reflectarray (which should be significantly oversized to overcome this problem). On the other hand, when the bifocal technique is applied to obtain the same beam spacing than in the equivalent single-focus design, the bifocal antenna provides better results for the extreme beams (in terms of SLL, C/I and beam shaping) than those of the monofocal case, while presenting similar values of gain and radiation efficiency.

After checking the results shown in the previous sections, it can be concluded that the most interesting option for the design of a multi-beam bifocal DRA would be

probably a combination of the two approaches. The bifocal technique can be applied to provide a reduced beam spacing (a high degree of beam spacing compression is not suitable, in order to maintain good radiation efficiency for the bifocal antenna), at the same time as correcting beam aberration and improving the performance of the extreme beams with respect to the equivalent single-focus antenna. Examples of this way of proceeding will be shown in Chapters 5 and 6.

3.5 Conclusions

A bifocal technique has been proposed for dual reflectarrays antennas. The problem of the 3D design is solved by starting from an axially-symmetrical geometry with parallel reflectarrays, in which a 2D GO algorithm is applied. The bifocal phase curves obtained in the offset plane can be rotated around the symmetry axis, and then, both centered and offset configurations are possible by selecting specific parts of these revolution surfaces. Offset configurations are preferable, as they allow to reduce blockage from the feeds and the sub-reflectarray, but the phase distributions present a high number of 360° cycles in this case, as a consequence of the large size of both reflectarrays and the initial conditions of the synthesis. For this reason, both reflectarrays can be tilted a certain angle, at the same time as their phases are adjusted to compensate the tilting and maintain the bifocal characteristic of the original design (the phase adjustment will provide smoother phase distributions on both reflectarrays).

A preliminary study on the capabilities of the bifocal technique to reduce beam spacing and improve the performance of the edge beams for multi-beam satellite antennas in Ka-band has been carried out. The results show that the bifocal technique allows to reduce beam spacing by a factor of 2 with respect to the equivalent single-focus antenna (and without using overlapping feeds), but at the cost of a lower radiation efficiency. The main reflectarray has to be significantly oversized (up to 4 m in diameter) to overcome this problem. On the other hand, the bifocal technique can be applied to correct beam aberration, providing the same beam spacing than in the single-focus case. As a result, a better performance is obtained for the extreme beams, with satisfactory results for the gain and radiation efficiency of the antenna.

Finally, note that the bifocal antennas have been designed to operate at 20 GHz (Tx frequency from a satellite in Ka-band), but a similar bifocal process can be carried out

to obtain the required phases on both reflectarrays at 30 GHz (Rx frequency in Ka-band), in order to generate the beams in reception. Then, appropriate reflectarray cells that will provide independent phase control at Tx and Rx frequencies can be used to implement the phase distributions on both reflectarrays, allowing for the design of transmit and receive multi-beam satellite antennas in Ka-band.

Chapter 4

Bifocal technique applied to dual transmitarray antennas

4.1 Introduction

The bifocal technique described in Chapter 3 can be applied to the design of dual transmitarray antennas (DTA), in a similar way to the case of dual reflectarrays. The use of transmitarrays provides some interesting advantages, as it eliminates blockage from the feeds and allows for the use of centered and rotationally-symmetrical geometries, thus simplifying the design process of the bifocal antenna. Besides these similarities with dielectric lenses, transmitarrays offer the same flexibility as reflectarrays when implementing the capabilities of the unit cells (flat panels, manufactured technology of printed circuits, dual polarization [45], [117], possibility of reconfiguration of the beam by adding controllable phase shifters in the transmitarray cells [47], [118], [119], etc.). The design of transmitarray cells requires finding a cell topology that provides at least 360° of phase range in a required bandwidth, at the same time as presenting full transmission and zero reflection within the operating band. In the literature, two main approaches have been reported to design transmitarray antennas (see Fig. 4-1): multiple stacked frequency selective surfaces (FSS) [43]-[45], [120]-[124] and transmitter-receiver antenna [42], [47], [125], [126].

The design process of a passive transmitarray based on stacked FSSs focuses on the synthesis and subsequent design of one filter for each phase state, assuming that the phase of the transmitted wave is discretized in a fix number of states, for example 8 states for 3-bit quantization, as in references [43], [44]. There are two main strategies to achieve the phase-shift. The first one consist on making that each filter provides the

same electrical performance, but centered at a different frequency, thus producing the required phase shift [120]-[122]. The other strategy consists on obtaining different slopes of the amplitude curves and bandwidths for each phase state, where all the states are centered at the same frequency [43]. The latter strategy usually requires larger ranges of variation of the equivalent capacities and inductances than the other one, so that it usually requires different types of metallizations to cover the entire range of 360° [43]. For both strategies, the approach based on filter theory exhibits electrical limitations, since it involves the use of canonical structures (Chebyshev, Elliptical, etc.).

Fig. 4-1 Example of the two approaches commonly used to design a transmitarray cell: (a) multiple stacked FSSs [121] and (b) transmitter-receiver antenna [40].

The theoretical limitations of canonical structures imply high order filters to achieve a phase-range of at least 360° . In [120], it was demonstrated that 4 or more layers are necessary to reach a phase range of 360° assuming an admissible level of losses of 1 dB (antenna efficiency of 56 %), whereas that the number of layers gets down to 3 if the admissible losses increases up to 3 dB (antenna efficiency of 35 %). These theoretical limitations are coupled with other physical limitations. By one hand, the values of the capacitances, inductances and topology, which result from the canonical synthesis of the filter, must be physically provided by the cell. Otherwise, the performance of the filter in terms of bandwidth and matching worsens, resulting in reduction of bandwidth and efficiency at antenna level. On the other hand, the synthesis of two canonical filters that exhibit similar bandwidth and ripple, but each one centred at a different frequency, implies different line lengths for each filter, so that each phase state of the transmitarray must have an appropriate thickness. However, the cells that compose the transmitarray antenna must present the same thickness, so a variable ripple of the bandpass must be assumed, resulting in a deterioration of the filtering structure and antenna performance.

Several works have been reported on increasing the performance of transmitarrays using the approach based on FSSs. Most of them are focused on improving both the bandwidth and the phase range using a relatively short number of layers, so that the efficiencies are relatively far of those required in space applications. In [121], a transmitarray antenna composed by four layers based on dual-resonant double square-rings elements is shown, which exhibits 1-dB gain bandwidth of 7.5% and antenna efficiency of 47% at 30 GHz. In [122], the use of polarization insensitive filters using thin thicknesses ($<0.6\lambda$) provides a 1-dB gain bandwidth of 10.2%, although at the expense of a reduced efficiency (34%) and the use of an additional layer (5 layers). The strategy of considering several cell topologies in the same array was proposed in [43] to improve the bandwidth, each of which designed to cover a certain phase interval of the complete 360° ; in this case, these topologies allow finding a large range of the required circuital variables after the synthesis, thus achieving 12.5% bandwidth of (-1 dB criteria) and 45% efficiency at 20 GHz using four layers.

Concerning the antenna efficiency, in [123] the efforts were made on reducing the losses of the cells (thus increasing the efficiency up to reach 55%), although at expense of assuming a small bandwidth (7.4%, -1 dB criterion). A trade-off between efficiency and bandwidth was also presented in [124], where both features are improved by using four layers and double square loop elements on dielectric materials, which exhibits 1-dB gain bandwidth of 11.7% and efficiency of around 48%. In this case, the improvement and the trade-off that includes the efficiency are achieved by optimizing the phase distribution on the aperture.

Other important objective for the developing of transmitarray antennas for space communications has been recently addressed in [45], where a transmitarray antenna exhibiting two independent beams for each linear polarization has been demonstrated. The transmitarray is composed by 3 stacked FSSs in X-band (see Fig. 4-2(a)), assuming canonical rectangular shapes for the patches, which is essential to ensure independence of the polarisations and low cross polar levels. However, stacked patches are able to provide only limited values of the circuital equivalent parameters, which implies limitations in terms of phase-range (210° , 1 dB of tolerable ripple), bandwidth (7%, -1 dB criterion), efficiency (which is especially small due to the phase quantization) or SLL. The difficulty of achieving independent dual linear polarization using FSSs is due to the fact that canonical shapes must be used, which is contrary to what is usually

needed to increase the phase-range or the bandwidth. In [43], the physical complexity of the elements used to improve the electrical performance of transmitarrays does not allow to obtain independent phase-shifting for each polarization.

The second approach to design transmitarray antennas consists on interconnecting the receiver and transmitter patches using slots on a common ground plane or delay lines. In [42], it was shown that the use of slots implies phase-ranges limited to 180° and a narrow bandwidth, being necessary the use of delay lines to overcome these limitations. Some references have also been focused on improving the electrical performance of transmitarrays using the antenna approach [117], [125]-[126], which are able to reach similar specifications than those obtained using FSSs with some additional advantages. These structures provide more flexibility to control the amplitude of the transmitted field or to include devices to electronically control the phase [125]. This flexibility allows for obtaining electrical features such as circular polarization [126], dual polarization (see Fig. 4-2(b)) [117] or dual band behaviour [46] more easily from the point of view of the design. However, the resulting antennas are usually more voluminous and heavier than the transmitarrays based on FSS, imply a more complex fabrication and can exhibit worse efficiencies due to the losses of the lines and the internal microwave circuitry.

Fig. 4-2 Two different transmitarray cells to achieve operation in dual polarization: (a) based on multiple stacked FSSs [45] and (b) based on transmitter-receiver concept and the use of PIN diodes [117].

In this chapter, ideal transmitarray elements will be considered, which present 360° of phase range to provide the required phase-shift, while presenting at the same time zero reflection and insertion losses within the operating band.

4.2 Bifocal design procedure for dual transmitarray antennas

Figure 4-3 shows the geometry of the dual transmitarray antenna and its main parameters, which must be fixed before starting the bifocal design procedure: directions of the radiated beams (θ_{b1} and θ_{b2}), distance between foci (d), distance between the foci and the first transmitarray (S_A) and distance between the two transmitarrays (S_B). As in the bifocal design of dual reflectarrays, a symmetrical arrangement of the foci along the x -axis is considered: the coordinates of the focal points are $(x_{F1}, y_{F1}, z_{F1}) = (-d/2, 0, 0)$ and $(x_{F2}, y_{F2}, z_{F2}) = (d/2, 0, 0)$. The beam directions are also symmetric with respect to the horizontal axis: $\theta_{b1} = -\theta_{b2}$.

Fig. 4-3 Geometry of the dual transmitarray antenna and example of performance of the bifocal ray tracing routine in the xz -plane.

Each transmitarray cell will introduce a certain phase shift between elements such that the incident wave coming from the feed with an angle θ_i will produce a transmitted wave with an output angle θ_o . As in the case of reflectarrays, it can be shown that the incidence and output angles will fulfill the following relation with the phase derivative along the transmitarray profile ($\partial\Phi/\partial x$):

$$\Phi'_x = \frac{\partial\Phi}{\partial x} = \frac{2\pi}{\lambda} \cdot (\sin \theta_i - \sin \theta_o) \quad (4-1)$$

An iterative ray-tracing algorithm that applies eq. (4-1) for both transmitarrays and alternates transmitted and received rays in the same way that is described in section

3.2.1 will provide a set of points (P_i and Q_i) that characterize each transmitarray, with the associated values of the phase derivative at those points. Due to the symmetry of the antenna configuration, the required phases on both transmitarrays will be even functions with respect to x , so their phase derivative will be zero at $x = 0$. Therefore, the first iteration of the bifocal algorithm will start with a transmitted ray from F_1 that impinges on the first transmitarray at $P_1 = (0, 0, S_A)$, with $\Phi'_x(P_1) = 0$, as shown in Fig. 4-3. Like in the design of dual reflectarray configurations, a second execution of the ray-tracing routine can be performed in order to double the number of points and improve the accuracy of the interpolation, starting with a received ray in the direction θ_{b2} that impinges on the second transmitarray at $Q_1' = (0, 0, S_A + S_B)$, also with $\Phi'_x(Q_1') = 0$.

The phase derivative samples obtained on both transmitarrays will be interpolated by polynomials, and then integrated to obtain the bifocal phase functions in the xz -plane. These phases will be rotated around z -axis to obtain a complete solution for the phase distributions on both transmitarrays. Note that the size of the first transmitarray for a proper illumination on the second transmitarray (or main transmitarray) can be estimated based on the relationship between the points obtained in the xz -plane for both transmitarrays, after applying the ray-tracing algorithm.

A bifocal dual transmitarray antenna, equivalent to the DRA system shown in section 3.2, has been designed at 20 GHz (transmission frequency in Ka-band) with the following parameters: $d = 20$ cm, $S_A = 1$ m, $S_B = 1.5$ m, $\theta_{b1} = 1.5^\circ$, and $\theta_{b2} = -1.5^\circ$. The geometry of the dual transmitarray configuration can be seen in Fig. 4-4, where the diameters of the transmitarrays are $D_1 = 0.55$ m (for the first transmitarray) and $D_2 = 1.8$ m (for the second transmitarray).

The bifocal phase curves obtained in the xz -plane for each transmitarray can be seen in Fig. 4-5. The behavior of the phases is very similar to the case of the dual reflectarray antenna shown in section 3.2: a convex response on the sub-transmitarray and a concave one on the main transmitarray. The points from $x = 0$ m to $x = 0.275$ m have been chosen for constituting the first reflectarray (0.55 m in diameter), and the associated points from $x = 0$ m to $x = 0.9$ m, for the main reflectarray (1.8 m in diameter). The bifocal phase distributions (in degrees) that must be implemented on each transmitarray are shown in Fig. 4-6. These distributions present a lower number of 360° cycles than in the equivalent dual reflectarray antenna with parallel reflectarrays.

Fig. 4-4 Geometry of the bifocal dual transmitarray antenna.

Fig. 4-5 Phases curves obtained with the bifocal technique in the xz-plane: (a) for the first transmitarray and (b) for the second transmitarray.

Fig. 4-6 Bifocal phase-shift distributions (in degrees) obtained for: (a) the first transmitarray and (b) the second transmitarray.

The simulated radiations patterns of the bifocal dual transmitarray antenna at 20 GHz in the elevation and azimuth orthogonal planes are shown in Fig. 4-7. As can be seen, a gain close to 50 dBi is reached for both beams, which present a 3-dB beamwidth of 0.57° and SLL lower than -25 dB respect to the maximum gain. The simulations have

been performed considering ideal transmitarray elements and a $\cos^q(\theta)$ function with $q = 50$ to model the electromagnetic field radiated by the feed-horns, which provides around -12 dB illumination on the edges of the first transmitarray (according to this illumination, the diameter of the horns is estimated at around 65 mm). Since the design has been performed considering ideal phases, there are no cross-polar components of the radiated field.

Fig. 4-7 Simulated radiation patterns for the dual transmitarray antenna: (a) in the elevation plane, (b) in the azimuth plane

The amplitude distributions of the incident field on both transmitarrays produced by the feeds placed at F_1 and F_2 is presented in Fig. 4-8. The illumination levels are close to -12 dB on the edges, thus maximizing the antenna gain and radiation efficiency. Finally, the simulated radiation patterns of the bifocal antenna at 20 GHz have been calculated for a linear array of six horns with 66.7 mm distance between their phase centers (where the second and the fifth horns are placed at F_1 and F_2 , respectively) and are shown in Fig. 4-9. As can be seen, the maximum gain is close to 50 dBi, and SLL is around -25

dB for all the beams. A separation of 1° is obtained between adjacent beams generated by contiguous feeds.

Fig. 4-8 Amplitude (dB) of the incident field on the first transmitarray produced by (a) F_1 and (b) F_2 , and on the main transmitarray produced by (c) F_1 and (d) F_2 .

Fig. 4-9 Simulated radiation patterns at 20 GHz in the XZ-plane for the bifocal antenna that provides 1° separation between adjacent beams.

4.3 Considerations on the design of bifocal dual transmitarray antennas

The bifocal dual transmitarray antenna presents some advantages with respect to the equivalent dual reflectarray. First, it avoids blockage from the feeds or the first transmitting structure. This makes it possible the design of centered and rotationally symmetrical configurations, where the required bifocal phase distributions present a lower number of 360° cycles. In the design with parallel reflectarrays, the choice of an offset configuration that minimizes blockage ends up with a very large number of 360° cycles in the phases, since the slope of the phase curves in the xz -plane becomes greater as they move away from $x = 0$. This reduces the potential bandwidth of the reflectarray and forces a modification of the antenna geometry by tilting both reflectarrays and correcting their phase distributions, which degrades the performance of the focal ring obtained after rotating the phases around z -axis (see section 3.2.3). On the other hand, the dual transmitarray antenna does not require any further adjustment on the bifocal phases obtained by rotation, allowing to work with a focal ring that contains F_1 and F_2 . This fact improves the antenna performance for the generation of multiple beams using a cluster of feeds, as it extends the focal region out of the xz -plane. Finally, the radiation patterns are less sensitive to surface deformations in the transmitarray than in the corresponding reflectarray configuration.

One of the problems of the dual transmitarray design consists on the large spacing of the resulting surfaces along the horizontal axis, as can be seen in Fig. 4-4. The use of reflectarrays allows for a more compact configuration, such as a Cassegrain or Gregorian reflector system, while the dual transmitarray requires more space to place the cluster of feeds and the two transmitarrays. Regarding the latter, there are two design alternatives that may be interesting to study: reducing the value of S_A distance, so as to integrate the feeds and the first transmitarray into the same sub-system; and using a small value of S_B distance, so that both transmitarrays can be held by the same supporting structure (this option resembles a dielectric lens). Two bifocal designs equivalent to the one previously shown in the previous section (Design 0) have been carried out at 20 GHz, preserving the F/D ratio of the antenna configuration: one with a small value of S_A (Design 1, see Fig. 4-10(a)) and the other with a short S_B distance (Design 2, see Fig. 4-10(b)). The initial parameters of the bifocal synthesis for the three designs are shown in Table 4-1. Note that the size of the first transmitarray increases as

it is placed closer to the second transmitarray: in Design 1, it is only 26 cm in diameter, while in Design 2 its diameter is 1.1 m. The comparison of the bifocal phase curves in the xz -plane obtained for each transmitarray in the three designs is shown in Fig. 4-11. As can be seen, the phase exhibits a concave response on the first transmitarray and a convex one on the main transmitarray.

TABLE 4-1
INITIAL PARAMETERS OF THE BIFOCAL SYNTHESIS

Parameter	Design 0	Design 1	Design 2
S_A	1 m	0.3 m	2.3 m
S_B	1.5 m	2.2 m	0.2 m
d	0.2 m	0.2 m	0.2 m
θ_{b1}	+1.5°	+1.5°	+1.5°
θ_{b2}	-1.5°	-1.5°	-1.5°

Fig. 4-10 Different design options for the bifocal dual transmitarray: (a) with a short S_A distance and (b) with a short S_B distance.

Fig. 4-11 Bifocal phase curves required for the three design configurations: (a) on the first transmitarray and (b) on the second transmitarray.

Regarding the performance of the ray-tracing algorithm, Design 2 provides the largest number of points to characterize each transmitarray, which facilitates an accurate interpolation of the phase derivative samples. However, it implies the use of a larger first transmitarray, and the bifocal phase curves present a steeper slope than in the other cases, which means more 360° cycles in the phase distributions. To obtain a smoother phase variation, a possible solution would be to increase the value of S_A for the same beam directions (larger F/D ratio).

On the other hand, Design 1 requires a smaller first transmitarray, which means less weight and lower fabrication costs. It presents a reasonable number of 360° cycles in the phases, but at the expense of obtaining a lower number of points in the bifocal

synthesis. This may lead to convergence problems if the initial parameters of the bifocal algorithm are modified when seeking different configurations to satisfy geometrical constraints or stringent antenna specifications, as will be shown in the next section.

4.4 Bifocal dual transmitarray antenna to reduce beam spacing

The previous results show the advantages of the proposed bifocal method for the design of multi-beam dual transmitarray antennas, but a smaller separation between adjacent beams (around 0.56°) is required in order to fulfill the stringent requirements of current multi-spot satellite antennas in Ka-band. The bifocal technique can be applied to reduce the spacing between beams generated with adjacent feeds in a way that would not be possible in a single-focus antenna, as the corresponding beam spacing would require overlapping feeds in that case.

The two design alternatives described in section 4.3 have been considered for the design of such an antenna. The first option (small S_A) was discarded after doing several tests with the bifocal algorithm, as it led to convergence problems with only a few points corresponding to a very small first transmitarray and a very large second transmitarray. Conversely, the second design alternative (small S_B) seems to be more feasible, so it has been the option chosen to start this study.

The bifocal technique has been applied to a dual transmitarray configuration in order to obtain 0.56° of separation between adjacent beams, considering a linear array of six horns of 54 mm diameter and the same characteristics than those used in section 3.4 for the design of multi-beam dual reflectarray antennas in Ka-band. In this case, the phase centers of the first and the sixth horns of the array are located at the foci of the bifocal antenna. The bifocal design algorithm has been executed with the initial parameters that are indicated in Table 4-2.

TABLE 4-2
INITIAL PARAMETERS OF THE BIFOCAL SYNTHESIS

Parameter	Value
S_A	3 m
S_B	0.3 m
d	0.27 m
θ_{b1}	$+1.4^\circ$
θ_{b2}	-1.4°

The geometry of the antenna is shown in Fig. 4-12. The value of D_2 has been set to 1.8 m in order to provide around 50 dBi gain at 20 GHz. The size of the first transmitarray is $D_1 = 1$ m and has been fixed (as a first approximation) to produce around -12 dB edge illumination on the main transmitarray. The phase distributions obtained for this bifocal antenna are shown in Fig. 4-13. As can be seen, the requirement of beam compression, combined with the small value of S_B , produces a faster variation in the phases of both transmitarrays (when compared with those shown in Fig. 4-6, for example). Note that the phase distributions are opposite in each transmitarray, the phase decreases from the center to the edge in the first transmitarray and the other way around in the second transmitarray.

Fig. 4-12 Geometry of the dual transmitarray antenna to achieve beam compression.

Fig. 4-13 Bifocal phase-shift distributions (in degrees) obtained for: (a) the first transmitarray and (b) the main transmitarray.

The simulated radiation patterns at 20 GHz for the two beams produced by the foci of the bifocal antenna, considering ideal phases in both transmitarrays and a $\cos^q(\theta)$

distribution with $q = 28$ for the horn model, are shown in Fig. 4-14. As can be seen, the required beam compression is achieved, but at the cost of a lower radiation efficiency of the bifocal antenna. Around 46.2 dBi gain is achieved with a 1.8 m main transmitarray, whereas in the patterns shown in Fig. 4-7 (with no beam compression) a gain close to 50 dBi was obtained for the same size of the main transmitarray. Apart from that, SLL is lower than -20 dB respect to the maximum, and both beams present well-shaped main lobes in the principal planes, as a result of rotating the bifocal phases around z -axis.

Fig. 4-14 Simulated radiation patterns for the bifocal antenna: (a) in the elevation plane, (b) in the azimuth plane.

The simulated radiation patterns at 20 GHz of the bifocal antenna, considering an array of six non-overlapping horns (54 mm diameter) to produce six beams with 0.56° of spacing in the xz -plane, are shown in Fig. 4-15 (solid lines). Also, the radiation patterns of the beams generated by an equivalent single-focus antenna, with the same aperture size (1.8 m) and the same feed separation (54 mm), have been calculated and included in Fig. 4-15 (dashed lines). As can be seen, the bifocal antenna provides the

required 0.56° separation between adjacent beams, while the beams from the equivalent single-focus antenna present around 1.1° separation. This means that the bifocal antenna is able to compress the beam spacing by a factor of 2 with respect to the single-focus case, which would require 50% feed overlap to provide the same beam separation. However, the gain of the beams is 4.25 dB lower than in the single-focus design, despite considering the same aperture size.

Fig. 4-15 Simulated radiation patterns at 20 GHz in the xz -plane for the bifocal antenna to provide 0.56° separation between beams (continuous lines) compared with the patterns for a monofocal equivalent antenna (dashed lines).

This reduction in the radiation efficiency is due to the high spillover on the first transmitarray, whose size is smaller than required for an optimum illumination from the feeds. The amplitude of the incident field over the symmetry axis of both transmitarrays produced by the horns placed at F_1 and F_2 can be seen in Fig. 4-16. The solution to the spillover problem is not trivial. If the first transmitarray is oversized to obtain -12 dB taper, the illumination on the main transmitarray increases significantly and reaches around -6 dB on the edges. Then, the main transmitarray should be also oversized, but this would result in a diameter close to 4 m, which is twice the size that was initially considered to provide 50 dBi gain at 20 GHz. Different dual transmitarray configurations have been studied in order to obtain a high degree of beam spacing compression, modifying the initial parameters of the bifocal synthesis, but the same illumination problem remains.

Fig. 4-16 Amplitude (dB) of the incident field on the two transmitarrays produced by F_1 and F_2 .

As an example, the bifocal algorithm has been executed with the initial parameters that are summarized in Table 4-3. The geometry of the resulting dual transmitarray antenna can be seen in Fig. 4-17, where $D_1 = 60$ cm and $D_2 = 1.8$ m. The reduction in the size of the first transmitarray associated to the 1.8 m main transmitarray with respect to the previous bifocal antenna ($D_1 = 1$ m) is due to the performance of the bifocal algorithm when the first transmitarray is moved towards the feeds, and constitutes the main reason why the illumination problems will remain (despite of reducing S_A distance, which was intended to increase the subtended angle from the feeds on the first transmitarray). The bifocal phase-shift distributions obtained for this design are shown in Fig. 4-18. As can be seen, the use of shorter S_A and larger S_B distances provides slightly smoother phase distributions than those shown in Fig. 4-13.

Fig. 4-17 Geometry of the dual transmitarray antenna to achieve beam compression.

TABLE 4-3
INITIAL PARAMETERS OF THE BIFOCAL SYNTHESIS

Parameter	Value
S_A	2 m
S_B	0.75 m
d	0.27 m
θ_{b1}	+1.4°
θ_{b2}	-1.4°

Fig. 4-18 Bifocal phase-shift distributions (in degrees) obtained for: (a) the first transmitarray and (b) the main transmitarray.

The simulated radiation patterns at 20 GHz in the principal planes for this bifocal dual transmitarray antenna are presented in Fig. 4-19. The results are very similar to those shown in Fig. 4-14; the maximum gain is close to 46 dB for the two beams generated from the foci, while SLL is slightly higher than before (around -19 dB).

Fig. 4-19 Simulated radiation patterns for the bifocal antenna: (a) in the elevation plane, (b) in the azimuth plane.

The reason for this reduction in the radiation efficiency of the bifocal antenna, as in the previous DTA configuration, are the high levels of illumination on the first transmitarray, which has to be oversized in order to reduce spillover. However, this modification leads to an increase of the illumination levels on the main transmitarray, which has to be also oversized (a diameter of around 4 m is required to reach around -12 dB taper on the edges). Therefore, the bifocal antenna is able to provide a high degree of beam spacing compression (by a factor of 2), but price to be paid is always the reduction in the radiation efficiency, as in the case of the bifocal dual reflectarray configuration that was shown in section 3.4.1.

4.5 Conclusions

The bifocal technique has been applied to dual transmitarray configurations for the design of multi-beam antennas in Ka-band. The design with transmitarrays brings some advantages, such as lower sensitivity to deformations and absence of blockage from the feeds or the first transmitting structure. The latter factor allows for the use of centered and rotationally-symmetrical geometries with a focal ring, which considerably simplifies the design process (no further adjustments are required after rotating the 2D design performed in the offset plane).

Two variations of the same baseline dual transmitarray geometry have been studied, in order to facilitate the practical implementation of such an antenna (integration of the feeds and the first transmitarray into the same sub-system, and holding of both transmitarrays by the same supporting structure). The variations in the dimensions and the required phase distributions for the dual transmitarray configurations have been evaluated in this chapter.

Although the design of bifocal dual transmitarray antennas allows for a high degree of beam spacing reduction, it presents the same efficiency problem than in the case of dual reflectarrays. The results of the simulations for a bifocal dual transmitarray antenna with a 1.8 m main transmitarray show that it is possible to achieve a high degree of beam spacing compression (by a factor of 2) with respect to the equivalent monofocal system, but at the cost of a reduced radiation efficiency. To increase the antenna efficiency for this degree of beam compression, the main transmitarray should be drastically oversized (around 4 m in diameter) to avoid spillover losses.

Chapter 5

General tridimensional bifocal method for dual reflectarray configurations

5.1 Introduction

The bifocal technique presented in Chapter 3 allows to overcome the problem of the 3D design of dual reflectarray configurations by rotation of a 2D design performed in the offset plane, considering an axially-symmetrical geometry with parallel reflectarrays and following an analogous approach to that used for the design of bifocal dual reflectors [80]. Then, the tilting of both reflectarrays was proposed in Chapter 3 to obtain smoother phase distributions. A novel phase correction routine was implemented to compensate the tilt of the reflectarrays and keep the bifocal characteristic of the original design, providing reasonable results for the radiation patterns of the bifocal antenna.

On the other hand, the technique proposed in Chapter 3 presents some limitations that may discourage its use for certain DRA geometries. First, the technique works if the relative tilt between the two reflectarrays is not very large. Also, it imposes some geometrical restrictions in the initial positions of the foci and the reflectarrays that remain throughout the design process of the bifocal antenna. These factors preclude the design of highly offset DRA configurations (as the compact-range geometry shown in [71]), which demand the implementation of a more general bifocal design procedure, valid for any possible arrangement of the foci and the two reflectarrays.

In this chapter, a general 3D bifocal technique is proposed for the design of dual reflectarray antennas, which makes it possible the direct synthesis of the phase

distributions in the selected antenna configuration, without geometrical constraints. First, the proposed bifocal method is validated for the axially-symmetrical geometry shown in section 3.2. Then, it is used to design a multi-beam dual reflectarray antenna in an offset compact-range configuration for operation in transmission in Ka-band. The phase distributions and radiation patterns of the bifocal antenna are studied for three different design cases (low beam spacing compression, high beam spacing compression and no beam spacing compression), and compared with those provided by the equivalent monofocal antenna. Also, the capabilities of the bifocal technique to provide a better multi-beam performance and a closer separation between adjacent beams are evaluated in the three cases. Although the results in this chapter are presented only for transmit antennas in Ka-band, the use of appropriate reflectarray cells that provide independent phase-shifts at 20 and 30 GHz will allow for the design of Tx and Rx antennas, with separate bifocal design processes for each frequency band.

5.2 Bifocal method for 3D design of dual reflectarray antennas

The implementation of a bifocal design method in 3D obliges to consider each reflectarray as a planar surface, which will be characterized by the partial derivatives of its phase distribution (Φ) with respect to the horizontal (y) and vertical (x) coordinates, according to the reference system of each reflectarray shown in Fig. 5-1. From the study of the reflecting properties of reflectarrays, it can be deduced that the partial derivatives of the phase at each reflectarray cell will satisfy the following relations with the incidence angles (θ_i, φ_i) and the reflection angles (θ_o, φ_o) of the rays [127]:

$$\Phi'_x = \frac{\partial \Phi(x, y)}{\partial x} = \frac{2\pi}{\lambda} \cdot (\sin \theta_i \cos \varphi_i - \sin \theta_o \cos \varphi_o) \quad (5-1)$$

$$\Phi'_y = \frac{\partial \Phi(x, y)}{\partial y} = \frac{2\pi}{\lambda} \cdot (\sin \theta_i \sin \varphi_i - \sin \theta_o \sin \varphi_o) \quad (5-2)$$

The previous expressions can be considered as the 3D extension of eq. (3-1), which was introduced in Chapter 3 for a 2D system. These expressions will provide upper and lower limits for the value of both partial phase derivatives, Φ'_x and Φ'_y , at each reflectarray cell, since they depend on sine and cosine functions of the incidence and reflection angles.

Concerning the geometry of the DRA system, it can be defined by the placement of the two reflectarrays (in parallel planes, with a certain tilting, etc.), the location of the focal points and the directions of the radiated beams associated to the foci. In this case, it will be assumed that the foci (F_1 and F_2) and their associated beam directions (θ_1 and θ_2) are both contained in the xz -plane (see Fig. 5-1), which will be the symmetry plane (or offset plane) of the DRA configuration.

Fig. 5-1 Geometry of an offset DRA configuration with tilted reflectarrays in the xz -plane, including the first iteration of the bifocal ray-tracing routine.

The geometrical parameters of the bifocal antenna must be fixed before starting the design procedure, according to the antenna specifications for the intended application. A reference single-focus design (dual reflector or dual reflectarray antenna) can be also used to estimate the dimensions of the bifocal antenna, at least as a first approximation. As was explained in Chapter 3, setting very stringent or improper initial conditions may lead to convergence problems of the bifocal algorithm, not being possible to reach a valid solution for the design of the bifocal antenna with such characteristics.

The partial phase derivatives Φ'_x and Φ'_y will be determined for a discrete grid of points on the surface of each reflectarray, by means of an iterative 3D ray-tracing routine that makes use of eqs. (5-1) and (5-2) and follows a similar procedure than in the bifocal design of dual reflectors [81]. The ray-tracing routine will be executed several times: each execution will provide a column of points on the surface of each reflectarray, and the addition of adjacent columns will form a grid. Then, the samples of the phase derivatives will be interpolated and properly integrated to obtain the required bifocal phase distributions on both reflectarrays. A block diagram with the steps of the proposed 3D bifocal design method is presented in Fig. 5-2. The different steps of the bifocal algorithm will be described in the following sections.

Fig. 5-2 Steps of the 3D bifocal design procedure.

5.2.1 Ray tracing procedure

The 3D ray-tracing routine can be executed several times, until obtaining a sufficient number of points to characterize each reflectarray. Each execution ‘ n ’ (with $1 \leq n \leq N$) requires an initial point on the sub-reflectarray plane, S_1^n , and the values of the partial phase derivatives associated to that point, $\Phi'_x(S_1^n)$ and $\Phi'_y(S_1^n)$. To ensure an efficient performance of the ray-tracing procedure, the points S_1^n must be located along the sub-reflectarray cross section (in the direction of y -axis, according to Fig. 5-1).

A transmitted ray from focus F_1 that impinges on S_1^n is reflected with angles (θ_o, ϕ_o) , which can be obtained by applying (5-1) and (5-2) at S_1^n . The reflected ray provides a new point on the main reflectarray, M_1^n , whose partial phase derivatives $\Phi'_x(M_1^n)$ and $\Phi'_y(M_1^n)$ are calculated by applying (5-1) and (5-2) at M_1^n with output angles $(\theta_1, 0^\circ)$. Similarly, a received ray in the direction $(\theta_2, 0^\circ)$ that impinges on M_1^n provides a new point on the sub-reflectarray, S_2^n , and its phase derivatives, $\Phi'_x(S_2^n)$ and $\Phi'_y(S_2^n)$, by applying (5-1) and (5-2) first at M_1^n , and then at S_2^n . The previous steps can be repeated starting with a transmitted ray from F_1 that impinges on S_2^n ; after M iterations, a row of M points is obtained on the surface of the main reflectarray, while $M+1$ points are achieved on the sub-reflectarray in a similar linear arrangement. In the end, the execution of the 3D ray-tracing routine N times will provide a grid of $N \cdot M$ points that characterize the main reflectarray, and $N \cdot (M+1)$ points for the sub-reflectarray. A flow chart summarizing the main steps of the 3D ray-tracing routine is shown in Fig. 5-3.

Due to the system’s symmetry with respect to the xz -plane, the 3D ray-tracing process can be performed only in the $y \geq 0$ region, and then the resulting points can be replicated in the other half-space ($y < 0$). For this purpose, note that the partial derivative Φ'_x will present even symmetry with respect to the xz -plane, while Φ'_y will present odd symmetry:

$$\Phi'_x(x_s, -y_s) = \Phi'_x(x_s, y_s) \quad (5-3)$$

$$\Phi'_y(x_s, -y_s) = -\Phi'_y(x_s, y_s) \quad (5-4)$$

As an example of the results that can be achieved, Fig. 5-4 shows the grids of points obtained for a DRA system with parallel reflectarrays (the results correspond to the DRA geometry analyzed in section 5.3). The samples of the normalized partial phase derivatives (multiplied by $-\lambda/2\pi$) associated to this geometry are shown in Fig. 5-5 for both sub- and main reflectarrays.

Fig. 5-3 Flow chart with the steps of the 3D ray-tracing procedure.

Fig. 5-4 Example of the grid of points obtained for each reflectarray after executing the 3D bifocal ray-tracing routine.

Fig. 5-5 Samples of the phase derivative: on the sub-reflectarray for (a) $\partial\Phi/\partial x$ and (b) $\partial\Phi/\partial y$, and on the main reflectarray for (c) $\partial\Phi/\partial x$ and (d) $\partial\Phi/\partial y$.

5.2.2 Setting of the initial values for the phase derivatives

There are several options to determine the value of the partial phase derivatives at the starting points of the 3D bifocal algorithm (S_1^n , with $1 \leq n \leq N$), so that the resulting phases make sense. In the more general case in which the DRA configuration does not present axial symmetry, as in the offset compact range geometry shown in [71], the equivalent single-focus design (considering the middle point between F_1 and F_2 as the new focus) can be used to fix the initial conditions for the 3D bifocal algorithm. The monofocal phase distributions can be calculated as described in [128], and then, differentiated with respect to x and y variables. The values of Φ'_x and Φ'_y for those points at the lower horizontal section of the sub-reflectorarray can be used as initial conditions to start each execution of the 3D bifocal algorithm (see section 5.4). The phase derivatives from the single-focus design will ensure that the beams are collimated in the azimuth plane, while the bifocal synthesis will be responsible for shaping and pointing the beams in elevation (xz -plane).

In a rotationally-symmetrical configuration, where the two reflectarrays are parallel to the xy -plane and the position of the foci and the beam directions are symmetrical with respect to z -axis, the values of Φ'_x and Φ'_y at the starting points S_1^n can be determined from the samples obtained after the first execution of the bifocal ray-tracing routine in the xz -plane. In that case, the algorithm starts at the point $S_1 = (0, 0, z_0)$ on the vertical axis of the sub-reflectorarray (placed at the plane $z = z_0$), with both Φ'_x and Φ'_y equal to zero (due to the axial symmetry conditions, as explained in section 3.2.1). After several iterations of the ray tracing, a set of points $S_i^1 = (x_i, 0, z_0)$ is obtained along the vertical axis of the sub-reflectorarray, with partial phase derivatives $(\Phi'_{xi}, 0)$. Note that Φ'_y will be null at the points S_i^1 , since it is an odd function with respect to y . Then, a set of points $S_i^1 = (0, y_i, z_0)$ placed along the horizontal axis which results from the intersection of the planes $x = 0$ and $z = z_0$ (thus being parallel to y -axis) is attained by rotating 90° around z -axis the points S_i^1 obtained in the xz -plane. The points S_i^1 are chosen as starting points at the plane $x = 0$ for the bifocal ray-tracing routine; the values of the phase derivatives associated to these points are obtained by applying the rotation symmetry properties:

$$\Phi'_x(0, y_i) = \Phi'_y(x_i, 0) \big|_{y_i=x_i} = 0 \quad (5-5)$$

$$\Phi'_y(0, y_i) = \Phi'_x(x_i, 0) \big|_{y_i=x_i} = \Phi'_{xi} \quad (5-6)$$

The starting points S_1^i obtained by this method correspond to the direction of the positive y -axis, but additional points can be obtained along the negative y -axis by applying (5-3) and (5-4), which allows to double the number of starting points by exploiting the symmetry conditions of both partial phase derivatives. Moreover, note that the points S_1^i placed at the plane $x = 0$ are used to start each execution of the bifocal ray-tracing routine, but the points selected to constitute the first row of the sub-reflectarray (its lower horizontal section) may be located at $x = x_0 > 0$, as in the Cassegrain design shown in section 3.2.

5.2.3 Integration of the partial phase derivatives

Once the phase derivatives samples are obtained on the surface of both reflectarrays, Φ'_x and Φ'_y must be interpolated by polynomials depending on x and y variables. Then, each partial derivative will be integrated with respect to x or y :

$$\int \Phi'_x(x, y) dx = f(x, y) + C_1(y) \quad (5-7)$$

$$\int \Phi'_y(x, y) dy = g(x, y) + C_2(x) \quad (5-8)$$

where $C_1(y)$ and $C_2(x)$ are the integration constants, which are polynomials of y and x , respectively. If the bifocal technique has been correctly applied and appropriate polynomials are selected for the interpolation of Φ'_x and Φ'_y , the terms of the form $x^p \cdot y^q$ will present the same coefficients in $f(x, y)$ and $g(x, y)$. Then, $f(x, y)$ may contain some terms of the form x^p , while $g(x, y)$ may contain some terms in y^q . The correct expression for the phase $\Phi(x, y)$ is a combination of the two partial integrals including the common terms and those depending on x^p and y^q , which can be expressed as:

$$\Phi(x, y) = f(x, y) + g(0, y) = f(x, 0) + g(x, y) \quad (5-9)$$

As an example, the unwrapped bifocal phase functions (in degrees) obtained for each reflectarray after the integration of the partial phase derivatives shown in Fig. 5-5 can be seen in Fig. 5-6. Note that the relation between the point distributions obtained in the bifocal synthesis for each reflectarray can be used to estimate the appropriate dimensions of both reflecting surfaces. The illumination will be concentrated in the regions where the points are very close one to each other, and therefore these regions should be included in the antenna to avoid a high spillover.

Fig. 5-6 Example of unwrapped bifocal phase distributions obtained for: (a) the sub-reflector and (b) the main reflector.

5.3 Validation in an axially symmetrical geometry

The 3D bifocal algorithm has been applied to the DRA geometry shown in section 3.2, which is an example of an axially-symmetrical configuration where the two reflectarrays are parallel to the xy -plane and the positions of the foci and beam directions are symmetrical with respect to z -axis (see Fig. 5-7). The two conjugate focal locations are: $F_1 = (-0.1, 0, 1)$ m and $F_2 = (0.1, 0, 1)$ m. The directions of the beams associated to the foci are: $(\theta_1 = 1.5^\circ, \phi_1 = 0^\circ)$ and $(\theta_2 = -1.5^\circ, \phi_2 = 0^\circ)$. The diameter of the main reflectarray is 1.8 m, and the position of its geometrical center in the reference coordinate system shown in Fig. 5-7 is $C_M = (1.7, 0, 0)$ m. The diameter of the sub-reflector is 60 cm and its center is located at $C_S = (0.5, 0, 1.5)$ m.

Fig. 5-7 Geometry of the axially-symmetrical DRA system under study.

In this kind of rotationally-symmetrical geometries, the bifocal procedure can be reduced to a 2D design problem, which only requires the execution of the ray-tracing routine in the xz -plane (see Fig. 5-7), as explained in Chapter 3. Due to the axial symmetry, the phase distribution on the sub-reflectarray will present a maximum or a minimum at $S_1 = (0, 0, z_0)$, so both partial phase derivatives Φ'_x and Φ'_y will be null at that point. After the execution of the ray-tracing routine starting at S_1 , a set of points $S_i = (x_i, 0, z_0)$ is obtained along the sub-reflectarray vertical axis in the xz -plane, and a set of points $M_j = (x_j, 0, 0)$ is obtained along the vertical axis of the main reflectarray. The values of Φ'_x associated to each set of points can be interpolated by polynomials depending on x variable, and then integrated to obtain the bifocal phase functions for each reflectarray in the xz -plane. These phase functions can be rotated around z -axis, and then, the design of an offset DRA configuration with parallel reflectarrays can be performed by selecting specific portions of the revolution surfaces.

The objective now is to validate the proposed 3D bifocal technique by comparing the phase distributions obtained for this geometry with those calculated by rotation of the bifocal phase curves in the xz -plane around z -axis. For this purpose, the points resulting from the execution of the bifocal ray-tracing routine in the xz -plane have been used to

obtain by a 90° rotation the starting points $S_1^n = (0, y_n, z_0)$, placed along the horizontal axis resulting from the intersection of the planes $x = 0$ and $z = z_0$ (which is also parallel to y -axis), as explained in section 5.2.2. Figure 5-8 shows the normalized values of Φ'_x and Φ'_y at the points S_1^n with respect to y coordinate. Note that Φ'_x is equal to zero for all points, while Φ'_y is an odd function.

Fig. 5-8 Normalized phase derivatives on the lower horizontal section of the sub-reflectarray, used as initial conditions for the 3D bifocal algorithm.

The grids of points obtained for each reflectarray, as well as the samples of the normalized phase derivatives corresponding to this DRA geometry have been shown in Fig. 5-4 and Fig. 5-5, respectively. The phase derivative samples have been interpolated using 9-degree polynomials and then integrated. The resulting bifocal phase functions have been shown in Fig. 5-6. The final phase-shift distributions corresponding to the 60-cm sub-reflectarray and the 1.8-m main reflectarray are presented in Fig. 5-9, normalized between -360° and 0° .

Fig. 5-9 Bifocal phase-shift distributions (in degrees) obtained by the 3D algorithm for: (a) the sub-reflectarray and (b) the main reflectarray.

Fig. 5-10 Difference (in degrees) between the phases obtained by the 3D algorithm and by the 2D algorithm with rotation of phase curves: (a) on the sub-reflectorarray and (b) main reflectarray.

The difference between the phases obtained by the 3D bifocal method and those obtained by rotation of the 2D bifocal phase curves in the xz -plane is shown in Fig. 5-10. As can be seen, this difference is lower than 1.2° for the main-reflectorarray elements, while in the sub-reflectorarray this difference is equal to 2° in the worst case (in the majority of the sub-reflectorarray elements, it is below 1°). Therefore, the difference between the two methods is almost negligible and can be attributed to the accuracy of the mathematical procedures involved in the synthesis of the bifocal phases. For example, this error could be reduced by using higher degree polynomials for the interpolation of the phase derivatives. In any case, the phase errors produced in the implementation of the phase with real reflectarray cells, or the manufacturing errors, will be larger than 2° (at least 10°).

The comparison of the simulated radiation patterns at 20 GHz in the principal planes for the two beams generated by the foci of the bifocal antenna are shown in Fig. 5-11 and Fig. 5-12. Note that the cross-polar radiation is not represented in the figures, because the radiation patterns have been computed assuming ideal reflectarray cells. The difference in the phases obtained by the two bifocal methods has virtually no effect on the radiation patterns, which present a very good agreement. These results prove the validity of the proposed 3D bifocal technique in a rotationally-symmetrical DRA configuration.

Fig. 5-11 Comparison of the simulated radiation patterns in the xz -plane for the 3D bifocal algorithm and the 2D algorithm with rotation of phase curves.

Fig. 5-12 Comparison of the simulated radiation patterns in the azimuth plane (orthogonal plane in the beam direction) for the 3D bifocal algorithm and the 2D algorithm with rotation of phase curves: (a) for the beam produced from F_1 ($\theta_1 = 1.5^\circ$), and (b) for the beam produced from F_2 ($\theta_2 = -1.5^\circ$).

5.4 Design of a multi-beam satellite antenna in Ka-band

The 3D bifocal technique has been applied to the offset compact-range configuration shown in Fig. 5-13, whose main geometrical parameters are summarized in Table 5-1, in order to design a multi-beam satellite antenna for operation in the Ka-band transmit frequencies. The antenna geometry has been adjusted to avoid blockage from the feeds or the sub-reflectarray. The period of the main reflectarray (main-RA) cells is 7.5 mm, while a 10 mm period has been used for the sub-reflectarray (sub-RA). The lack of axial symmetry and the large tilt angle of the sub-reflectarray with respect to the main reflectarray prevent from the application of the bifocal procedure presented in Chapter 3. Therefore, the synthesis of the required phase distributions on each reflectarray will be carried out by the 3D bifocal method proposed in section 5.2.

TABLE 5-1
MAIN GEOMETRICAL PARAMETERS OF THE COMPACT-RANGE SYSTEM

Parameter	Value
Size Main-RA	1.80 x 1.59 m (240 x 212 elements)
Size Sub-RA	1.31 x 1.23 m (131 x 123 elements)
Angle of tilting Sub-RA	47.55°
Coordinates center Main-RA	[0, 0, 0] mm
Coordinates center Sub-RA	[-644, 0, 1017] mm
Phase center F_1 (focus 1 BDRA)	[208, 0, 1108] mm
Phase center F_5 (focus 2 BDRA)	[412, 0, 1192] mm
Distance F_1 - F_5	216 mm
Phase center F_3 (focus MDRA)	[310, 0, 1150] mm
Virtual focus related to F_3	[-2546, 0, 4721] mm

Fig. 5-13 Geometry of the compact-range DRA system under study.

A linear arrangement of 7 adjacent feeds, from F_0 to F_6 , has been considered. The feed-horns present the same characteristics than those used in Chapter 3 (a realistic feed model characterized by Astrium [82] has been used as a reference). They have 54 mm diameter and provide -12 dB illumination on the sub-reflectorarray edges at 20 GHz when the sub-reflectorarray is illuminated with a subtended angle of 36° from the feed.

Prior to the design of the bifocal antenna, a single-focus reference design has been performed at 19.7 GHz for the central feed, F_3 , to radiate at $(\theta_3 = 28^\circ, \varphi_3 = 0^\circ)$ with respect to the normal vector to the main reflectarray surface (see Fig. 5-13). The required phase-shift distributions for this monofocal dual reflectarray antenna (MDRA) have been obtained by the method described in [128] and are shown in Fig. 5-14.

Fig. 5-14 Monofocal phase distributions (in degrees) required on the: (a) sub-reflectorarray and (b) main reflectarray.

5.4.1 Bifocal antenna with small beam spacing compression

The 3D bifocal algorithm has been used to synthesize the phases on the two reflectarrays, considering the phase centers of the feeds F_1 and F_5 as the foci of the bifocal dual reflectarray antenna (BDRA), which will generate two beams at 19.7 GHz in the directions $(\theta_1 = 30.24^\circ, \varphi_1 = 0^\circ)$ and $(\theta_5 = 25.76^\circ, \varphi_5 = 0^\circ)$. The designed BDRA will provide a separation of 1.12° between beams generated by contiguous feeds, whereas the equivalent MDRA will provide around 1.24° of beam spacing (as will be shown latter, see Fig. 5-22). This means that the BDRA will compress the beams by a factor of $1.24^\circ/1.12^\circ = 1.11$.

Due to the non-rotationally symmetrical geometry of the DRA system, the phase distribution on the sub-reflectorarray in the equivalent MDRA has been used to compute

the value of the partial phase derivatives on its lower horizontal section (parallel to y -axis), in order to provide a set of starting points for the 3D bifocal algorithm. The phase-shift distribution shown in Fig. 5-14(a) has been extended to consider a rectangular sub-reflectarray of size 1.31 x 1.23 m instead of a slightly elliptical one (see Fig. 5-15), so as to increase the number of elements in the first row of the sub-reflectarray. Then, the unwrapped monofocal phases have been differentiated with respect to x and y variables.

Fig. 5-15 Monofocal phase-shift distribution (in degrees) on the rectangular sub-reflectarray.

Figure 5-16 shows the normalized curves obtained for Φ'_x and Φ'_y with respect to y coordinate at the lower horizontal section of the sub-reflectarray. Note that Φ'_x presents even symmetry and Φ'_y is an odd function. A constant $\gamma_0 = -0.123$ has been previously added to the normalized Φ'_x curve so that its maximum is now $\Phi'_x(y = 0) = 0$. This adjustment does not affect the performance of the bifocal procedure, since the shape of the Φ'_x curve remains the same, but it allows to reduce the number of 360° cycles in the resulting phase-shift distributions (as will be shown later).

Fig. 5-16 Normalized phase derivatives on the lower horizontal section of the sub-reflectarray, used as initial conditions for the 3D algorithm.

Fig. 5-17 Grid of points obtained for each reflectarray after executing the 3D bifocal ray-tracing routine.

Fig. 5-18 Samples of the phase derivative: on the sub-reflectarray for (a) $\partial\Phi/\partial x$ and (b) $\partial\Phi/\partial y$, and on the main reflectarray for (c) $\partial\Phi/\partial x$ and (d) $\partial\Phi/\partial y$.

The grids of points obtained for the two reflectarrays after applying the ray-tracing routine to the current geometry can be seen in Fig. 5-17. The samples of the partial phase derivatives associated to each reflectarray are shown in Fig. 5-18. The unwrapped phase functions that arise from the interpolation and integration of the phase derivatives are shown in Fig. 5-19, and the final bifocal phase-shift distributions to be implemented on each reflectarray (normalized between -360° and 0°) are shown in Fig. 5-20.

Fig. 5-19 Unwrapped bifocal phase functions obtained for the: (a) sub-reflectorarray and (b) main reflectarray.

Fig. 5-20 Bifocal phase distributions (in degrees) required on the: (a) sub-reflectorarray and (b) main reflectarray.

The DRA system has been analyzed by applying the modular technique described in [70] and assuming ideal reflectarray cells in both surfaces. A $\cos^q(\theta)$ model with $q = 28$ has been used to simulate the electromagnetic field radiated by the 54-mm horns. The simulated radiation patterns of the seven beams produced by the BDRA at 19.7 GHz in elevation (xz -plane) and azimuth (orthogonal plane in the direction of the beam) are shown in Fig. 5-21. The maximum gain varies from 47.6 dBi to 47.85 dBi, and side-lobe levels are lower than -23 dB for all the beams.

Fig. 5-21 Simulated radiation patterns for the BDRA to provide 1.12° of beam spacing at 19.7 GHz: (a) superposition of cuts in the azimuth plane, and (b) cut in the xz -plane.

Moreover, Fig. 5-22 shows the comparison of the radiation patterns in the xz -plane for the beams generated by the BDRA (solid lines) and those produced by the equivalent MDRA (dashed lines). As can be seen, the BDRA provides a better performance for the extreme beams (around 0.45 dB larger gain and 3 dB lower SLL), while reducing at the same time the beam spacing in the xz -plane from 1.24° to 1.12° . So, beam spacing is compressed by a factor of $1.24^\circ/1.12^\circ = 1.11$ with respect to the MDRA for the same feed spacing. These results reinforce the validity of the proposed 3D bifocal technique when applied to non-rotationally symmetrical geometries, as well as its main advantages over the equivalent monofocal design.

The amplitude distributions of the incident electric field on both reflectarrays produced by the two feeds placed at the foci of the BDRA (F_1 and F_5) are shown in Fig. 5-23. The module of the incident field is close to -12 dB on the edges of both reflectarrays and the estimated radiation efficiency is around 55%, considering the expected gain from an elliptical aperture of 1.8×1.59 m.

Fig. 5-22 Simulated radiation patterns at 19.7 GHz in the xz -plane for the beams produced by the BDRA (solid lines) and by the MDRA (dashed lines).

Fig. 5-23 Amplitude (dB) of the incident field on the sub-reflectorarray when the antenna is illuminated from (a) F_1 and (b) F_5 , and on the main reflectarray for illumination from (c) F_1 and (d) F_5 .

The results presented in Chapter 3 showed that a high degree of beam spacing compression (by a factor of 2) leads to a reduction in the radiation efficiency of the bifocal antenna. However, the bifocal technique can be safely applied with quite good results for the antenna efficiency when beam spacing is reduced by a relatively low factor, as in this case. Also, note that if the reflectarrays' capability to discriminate in polarization is used, a second bifocal design process can be performed in the orthogonal polarization to generate the interleaved beams for a 0.56° final separation in the xz -plane (as will be shown in Chapter 6, concerning the design of a small-scale BDRA demonstrator).

Concerning the correction performed in the normalized value of Φ'_x at the starting points of the 3D ray-tracing routine (adding a constant $\gamma_0 = -0.123$), a second bifocal design procedure has been carried out for the same DRA system, but considering the initial values of Φ'_x and Φ'_y on the lower horizontal section of the sub-reflectarray as directly calculated from the equivalent single-focus design. In this case, the bifocal phase-shift distributions shown in Fig. 5-24 are obtained.

Fig. 5-24 Bifocal phase distributions (in degrees) obtained on the (a) sub-reflectarray and (b) main reflectarray without any correction in the initial condition for the Φ'_x curve.

The effect of adding a constant γ_0 to the initial values of the normalized Φ'_x curve on the sub-reflectarray is equivalent to the inclusion of a progressive phase term of the form $\alpha_0 \cdot x$ in the phase distribution of the single-focus design, where $\alpha_0 = (-2\pi/\lambda) \cdot \gamma_0$. Note that this term will not affect the value of the Φ'_y derivative, as it only depends on x . Figure 5-25 shows the variation in the monofocal phase-shift distribution of the sub-reflectarray before and after adding the progressive phase term resulting from the adjustment performed in the initial Φ'_x curve ($\gamma_0 = -0.123$).

Fig. 5-25 Phase-shift distributions (in degrees) on the sub-reflectarray for the reference monofocal antenna: (a) in the original monofocal design, and (b) after adding a progressive phase term.

The progressive phase term added to the phase distribution of the sub-reflectarray would produce a shifting in the directions of the radiated beams; however, the bifocal algorithm is able to compensate this variation in the sub-reflectarray phases by adding a progressive phase term of opposite sign in the phase distribution of the main reflectarray. This fact can be checked by subtracting the phase-shift distributions shown in Fig. 5-20 (with the corrected Φ'_x curve, $\gamma_0 = -0.123$) to those shown in Fig. 5-24 (with Φ'_x directly calculated from the single-focus reference design, $\gamma_0 = 0$). The results can be seen in Fig. 5-26.

Fig. 5-26 Difference (in degrees) between the phase distributions with and without correcting the initial condition for the Φ'_x curve: (a) on the sub-reflectarray, and (b) on the main reflectarray.

Furthermore, the simulated radiation patterns at 19.7 GHz in the xz -plane for the second BDRA design (with the phase-shift distributions in Fig. 5-24) have been obtained. The comparison with the patterns shown in Fig. 5-21 for the initial BDRA, designed with the corrected Φ'_x curve, is presented in Fig. 5-27.

Fig. 5-27 Comparison of the radiation patterns at 19.7 GHz in the xz -plane for the beams produced by the BDRA with modified Φ'_x curve (solid lines) and by the BDRA with original Φ'_x curve (dashed lines).

Therefore, it can be concluded that the correction introduced in the value of the Φ'_x curve used as initial condition for the 3D bifocal algorithm allows to obtain more centered phase distributions for both reflectarrays (meaning that the elements with the smoothest phase variations are near to the geometrical centers of both reflectarrays), without any effect on the radiation patterns of the bifocal antenna. It has been checked that a larger value of γ_0 causes a greater variation in the bifocal phase distributions with respect to those calculated for $\gamma_0 = 0$. However, note that the allowed values of γ_0 will be limited by the application of eq. (5-1) in the subsequent 3D ray-tracing procedure, which fixes a limit for the value of Φ'_x . These results are very interesting because they help us to understand better the behavior of the bifocal algorithm.

5.4.2 *Bifocal antenna with large beam spacing compression*

To continue with the study of the bifocal technique, a new design has been performed applying the 3D bifocal method for the same DRA configuration in order to reduce the beam spacing to 0.56° (typical value for current multi-spot satellite applications in Ka-band). The high ratio of beam spacing compression ($\text{BCR} = 1.24/0.56^\circ = 2.21$) will be responsible for the appearance of a larger number of 360° cycles in the bifocal phase-shift distributions of both reflectarrays. This can be partially corrected by adding a certain constant γ_0 to the normalized Φ'_x curve in the lower horizontal section of the sub-reflectarray, as explained in the previous section. In this case, the best results for the phase distributions are obtained with $\gamma_0 = -0.563$ (see Fig. 5-28), which is the minimum allowed value for γ_0 . After this adjustment, the maximum

of the normalized Φ'_x curve is $\Phi'_x(y=0) = -0.44$, while the shape of the curve remains the same than in Fig. 5-16.

Fig. 5-28 Required phase-shift distributions (in degrees) for the bifocal antenna to provide 0.56° of beam spacing: (a) on the sub-reflector array, and (b) on the main-reflector array.

The simulated radiation patterns in the elevation and azimuth orthogonal planes for this BDRA are shown in Fig. 5-29. The gain varies from 42.8 dBi to 44.5 dBi, and the SLL is equal to or lower than -16 dB for all the beams. Despite achieving the required beam compression (0.56° separation between adjacent beams), the antenna presents a serious efficiency problem, which can be noticed in the reduced gain of the beams.

A comparison of the simulated radiation patterns at 19.7 GHz in the xz -plane for the beams generated by the BDRA (solid lines) and those produced by the equivalent MDRA (dashed lines) is shown in Fig. 5-30. As can be seen, the BDRA is able to reduce beam spacing from 1.24° to 0.56° , so the beams are compressed by a factor of $1.24/0.56 = 2.21$ with respect to the MDRA. On the other hand, the beams from the BDRA present around 3-4 dB lower gain than the beams from the MDRA (about 47.7 dBi gain), and SLL is between 6 and 7 dB higher than in the monofocal case.

The amplitude distributions of the incident field on both reflectarrays considering illumination from F_1 and F_5 are shown in Fig. 5-31. Note that the illumination levels on the edges of the main reflectarray are very high, which produces a lot of spillover, and thus, a reduction in the radiation efficiency of the bifocal antenna. These results are caused by the application of the bifocal technique for a high degree of beam spacing compression ($BCR = 2.21$), which spreads the illumination along the vertical axis of the main reflectarray (the direction in which the beams are compressed) with respect to the amplitude distributions shown in Fig. 5-23 for the BDRA with $BCR = 1.11$.

Fig. 5-29 Simulated radiation patterns for the BDRA to provide 0.56° of beam spacing at 19.7 GHz: (a) superposition of cuts in the azimuth plane, and (b) cut in the xz -plane.

Fig. 5-30 Comparison of the radiation patterns in the xz -plane for beams generated at 19.7 GHz by the BDRA (solid lines) and by the equivalent MDRA (dashed lines).

Fig. 5-31 Amplitude (dB) of the incident field on the sub-reflectarray for (a) F_1 and (b) F_5 , and on the main reflectarray for (c) F_1 and (d) F_5 .

As indicated in Chapter 3, introducing some modifications in the initial conditions of the bifocal synthesis or changing the antenna configuration does not mitigate the illumination problem. The only possible solution would be to increase the size of the main reflectarray, which implies using a very large reflectarray, around 3.5 m in diameter (twice its original size). However, enlarging the size of the main reflectarray in the current geometry will produce blockage from the feeds (see Fig. 5-13). Therefore, the antenna geometry should be modified to allow the implementation of a bifocal antenna with an oversized main reflectarray, which will reduce spillover and increase the radiation efficiency of the antenna (as will be shown in Chapter 7).

5.4.3 *Bifocal antenna with no beam compression*

A third bifocal design has been carried out in which the separation between adjacent beams will be the same than in the single-focus reference design: 1.24° (so, $BCR = 1$). The idea is to apply the bifocal algorithm to correct beam aberration and improve the multi-beam performance with respect to the equivalent MDRA. In this design, the

extreme feeds of the array (F_0 and F_6) have been considered as the foci of the bifocal antenna, which will generate two beams at 19.7 GHz in the directions ($\theta_0 = 31.72^\circ$, $\varphi_0 = 0^\circ$) and ($\theta_6 = 24.28^\circ$, $\varphi_6 = 0^\circ$). This modification has been performed to make more evident the differences between the BDRA and the equivalent MDRA, since the extreme beams of the MDRA are the most degraded (they are produced by the feeds with the largest separations from the antenna focus). On the other hand, the BDRA will ensure perfect focusing for those two beams, at the cost of a slightly worst performance for the central beam. The bifocal algorithm has been applied considering as initial condition the values of Φ'_x and Φ'_y directly obtained from the phase distribution of the sub-reflectorarray in the MDRA (without adding any constant). The resulting phase-shift distributions are shown in Fig. 5-32. It is very interesting that both phase distributions are quite centered and present a relatively low number of 360° cycles.

Fig. 5-32 Required phase-shift distributions (in degrees) for the bifocal antenna to provide 1.24° of beam spacing: (a) on the sub-reflectorarray, and (b) on the main-reflectorarray.

Therefore, it can be concluded that applying the bifocal method with the initial values of Φ'_x and Φ'_y directly calculated from the equivalent MDRA and with $\text{BCR} = 1$ (same beam spacing than in the MDRA) will result in optimal phase-shift distributions for the BDRA. When a certain degree of beam compression is required, the initial condition for Φ'_x can be adjusted by means of γ_0 in order to reduce the number of 360° and obtain more centered phase distributions, although this adjustment is limited by the range of allowed values for γ_0 . If the BCR is very high (as in the design shown in section 5.4.2, with $\text{BCR} = 2.21$), the large number of 360° cycles in the phase-shift distributions of both reflectarrays cannot be completely avoided. The simulated radiation patterns at 19.7 GHz in the principal planes for this BDRA with $\text{BCR} = 1$ are shown in Fig. 5-33. The gain varies from 47.45 dBi to 47.78 dBi, being the central beam

the one with the highest gain, and side-lobe levels are lower than -23.5 dB with respect to the maximum for all the beams. Note that all the beams present similar values of gain, beamwidth and SLL.

Fig. 5-33 Simulated radiation patterns for the BDRA to provide 1.24° of beam spacing at 19.7 GHz: (a) superposition of cuts in the azimuth plane, and (b) cut in the xz -plane.

The comparison between the beams generated by the BDRA and those generated by the equivalent MDRA is shown in Fig. 5-34. As can be seen, the beams produced by the BDRA and the MDRA present almost the same gain, although the bifocal antenna provides 0.3 dB higher gain for the beams produced by F_0 and F_1 . The main advantages of the bifocal design over the monofocal one are a better shaping of the main lobes and lower SLL, especially for the beams generated from F_0 , F_1 , F_5 and F_6 . On the other hand, the central beam produced by the BDRA presents around 3 dB higher SLL than the one produced by the MDRA. The performance of the bifocal antenna provides higher values of single-entry C/I (where the interference is produced by the radiation of the adjacent beams) with respect to the equivalent MDRA.

Fig. 5-34 Comparison of the radiation patterns in the xz -plane for the beams generated at 19.7 GHz by the BDRA (solid lines) and the equivalent MDRA (dashed lines).

The main characteristics of the bifocal beams, together with the improvement in the single-entry C/I with respect to the MDRA, are summarized in Table 5-2. The single-entry C/I has been measured within a 0.65° beamwidth, which corresponds to a gain of around 44.75 dBi. Note that there are some small pointing errors in the central beams of the bifocal design, but they are lower than the pointing errors in some of the extreme beams of the equivalent MDRA (-0.04° error is obtained for the beam no. 4, -0.1° error for the beam no. 5, and -0.15° error for the beam no. 6). The pointing errors in the BDRA are in general smaller than the pointing errors of the satellite and can be corrected in a second step by slightly adjusting the position of the feeds during the detailed design of the antenna.

TABLE 5-2
MAIN CHARACTERISTICS OF THE BEAMS (BCR = 1)

No. beam	Gain (dBi)	Beamwidth at 44.7 dBi ($^\circ$)	C/I (dB)	Δ C/I (dB)	Pointing error ($^\circ$)
0	47.62	0.657	24.23	-	0
1	47.73	0.659	23.69	+4.63	+0.02
2	47.76	0.661	23.17	+2.71	+0.06
3	47.78	0.662	22.73	-0.53	+0.06
4	47.71	0.663	22.54	+1.86	+0.06
5	47.63	0.663	23.71	+4.32	+0.02
6	47.45	0.659	24.31	-	0

The amplitude distributions of the incident field on the two reflectarrays produced by the extreme feeds (F_0 and F_6) are shown in Fig. 5-35. As can be seen, the module of the incident field is close to -12 dB on the edges of both reflectarrays, except for the illumination produced by F_0 on the main reflectarray, which is moved to the upper reflectarray edge. To overcome this effect, either the main reflectarray should be slightly oversized, or the inclination of the feeds adjusted to point to different parts of the sub-reflectarray, so as to achieve a more centered illumination on the main reflectarray for all the feeds.

Fig. 5-35 Amplitude (dB) of the incident field on the sub-reflectarray for (a) F_0 and (b) F_6 , and on the main reflectarray for (c) F_0 and (d) F_6 .

5.4.4 *Radiation patterns of the bifocal antenna in the azimuth plane*

It has been checked that the radiation patterns in the azimuth plane of the three BDRAs designed in the previous sections (with $BCR = 1$, $BCR = 1.11$ and $BCR = 2.21$) are very similar to those obtained for the equivalent MDRA with 1.24° of beam separation. Particularly, the same 3-dB beamwidth is achieved for the azimuth patterns shown in Fig. 5-21, Fig. 5-29 and Fig. 5-33. To illustrate this point, Fig. 5-36 shows the

superposition of cuts in the azimuth plane at 19.7 GHz for the beam produced by F_3 (central feed) in the MDRA and in the three designed BDRA. As can be seen, quite good agreement is achieved for the main lobe and the adjacent secondary lobes of the patterns, except for the beam produced by the BDRA with $BCR = 2.21$, which presents around 4 dB lower gain due to the high BCR. Furthermore, an additional beam has been obtained in the azimuth plane, adjacent to the central beam, for the bifocal antennas designed with $BCR = 1.11$ and $BCR = 2.21$, see Fig. 5-37. This beam is produced by a 54 mm feed-horn placed contiguous to F_3 in the direction of y -axis, which is named F_3^* . The spacing between the beams generated from F_3 and F_3^* in both BDRA is around 1.19° , which is almost the same beam spacing provided by the MDRA. However, the main beam is slightly broadened for the BDRA with a high beam compression ratio.

Fig. 5-36 Comparison of the radiation patterns in the azimuth plane for the central beam generated in all the previous DRA designs.

Fig. 5-37 Simulated radiation patterns at 19.7 GHz in the azimuth plane: (a) for the BDRA with 1.12° of beam spacing in the xz -plane, and (b) for the BDRA with 0.56° of beam spacing in the xz -plane.

Therefore, it can be concluded that the BDRA's designed by using the 3D bifocal technique are able to provide the required degree of beam spacing compression in the xz -plane, while the monofocal characteristic of the equivalent MDRA is almost preserved in the plane orthogonal to the xz . The reason for this behaviour is that the phase distributions from the MDRA have been used to obtain the initial condition for the phase derivatives along the sub-reflectarray cross section, when performing the 3D bifocal design process. Changing the initial conditions for the bifocal algorithm (i. e., using a different monofocal design, or even a bifocal design, as reference to compute the phase derivatives) will modify the performance of the BDRA in the azimuth plane.

5.4.5 Conclusions

The proposed 3D bifocal technique has been used for the design of a multi-beam dual reflectarray antenna in an offset compact-range configuration. The antenna has been designed at 19.7 GHz to operate in transmission (from the satellite) in Ka-band. The 3D bifocal algorithm has been applied considering different degrees of beam spacing compression with respect to the equivalent monofocal antenna: no beam compression, low beam compression and high beam compression. It has been shown that:

- The use of the monofocal phases to obtain the initial conditions for the 3D bifocal algorithm (the partial phase derivatives Φ'_x and Φ'_y along the sub-reflectarray cross section) provides centered phase-shift distributions and a low number of 360° cycles in the case of designing with $BCR = 1$ (no beam compression).
- The phase distributions obtained with the 3D bifocal method when $BCR > 1$ can be corrected by adding a progressive phase term depending on x variable to the monofocal phase distribution on the sub-reflectarray, without affecting the radiation patterns of the bifocal antenna.
- The bifocal antenna with $BCR = 1$ improves the performance of the extreme beams with respect to the equivalent monofocal antenna when it is used for the generation of multiple beams, providing a better shaping of the main lobe, lower SLL and higher C/I.

- The bifocal antenna with low beam compression ($\text{BCR} = 1.11$) allows to obtain closer beams with non-overlapping feeds than in the equivalent monofocal antenna, at the same time as improving the performance of the extreme beams. The first factor allows to reduce the required antenna size with respect to conventional antennas for the same beam spacing.
- The bifocal antenna with high beam compression ($\text{BCR} = 2.21$) is able to provide 0.56° of beam spacing, which is the required value for current multi-spot satellite applications in Ka-band, although the radiation efficiency of the antenna is reduced and the phases present a large number of 360° cycles. The main reflectarray should be oversized to avoid the first problem.
- The radiation patterns of the bifocal antenna in the azimuth plane are determined by the performance of the equivalent monofocal antenna, whose phase distributions have been used as initial condition for the 3D bifocal method. The beam spacing provided by the BDRA in the azimuth plane will be the same than the one provided by the MDRA.

5.5 Conclusions

A 3D bifocal design technique has been proposed for dual reflectarray antennas, which makes it possible the direct synthesis of the required phase distributions in the selected antenna configuration, without imposing any geometrical restrictions in the positions of the focal points or the reflectarrays. The technique is based on an iterative 3D ray-tracing routine that alternates transmitted and received rays to provide a grid of points that characterize each reflectarray (this iterative GO-based routine can be considered as an extension of the 2D procedure shown in Chapter 3). The samples of the partial phase derivatives Φ'_x and Φ'_y associated to the previous points are interpolated by polynomials, and then, properly integrated to obtain the required bifocal phase functions on each reflectarray. The initial condition for the partial phase derivatives along the sub-reflectarray cross section can be determined, in the more general case, from the equivalent single-focus design, considering the mid-point between the foci of the bifocal antenna as the focus of the monofocal antenna.

The 3D bifocal technique has been validated for the axially-symmetrical dual reflectarray configuration presented in Chapter 3 (section 3.2). The results obtained by

the 3D bifocal technique have been compared with those achieved by applying the 2D bifocal algorithm in the offset plane followed by the rotation of the resulting phase curves around the antenna symmetry axis. The results of this comparison are very satisfactory, as the differences between the two design methods are practically negligible.

A multi-beam dual reflectarray antenna has been designed to operate in transmission in Ka-band (19.7 GHz) with a 1.8 x 1.6 m main reflectarray and a 1.3 x 1.2 m sub-reflector. The antenna presents a dual offset compact-range configuration, which precludes the application of the bifocal method shown in Chapter 3. The required phase-shift distributions on both reflectarrays have been obtained by means of the proposed 3D bifocal technique, considering three different degrees of beam spacing compression: no compression (1.24° of beam spacing), low compression (from 1.24° to 1.12°) and high compression (from 1.24° to 0.56°). The results have been compared with those provided by the equivalent single-focus antenna with the same configuration. The bifocal antenna with no beam compression improves the performance of the extreme beams with respect to the equivalent monofocal antenna in terms of beam shaping, SLL and C/I. The bifocal antenna with low beam compression allows to obtain closer beams with non-overlapping feeds than in the equivalent monofocal antenna, at the same time than improving the performance of the extreme beams. Finally, the bifocal antenna with high beam compression is able to provide the required 0.56° spacing for the current multi-spot applications in Ka-band, although the radiation efficiency of the antenna is reduced.

Among the three cases studied, the bifocal antenna with low beam spacing compression is probably the most promising for multi-beam antennas in Ka-band, as it combines the smaller beam spacing with an improvement of the antenna performance for the generation of multiple beams. The first factor will allow to reduce the antenna size with respect to conventional reflectors to provide the same beam spacing. Moreover, the manufacturing of the bifocal dual reflectarray antenna will involve the same conventional processes used for printed reflectarrays (unlike the case of bifocal dual reflectors, which require expensive custom moulds for the two shaped reflectors). For these reasons, the bifocal antenna with low beam compression has been the selected option for the design, manufacturing and testing of a DRA prototype that will be used to validate the proposed bifocal technique, as will be shown in the following chapter.

Chapter 6

Design, manufacturing and test of a bifocal dual reflectarray antenna demonstrator

6.1 Introduction

A bifocal dual reflectarray antenna (BDRA) demonstrator that operates in dual-linear polarization in the 19.2-20.2 GHz band (transmission frequencies from a satellite in Ka-band) has been designed, manufactured and tested, in order to validate the bifocal technique and show its main advantages with respect to the equivalent single-focus design: a better antenna performance for the generation of multiple beams and a closer separation between adjacent beams for the same feed spacing.

The antenna configuration is based on an offset compact-range geometry, which is a small-scale version of the one used in Chapter 5 for the design of a multi-beam transmit antenna in Ka-band. The absence of axial symmetry and the large tilting angle of the sub-reflectarray with respect to the main reflectarray preclude the application of the bifocal design procedure proposed in Chapter 3. Therefore, the synthesis of the required phase-shift distributions on each reflectarray will be carried out by means of the 3D bifocal method presented in Chapter 5.

The BDRA demonstrator generates 10 beams alternating in dual-linear polarization when the antenna is illuminated by 5 contiguous feed-horns, but the method can be extended to generate adjacent beams in dual-circular polarization by implementing a sequential rotation of the reflectarray elements [59]. The spacing between adjacent beams in the same polarization is 3.8° , which means that the beams are compressed by a factor of 1.2 with respect to the equivalent monofocal design (4.6° of spacing). Moreover, the fabrication of the BDRA prototype involves the same conventional

processes used for printed reflectarrays, thus avoiding the high manufacturing costs of shaped bifocal dual reflectors. Finally, the same concept can be used for the design of transmit and receive antennas if appropriate reflectarray cells that enable independent phasing at Tx and Rx frequencies are employed [60], [68], [129], [130].

6.2 Design of the bifocal dual reflectarray antenna demonstrator

A BDRA demonstrator has been designed, manufactured and tested in order to validate the proposed bifocal technique and show its ability to reduce beam spacing, while providing a better performance for the extreme beams than the equivalent monofocal dual reflectarray antenna (MDRA). The prototype has been designed at 19.7 GHz to operate in the 19.2-20.2 GHz band, which are transmission frequencies from the satellite in Ka-band. A separate design process has been performed for each linear polarization, in order to obtain the required bifocal phase-shift distributions that must be implemented on each reflectarray.

6.2.1 Antenna definition

Figure 6-1 shows the geometry of the BDRA demonstrator, whose main parameters are summarized in Table 6-1. An arrangement of 5 contiguous feeds (from F_1 to F_5) placed in the xz -plane has been considered to illuminate the antenna. The diameter of the feed-horns is 60 mm, so the separation between the phase centers of adjacent feeds has been set to 61 mm, in order to allow a 1 mm margin to properly accommodate the horns. The antenna geometry has been adjusted to avoid blockage from the feeds or the sub-reflectarray.

Fig. 6-1 Geometry of the DRA demonstrator.

TABLE 6-1
MAIN GEOMETRICAL PARAMETERS OF THE BDRA DEMONSTRATOR

Parameter	Value
Size Main-RA	570 x 420 mm (76 x 56 elements)
Size Sub-RA	390 x 352.5 mm (52 x 47 elements)
Angle of tilting Sub-RA	51.1°
Center Main-RA	[0, 0, 0] mm
Center Sub-RA	[-240.3, 0, 330.7] mm
Phase center F_1 (focus 1 BDRA)	[337.1, 0, 47.8] mm
Phase center F_5 (focus 2 BDRA)	[403.2, 0, 282.8] mm
Distance F_1 - F_5	244 mm
Phase center F_3 (focus MDRA)	[370.1, 0, 165.3] mm
Virtual focus related to F_3	[-1321.4, 0, 2384.5] mm

As previously explained, an offset compact-range configuration has been selected for the BDRA demonstrator. This configuration is characterized by its large F/D ratio (see Fig. 6-2), associated to the virtual focus of the equivalent parabolic main reflector. The large focal distance will allow for smoother phase distributions (with a low number of 360° cycles) on the main reflectarray, as will be shown later. Moreover, the antenna requires less space than a Cassegrain or Gregorian configuration, thanks to the compact arrangement of the two reflectarrays and the feeds. The size of the sub-reflectarray has been selected for a subtended angle of around 30° from the feeds, which will provide around -12 dB edge illumination on its edges. Also, both reflectarrays are slightly elliptical, in order to ensure proper illuminations.

Fig. 6-2 Compact-range dual reflectarray configuration with large F/D.

6.2.2 Characterization of the feed

A 60-mm diameter horn manufactured by ANTERAL [131] has been used to illuminate the dual reflectarray demonstrator (see Fig. 6-3). The horn meets all the requirements for current multi-spot satellite applications in Ka-band, so it operates both at Tx and Rx frequency bands (17.6-20.3 GHz for Tx band and 27.3-30.1 GHz for Rx band), providing very low levels of cross-polar radiation and a high aperture efficiency. The inner profile of the horn is shown in Fig. 6-4, together with the position of its phase center at each frequency band. For design purposes, the position of the phase center at 19.7 GHz has been estimated at 5 mm inside the aperture plane.

Fig. 6-3 Feed-horn antenna.

Fig. 6-4 Inner profile of the horn and position of its phase center at each frequency band [131].

A simplistic model based on a $\cos^q(\theta)$ distribution has been then used to simulate the electromagnetic field radiated by the horn. Note that the sub-reflectarray elements are placed in the far field region of the feeds, since $2 \cdot D^2/\lambda = 473$ mm and the distances from the horn locations (F_1 to F_5) to the sub-reflectarray elements are comprised between 560 mm and 775 mm. The radiation patterns of the horn provided by the manufacturer at

18.9 and 20.3 GHz are shown in Fig. 6-5. The horn presents a directivity of 20.9 dBi at 18.9 GHz, while 21.6 dBi directivity is achieved at 20.3 GHz. The aperture efficiency is close to 0.89 and cross-polar levels are lower than -25 dB respect to the co-polar maximum within a 12-dB beamwidth. The radiation patterns present a different beamwidth in the two principal planes when operating in linear polarization, as can be seen in Fig. 6-5, where the dashed lines indicate the half-subtended angle to get an edge illumination level of -12 dB on the sub-reflectarray. The 12-dB beamwidth of the horn at 19.7 GHz has been estimated at 29° in the E-plane and at 32.2° in the H-plane.

Fig. 6-5 Radiation patterns of the feed at: (a) 18.9 GHz and (b) 20.3 GHz [131].

According to the previous data, the electromagnetic field radiated by the horn at 19.7 GHz has been modelled by a $\cos^q(\theta)$ function with $q = 43$ in the xz -plane (which corresponds to a 12-dB beamwidth of 29° in the E-plane of the horn) and $q = 35$ in the azimuth plane (12-dB beamwidth of 32.2° in the H-plane of the horn) for X-polarization, while the q values are reversed for Y-polarization, since the field of the horn is rotated 90 degrees.

6.2.3 Design of the reflectarray unit cell

A single-layer unit cell based on orthogonal arrangements of parallel dipoles has been used for the design of both sub- and main reflectarrays (see Fig. 6-6). There are three parallel dipoles for controlling the phases that will be introduced in each linear

polarization at 19.7 GHz. The side dipoles of each set present the same dimensions, so as to keep low levels of cross-polar radiation [30]. The period is fixed at $P_X = P_Y = 7.5$ mm, which is $\lambda/2$ at 20 GHz. This allows to avoid the appearance of grating lobes and provide enough range of phase-shift at 19.7 GHz, while preventing orthogonal dipoles from overlapping.

Fig. 6-6 View of the reflectarray periodic structure, including four unit-cells for X-polarization and one unit-cell for Y-polarization.

The geometrical parameters of the cell have been adjusted to provide a linear response of the phase with respect to the dipoles' lengths and robustness against variations in the angle of incidence. In this sense, a width of 0.5 mm has been selected for all dipoles, the center-to-center separations are $S_A = S_B = 1$ mm, and a scale factor of 0.78 is considered for the lengths of side dipoles with respect to the length of the central ones ($l_{A1} = 0.78 \cdot l_{A2}$ and $l_{B1} = 0.78 \cdot l_{B2}$). The substrate layer has been implemented by a Diclاد 880B sheet with a thickness (h) of 1.524 mm and a loss tangent ($\tan \delta$) of 0.001. For design purposes, the measured value of the dielectric constant at 20 GHz ($\epsilon_r = 2.3$) has been used, instead of the nominal value ($\epsilon_r = 2.17$ at 10 GHz) provided by the manufacturer.

The analysis and design of the reflectarray element has been carried out by means of a home-made SD-MoM (Method of Moments in the Spectral Domain) code, which uses entire domain basis functions to approximate the current densities on the dipoles, and assumes an infinite periodic array model for the analysis of each cell [101]. The SD-MoM code takes into account the real angles of incidence on each reflectarray cell to compute the co- and cross-polar coefficients that form the reflection matrix of the cell (further details can be found in Chapter 2). Figure 6-7 shows the magnitude and phase of the co-polar reflection coefficient for X and Y polarizations under different angles of

incidence, which are the highest angles from the feeds on the sub-reflectarray cells and from the virtual focus on the main reflectarray cells. As can be seen, the phase curves are quite linear in a range larger than 360° , and variations with the angle of incidence are lower than 25° in the worst case.

Fig. 6-7 Magnitude and phase of the co-polar reflection coefficient at 19.7 GHz, considering the most critical angles of incidence: (a) for X-polarization and (b) for Y-polarization.

6.2.4 *Design of the dual reflectarray antenna*

Considering the geometry of the DRA demonstrator shown in Fig. 6-1, a single-focus reference design has been performed at 19.7 GHz for the central feed (F_3) to produce two beams in the directions ($\theta_{3X} = 27.8^\circ$, $\varphi_{3X} = 0^\circ$) for X-polarization and ($\theta_{3Y} = 29.7^\circ$, $\varphi_{3Y} = 0^\circ$) for Y-polarization. The required phase-shift distributions on each reflectarray associated to this monofocal antenna are shown in Fig. 6-8 for both polarizations. The calculation of these phases has been performed following the same procedure than in [71], as in the case of the MDRA design presented in Chapter 5.

Fig. 6-8 Monofocal phase distributions (in degrees) at 19.7 GHz on the sub-reflectarray (a) in X-pol. and (b) in Y-pol.; and on the main reflectarray (c) in X-pol. and (d) in Y-pol.

The bifocal phase distributions that must be implemented on each reflectarray have been calculated by performing two independent bifocal design processes, one for each polarization. For this purpose, the phase centers of the feeds F_1 and F_5 have been selected as the foci of the bifocal antenna. The directions of the beams generated by the foci in each linear polarization are: ($\theta_{1X} = 35.4^\circ$, $\varphi_{1X} = 0^\circ$) and ($\theta_{1Y} = 37.3^\circ$, $\varphi_{1Y} = 0^\circ$) for feed F_1 ; and ($\theta_{5X} = 20.2^\circ$, $\varphi_{5X} = 0^\circ$) and ($\theta_{5Y} = 22.1^\circ$, $\varphi_{5Y} = 0^\circ$) for feed F_5 . Note that the beams in Y-polarization are shifted 1.9° with respect to those in X-polarization. The

spacing between beams in the same polarization produced by contiguous feeds is 3.8° , while it is 4.6° in the equivalent MDRA (as will be shown later, see Fig. 6-10). So, the BDRA demonstrator reduces the beam spacing provided by the MDRA by a factor of $4.6^\circ/3.8^\circ = 1.2$, for the same feed spacing. The resulting bifocal phase distributions to be implemented at 19.7 GHz in X and Y polarizations are shown in Fig. 6-9.

Fig. 6-9 Bifocal phase distributions (in degrees) to be implemented at 19.7 GHz on the sub-reflectarray (a) in X-pol. and (b) in Y-pol., and on the main reflectarray (c) in X-pol. and (d) in Y-pol.

The lengths of the dipoles are obtained in each reflectarray cell by iteratively calling an analysis routine that adjusts the lengths and analyzes the cell by a SD-MoM method assuming local periodicity to match the required phase-shift at each cell obtained by the bifocal technique, see Fig. 6-9. The variable l_{A2} is used to control X-polarization, while the variable l_{B2} controls Y-polarization. These variables can be separately adjusted thanks to the uncoupling of the phase response between orthogonal sets of dipoles [30]. In the design of the sub-reflectarray cells, the incidence angles from the central feed (F_3) have been taken into account, in order to minimize variations in the phase of the reflection coefficient when illuminating with a different feed. Similarly, the incidence angles from the virtual focus of the DRA system related to F_3 have been considered in the design of the main reflectarray cells.

6.2.5 Comparison with the equivalent single-focus antenna

Prior to the manufacturing of the demonstrator, some simulations have been performed by assuming ideal reflectarray cells in order to compare the beams from the BDRA demonstrator with those from the equivalent MDRA. The two DRA systems have been analyzed by applying the modular technique described in [70], which has been previously validated in the design of other dual reflectarray demonstrators [71]. A $\cos^q(\theta)$ function with $q = 43$ has been used to model the field radiated by the 60 mm feed-horns, which provides around -12 dB illumination on the edges of the sub-reflectarray. The comparison of the simulated radiation patterns in the xz -plane for the designed BDRA and the equivalent MDRA is shown in Fig. 6-10(a) for the beams in X-polarization, and in Fig. 6-10 (b) for the beams in Y-polarization.

Fig. 6-10 Simulated radiation patterns at 19.7 GHz in the xz -plane for the beams generated by the BDRA (solid lines) and by the equivalent MDRA (dashed lines): (a) in X-polarization, (b) in Y-polarization.

As can be seen, the bifocal beams present similar values of gain (a difference of 1.2 dB between the central beam and the beam produced by F_5 , which presents the lowest gain), beamwidth (around 2.6° at -3 dB) and SLL (around -23 dB with respect to the maximum). On the other hand, the MDRA provides a better performance for the central beam (-31 dB of SLL), but the extreme beams are quite broadened and the gain variation with respect to the central beam is around 1.5 dB. Moreover, the MDRA presents around 4.6° of separation between adjacent beams in the same polarization, while the BDRA is able to reduce this value to 3.8° for the same feed spacing. The ratio of beam spacing compression provided by the BDRA is $4.6^\circ/3.8^\circ = 1.2$.

Figure 6-11 shows the superposition of the radiation patterns in the xz -plane for the 10 beams generated by the BDRA in both X and Y polarizations, where the separation between adjacent beams in orthogonal polarizations is 1.9° .

Fig. 6-11 Simulated radiation patterns in the xz -plane for the beams in X and Y polarizations generated by the BDRA (ideal phases).

Figure 6-12 shows the amplitude distributions of the incident field on each reflectarray, considering illumination from the foci of the BDRA (F_1 and F_5), and from the central feed (F_3). As can be seen, the edge illumination levels on the sub-reflectarray are close to -12 dB for the three cases. The feed placed at F_1 produces a higher illumination on the upper part of the main reflectarray, while the feed placed at F_5 illuminates more its bottom part. In the latter case, the illumination levels reach -8 dB on the lower edge of the main reflectarray (the main reflectarray should be slightly oversized to reduce spillover).

Fig. 6-12 Amplitude (dB) of the incident field: (a) on the sub-reflectorarray produced by F_1 , (b) on the main reflectarray produced by F_1 , (c) on the sub-reflectorarray produced by F_3 , (d) on the main reflectarray produced by F_3 , (e) on the sub-reflectorarray produced by F_5 , (f) on the main reflectarray produced by F_5 .

Finally, Fig. 6-13 shows the superposition of the radiation patterns in the azimuth plane (the one orthogonal to the xz -plane in the direction of the beam) for the BDRA and the equivalent MDRA, corresponding to the beam produced by F_3 in X-polarization at 19.7 GHz. The patterns show a very good agreement, since the phases of the MDRA

have been used to obtain the values of the partial phase derivatives at the starting points of the bifocal algorithm. Therefore, the radiation patterns of the BDRA in the azimuth plane maintain the same characteristics than those of the MDRA.

Fig. 6-13 Simulated radiation patterns in the xz -plane for the beams in X and Y polarizations generated by the BDRA (ideal phases).

6.3 Manufacturing of the dual reflectarray demonstrator

The designed BDRA demonstrator has been manufactured and tested. Both sub- and main reflectarrays are made on a single layer of DiClad 880B. The dipoles are printed on the upper side of the substrate by conventional photo-etching process. Figure 6-14 shows the lateral view of the sandwich configuration for both reflectarrays.

Fig. 6-14 Sandwich configuration of both reflectarrays.

The dimensions of the reflectarray panels have been slightly increased, in order to allow an outer frame without dipoles that makes it possible to fix the antennas on the supporting plates with nylon screws. The dimensions of the main reflectarray panel are 570 x 510 mm (the main reflectarray size is 570 x 420 mm, as indicated in Table 6-1), while the sub-reflectarray panel has 405 x 427.5 mm (the sub-reflectarray size is 390 x

352.5 mm). The layouts with the dimensions of the dipoles are generated in AutoCAD, using the dipoles' lengths that were computed in the design of the demonstrator. The photo-etching masks of both sub- and main reflectarrays are shown in Fig. 6-15 and Fig. 6-16, respectively.

Fig. 6-15 Photo-etching mask for the sub-reflectarray and detail of the dipoles.

Fig. 6-16 Photo-etching mask for the main reflectarray.

The two reflectarrays have been manufactured and assembled on an aluminum structure, which also supports the 60-mm feed-horn from ANTERAL. Three different feed supports have been manufactured, in order to place the horn at positions F_1 , F_3 and F_5 given in Table 6-1 (the two foci of the bifocal antenna and the central feed). An AutoCAD scheme with the structure of the dual reflectarray demonstrator is shown in Fig. 6-17, with the feed placed at position F_1 . The resulting breadboard at the facilities of the Applied Electromagnetics Group of Universidad Politécnica de Madrid can be seen in Fig. 6-18.

Fig. 6-17 AutoCAD scheme with the structure of the BDRA demonstrator with the feed-horn placed at position F_1 .

Fig. 6-18 Manufactured BDRA demonstrator with the feed-horn placed at position F_5 .

6.4 Measurement of the dual reflectarray demonstrator and comparison with simulations

The BDRA prototype has been measured at the facilities of Universidad Politécnica de Madrid (UPM), in a compact-range measurement system. Figure 6-19 shows three pictures of the breadboard in the anechoic chamber of UPM with the horn placed at positions F_1 , F_3 and F_5 .

(a)

(b)

(c)

Fig. 6-19 Pictures of the BDRA demonstrator in the compact-range anechoic chamber with the feed-horn placed at: (a) position F_1 , (b) position F_3 and (c) position F_5 .

The comparison of the simulated and experimental radiation patterns at 19.7 GHz in the xz -plane for the beams produced in X and Y polarizations is presented in Fig. 6-20, considering illumination from F_1 , F_3 and F_5 . Then, Fig. 6-21 shows the comparison of the radiation patterns at 19.7 GHz in the azimuth plane for the beams produced in both polarizations when the feed is placed at F_1 (similar results are obtained for the beams generated from F_3 and F_5). The patterns include the co- and cross-polar components for each linear polarization. The simulations take into account the final design of the dual reflectarray antenna with the (3+3) dipole element, i. e., the tangential reflected field on each reflectarray cell has been computed by means of the aforementioned SD-MoM code. The effect of the outer frame without dipoles in both reflectarray panels has been also considered in the simulations. According to what was exposed in section 6.2.2, the field radiated by the horn has been modelled by a $\cos^q(\theta)$ function with $q = 43$ in the E-plane (12-dB beamwidth of 29°), which means in xz -plane for X-polarization and in azimuth plane for Y-polarization, and $q = 35$ in the H-plane (12-dB beamwidth of 32.2°), which means in xz -plane for Y-polarization and in azimuth plane for X-polarization. As can be seen in Fig. 6-20, a quite good agreement is achieved between simulations and measurements in the xz -plane, where the bifocal antenna is able to produce the beams in the required directions in both polarizations. In the case of the azimuth patterns (see Fig. 6-21), the main difference between measurements and simulations is the filling of the first nulls adjacent to the main lobe. An overview of the antenna performance at 19.7 GHz is shown in Table 6-2, which compares the simulated and experimental values of gain, SLL and XPD (measured within a 3-dB beamwidth) for the six beams generated from F_1 , F_3 and F_5 .

TABLE 6-2
COMPARISON OF MAIN ANTENNA PARAMETERS AT 19.7 GHz

Feed	Gain simul. (dB)	Gain meas. (dB)	SLL simul. (dB)	SLL meas. (dB)	XPD simul. (dB)	XPD meas. (dB)
Pol. X						
1	36.28	36.11	21.02	20.46	37.93	31.18
3	36.41	36.26	19.75	19.43	31.81	29.87
5	35.18	34.85	19.49	17.85	35.78	33.21
Pol. Y						
1	36.29	36.15	19.18	17.28	32.87	27.84
3	36.52	36.31	19.23	17.62	30.85	26.12
5	35.27	35.13	16.74	16.63	28.57	23.94

Fig. 6-20 Measured and simulated radiation patterns at 19.7 GHz in the xz -plane considering illumination from: (a) F_1 , (b) F_3 and (b) F_5 .

Fig. 6-21 Measured and simulated radiation patterns at 19.7 GHz in the azimuth plane for the beams produced by the feed at F_1 : (a) in X-polarization and (b) in Y-polarization.

The discrepancies between the measured and simulated radiation patterns of the bifocal antenna in the azimuth plane (filling of the first nulls adjacent to the main lobe) are mainly caused by the simplistic modelling of the horn, which has been simulated by a $\cos^q(\theta)$ distribution that considers the same position for the horn phase center in the two orthogonal planes. A shifting of the phase center in the azimuth plane (not taken into account in the simulations) would produce such effects on the antenna radiation patterns, which could be also worsened by phase errors due to tolerances in the dipole dimensions and in the dielectric constant of the substrate. A full-wave simulation of the feed-horn should be used to compute accurately the incident field on the sub-reflectarray for each feed.

Fig. 6-22 Measured radiation patterns in the xz -plane at the central and extreme frequencies of the 19.2-20.2 GHz band considering illumination from: (a) F_1 , (b) F_3 and (b) F_5 .

After validating the bifocal technique by comparison of the measured and simulated radiation patterns at 19.7 GHz (design frequency), the radiation patterns of the bifocal antenna in the xz -plane have been measured at the extreme frequencies of the 19.2-20.2 GHz band, in the cases where illumination from F_1 , F_3 and F_5 is considered. The results are shown in Fig. 6-22. It has been checked that some of these patterns present small deviations in the beam directions lower than $\pm 0.1^\circ$. Also, there is a gain variation lower than 1.2 dB, and around 2-3 dB higher SLL with respect to the patterns at the central frequency of the band.

The gain of the bifocal antenna has been measured within the prescribed band for the six beams generated from F_1 , F_3 and F_5 , and the results are presented in Fig. 6-23. The maximum gain variation is equal to or lower than 1.2 dB for all the beams, showing a steady response within the band. It can be noticed that the two beams produced from F_5 present around -1 dB lower gain than the beams from F_1 and F_3 , although this was expected after checking the results of the simulations with ideal phases (see Fig. 6-10). The reduction in gain for the beams associated to F_5 is produced because the illumination on the main reflectarray is moved to the reflectarray edge. To overcome this effect, the main reflectarray should be slightly oversized. The radiation efficiency of the bifocal antenna, calculated as the ratio between the measured gain and the maximum directivity, varies from 66% to 70% within the prescribed frequency band.

Fig. 6-23 Measured gain versus frequency for the six beams generated by F_1 , F_3 and F_5 in X and Y polarizations.

The breadboard has been also tested in a spherical near-field measurement system at UPM to compute 3D radiation patterns, in the case where the horn is placed at position

F_1 (see Fig. 6-24). The radiated field has been measured for both X and Y polarizations in the angular range $-80^\circ < \theta < 80^\circ$, $-180^\circ < \varphi < 180^\circ$. The simulated and measured radiation patterns in (u, v) coordinates at 19.7 GHz for the co-polar components of both polarizations generated from F_1 are in good agreement and can be seen in Fig. 6-25.

Fig. 6-24 Picture of the BDRA demonstrator in the spherical near-field measurement system.

Fig. 6-25 Radiation patterns (in dB) for the co-polar component produced by F_1 at 19.7 GHz: for X-polarization (a) simulated and (b) measured, and for Y-polarization (c) simulated and (d) measured.

The comparison of the simulated and measured radiation patterns in (u, v) coordinates at 19.7 GHz for the cross-polar components of both polarizations generated from F_1 is presented in Fig. 6-26. The patterns show very low levels of cross-polarization in all directions (no conversion from co-polar to cross-polar radiated fields), which makes it difficult to reach a very good accordance with the simulations.

Fig. 6-26 Radiation patterns (in dB) for the cross-polar component produced by F_1 at 19.7 GHz: for X-polarization (a) simulated and (b) measured, and for Y-polarization (c) simulated and (d) measured.

The superposition of the simulated and measured radiation pattern contours at -3 dB (respect to the maximum gain) for the two beams produced by F_1 at 19.7 GHz is presented in Fig. 6-27, and the measured -3 dB contours of the same beams at the central and extreme frequencies of the prescribed frequency band (19.2-20.2 GHz) are shown in Fig. 6-28. As can be seen, there is only a slight beam squint ($+0.09^\circ$) at the higher frequency. These results show that the bifocal antenna is able to produce two adjacent beams in dual-linear polarization for each feed. The beams can also be generated in dual-circular polarization if the orthogonal CP are discriminated by means of the variable rotation technique, as proposed in [59]. This concept can be used for the

design of a multi-beam antenna in Ka-band to provide adjacent beams in orthogonal polarizations with a smaller spacing than the equivalent single-focus antenna

Fig. 6-27 Comparison of the simulated and measured -3 dB pattern contours at 19.7 GHz for the beams produced from F_1 in X and Y polarizations.

Fig. 6-28 Measured -3 dB contours at the central and extreme frequencies of the prescribed band for the beams produced from F_1 in X and Y polarizations.

6.5 Conclusions

A bifocal dual-reflectarray antenna demonstrator with a main reflectarray of dimensions 57 x 42 cm has been designed, manufactured and tested for the first time. The prototype has been designed at 19.7 GHz to operate in the 19.2-20.2 GHz band (transmission frequencies from the satellite in Ka-band). Due to the large offset of the antenna geometry, the calculation of the required phase-shift distributions on each reflectarray has been performed by means of the 3D bifocal method presented in Chapter 5. Two separate bifocal design processes have been carried out, one for each

linear polarization. A single-layer reflectarray element, consisting of two orthogonal sets of three parallel dipoles, has been used to implement the required phase-shifts on both sub- and main reflectarrays by properly adjusting the lengths of the parallel dipoles. The BDRA is designed to provide 10 beams alternating in dual-linear polarization when the antenna is illuminated by 5 feed-horns, but the technique can be used to generate adjacent beams in dual-circular polarization by using a sequential rotation method.

The measured patterns show both a good agreement with the simulations and a steady response within the prescribed frequency band (19.2-20.2 GHz), with a maximum gain variation of 1.2 dB. The bifocal antenna provides the required beam directions, with 3.8° of separation between beams generated by contiguous feeds in the same polarization (1.9° between adjacent beams in orthogonal polarizations). The side-lobe levels are lower than -17 dB below the co-polar maximum, while the cross-polar discrimination varies from 24 to 34 dB. There are slight discrepancies between the measured and simulated radiation patterns in the azimuth plane, which are a consequence of the use of a simplistic electromagnetic model for the horn. However, these discrepancies do not have any effect on the antenna performance in the xz -plane (the plane that contains the radiated beams), which is determined by the bifocal technique. The BDRA demonstrator has validated the capabilities of the bifocal technique to improve the multi-beam performance in two directions:

- 1) Reduction of the beam spacing in a certain degree (in this case, by a factor of 1.2) with respect to the equivalent monofocal antenna, with a similar radiation efficiency (close to 70%) and using non-overlapping feeds. The smaller beam spacing will make it possible a reduction in the required antenna size with respect to conventional reflectors for the same beam spacing.
- 2) Improvement of the beams far away of the focal position, which present better results for the gain (+0.3 dB), SLL (-2 dB) and 3-dB beamwidth (2.6° , similar to the other beams) than in the equivalent monofocal antenna. A further optimization could be carried out in order to fulfill additional SLL or XPD requirements for current multi-beam satellite antennas in Ka-band, although this optimization should not disturb the phase distributions obtained by the bifocal technique.

Furthermore, the manufacturing of the BDRA involves the same conventional photo-etching processes used for printed reflectarrays, as the only difference with respect to the equivalent MDRA are the dimensions of the printed elements. This will make it possible to improve the multi-beam performance of current satellite antennas in Ka-band (when designed for a single focus) without increasing the costs of the antenna system. In the case of bifocal dual reflectors, both reflectors are shaped as a consequence of the bifocal design process, thus requiring expensive metallic moulds that should be manufactured specifically for each mission. But this problem is avoided using the proposed bifocal dual-reflectarray antenna.

Finally, the proposed concept can be extended to the design of transmit and receive antennas in Ka-band by the use of dual-frequency reflectarray cells that enable independent phasing at Tx (20 GHz) and Rx (30 GHz) frequencies (as those presented in Chapter 2). In this case, independent bifocal processes will be carried out for each frequency band, in order to obtain the required phase distributions to be implemented by the reflectarray elements.

Chapter 7

Bifocal antenna with elliptical main reflectarray for multi-spot coverage in Ka-band

7.1 Introduction

This chapter describes the design of a multi-beam dual reflectarray antenna to provide multi-spot coverage for transmission from a geostationary satellite operating in Ka-band (20 GHz). The beam spacing in the offset plane is reduced by a factor of 1.8 with respect to the equivalent monofocal antenna, thanks to the use of the bifocal technique. This allows to provide the required 0.56° separation between adjacent beams, but as shown in Chapters 3 and 5, the high degree of beam spacing compression forces to oversize the main reflectarray in the same dimension where the beams are going to be compressed, in order to ensure low spillover and good radiation efficiency. Therefore, an elliptical main reflectarray of dimensions 3.55×1.8 m is considered.

The radiation patterns of the multi-beam antenna combine the bifocal characteristic in the xz -plane with a monofocal characteristic in the orthogonal plane, where the angular separation between adjacent beams is around 1.1° . The interleaved beams for providing full multi-spot coverage are generated in the orthogonal polarization, assuming that appropriate reflectarray cells that allow independent control of the phase-shift introduced in each polarization (as those used in Chapter 6 for the design of the bifocal antenna demonstrator) are employed.

As a result of using an elliptical main aperture, the designed antenna produces elliptical beams, which are narrower in the dimension where the beams are compressed by the bifocal technique. For this reason, a preliminary study will be carried out to

broaden the beams in the offset plane in order to obtain circular spots, by introducing a quadratic phase adjustment in the bifocal phase distribution on the main reflectarray.

Finally, the multi-beam performance of the proposed bifocal antenna with the elliptical main reflectarray and polarization discrimination will be compared with that of an oversized shaped reflector shown in [90]. The oversized reflector (around 4.5 m in diameter) provides all the beams required for multi-spot coverage in Ka-band by using a SFPB scheme with a large focal distance, but it has to be shaped to produce wider spots. On the other hand, the proposed bifocal antenna requires a smaller main aperture (an elliptical instead of a circular one) and the use of flat printed reflectarrays will allow for a relatively inexpensive fabrication and an easier deployment on the satellite.

7.2 Design of a bifocal dual reflectarray antenna to provide multi-spot coverage in Ka-band

A Cassegrain configuration has been selected for the design of the bifocal dual reflectarray antenna (see Fig. 7-1), which consists of an elliptical main reflectarray (3.55 x 1.81 m) and a circular sub-reflectorarray (79 x 79 cm), with a relative tilting of 10° between them. The main geometrical parameters of the dual reflectarray antenna (DRA) system are summarized in Table 7-1. A cell period of 10 mm is considered for both reflectarrays in order not to work with an excessive number of reflectarray elements; however, in a realistic design the period should be lower (for example, around 7.5 mm, which is $\lambda/2$ at 20 GHz), in order to avoid grating lobes and provide larger bandwidth.

Fig. 7-1 Geometry of the DRA configuration with an elliptical main reflectarray.

TABLE 7-1
MAIN GEOMETRICAL PARAMETERS OF THE DRA SYSTEM

Parameter	Value
Size Main-RA	3.55 x 1.81 m (355 x 181 elements)
Size Sub-RA	79 x 79 cm (79 x 79 elements)
Angle of tilting Sub-RA	-10°
Center Main-RA	[-65, 0, 0] mm
Center Sub-RA	[1288, 0, 3003] mm
Phase center F_1 (focus 1 BDRA)	[-2043, 0, 2222] mm
Phase center F_5 (focus 2 BDRA)	[-1830, 0, 2185] mm
Distance F_1 - F_5	220 mm
Phase center F_3 (focus MDRA)	[-1937, 0, 2204] mm
Virtual focus related to F_3	[-1854, 0, 4847] mm

An array of five contiguous feed-horns (from F_1 to F_5) placed in the xz -plane is initially considered for the illumination of the antenna. The horns present 54 mm diameter and the same characteristics than those used in Chapters 3 and 5 [82]. They provide around -12 dB on the sub-reflector edges at 20 GHz for a subtended angle of 36°. A $\cos^q(\theta)$ distribution with $q = 28$ is used to model the electromagnetic field radiated by the horns. The separation between the phase centers of adjacent feeds is set to 55 mm, allowing 1 mm margin to accommodate the horns.

The reason for choosing a Cassegrain instead of a compact-range configuration (as the one used for the design of the demonstrator in Chapter 6) is because it allows to oversize the main reflectarray without producing blockage from the feeds or the sub-reflector. For this purpose, the antenna geometry has been adjusted after carrying out a trade-off with the bifocal algorithm, in order to ensure an appropriate illumination on the main reflectarray when applying the bifocal technique to obtain adjacent beams (0.56° separation in the xz -plane), at the same time as minimizing blockage from the sub-reflector. In this case, starting from the geometry of a reference single-focus antenna can lead to a non-centered illumination on the main reflectarray when the bifocal technique is applied.

Also, the relative tilting between the two reflectarrays has been adjusted to provide the smoothest variation in the required phase distributions for both reflecting surfaces. As shown in Chapter 3, the design with parallel reflectarrays leads to a larger number of 360° cycles in the phase-shift distributions of both reflectarrays, which makes more

difficult their practical implementation and reduces the potential bandwidth of the antenna.

7.2.1 *Reference single-focus antenna*

A single-focus reference design has been performed at 20 GHz for the previous DRA geometry, considering the central feed (F_3) to radiate at $(\theta_{b3} = 19^\circ, \varphi_{b3} = 0^\circ)$ with respect to the normal vector to the main reflectarray surface, $\hat{z}_M$ (see Fig. 7-1). The purpose of this design is to achieve around 1° separation between adjacent beams in the azimuth plane, in order to use the monofocal phases for performing a bifocal design with 0.56° beam spacing in the xz -plane. The required phase-shift distributions on each reflectarray associated to this monofocal design, which have been calculated according to the method described in [71], can be seen in Fig. 7-2.

Fig. 7-2 Required phase-shift distributions (in degrees) for the monofocal antenna on the (a) sub-reflector array and on the (b) main reflectarray.

The simulated radiation patterns at 20 GHz for the five beams produced by the monofocal DRA with the elliptical main reflectarray both in elevation (xz -plane) and azimuth (orthogonal plane in the direction of the beam) planes are shown in Fig. 7-3. Note that there are no cross-polar components in the radiation patterns, as the design has been performed considering ideal reflectarray elements. The angular spacing between adjacent beams in the xz -plane is between 0.9° - 1° for the beams radiating at the right of the central one, and between 1° - 1.1° for the beams at the left. The gain of the beams varies between 46.03 dBi and 48.5 dBi. As can be seen, the extreme beams are considerably broadened and defocused, because the corresponding feeds are placed out of the antenna focus.

Fig. 7-3 Simulated radiation patterns for the monofocal DRA at 20 GHz: (a) superposition of cuts in the azimuth plane, and (b) cut in the xz -plane.

The amplitude distributions (dB) of the incident field on the two reflectarrays produced by the extreme feeds of the array, F_1 and F_5 , are shown in Fig. 7-4. As can be seen, the module of the electric field is close to or lower than -12 dB on the edges of the sub-reflectarray, in order to reduce spillover. On the other hand, the main reflectarray has been significantly oversized along its vertical dimension, so only a small portion is illuminated above -12 dB.

Fig. 7-4 Amplitude (dB) of the incident field on the sub-reflectarray when the antenna is illuminated from (a) F_1 and (b) F_5 , and on the main reflectarray for illumination from (c) F_1 and (d) F_5 .

A conventional single-focus antenna with the same characteristics would use a circular main reflectarray, around 1.8 m in diameter, instead of an elliptical one. The subsequent application of the bifocal technique in the xz -plane with a high degree of beam spacing compression is the reason for oversizing the main reflectarray. The monofocal DRA design will not be implemented in the practice, but it is necessary for obtaining the bifocal phases required on each reflectarray.

7.2.2 *Bifocal antenna with high beam spacing compression*

The bifocal technique has been applied to the current DRA geometry in order to obtain 0.56° separation between adjacent beams in the xz -plane, which implies reducing the beam spacing by a factor of $1^\circ/0.56^\circ = 1.8$ with respect to the equivalent monofocal antenna shown in the previous section. The phase centers of the extreme feeds (F_1 and F_5) have been selected as the foci of the bifocal antenna. The directions of the beams associated to the foci are $(\theta_{b1} = 20.12^\circ, \phi_{b1} = 0^\circ)$ and $(\theta_{b5} = 17.88^\circ, \phi_{b5} = 0^\circ)$. The bifocal design algorithm has been used to obtain the required phases for both reflectarrays, considering a monofocal phase condition in the orthogonal plane for the starting points, as in the bifocal designs performed in Chapter 5. The resulting bifocal phase-shift distributions can be seen in Fig. 7-5. Note that the rotation of a 2D bifocal design in an axially-symmetrical configuration would produce a bifocal antenna with a focal ring and the same beam spacing (0.56°) in the two orthogonal planes, as in the designs shown in Chapter 3, but it would require an oversized circular main reflectarray (instead of an elliptical one) which would produce narrower spots.

Fig. 7-5 Required phase-shift distributions (in degrees) for the bifocal antenna on the (a) sub-reflectorarray and on the (b) main-reflectorarray.

The simulated radiation patterns in the principal planes (elevation and azimuth) for the designed bifocal DRA are shown in Fig. 7-6. As can be seen, the separation between beams generated by contiguous feeds in the xz -plane is now 0.56° . There are some small pointing errors (0.05° - 0.08°) in the central beams that can be corrected in a more detailed design of the antenna by adjusting the feed positions. The gain varies from 49.65 dBi (for the beam produced from F_1) to 50.44 dBi (for the beam produced from F_4). The SLL is lower than -21 dB with respect to the maximum for all the beams. The radiation patterns in the azimuth plane present similar characteristics than those shown in Fig. 7-3(a) for the monofocal antenna.

Fig. 7-6 Simulated radiation patterns at 20 GHz for the bifocal DRA to provide 0.56° separation between adjacent beams: (a) superposition of cuts in the azimuth plane, and (b) cut in the xz -plane.

A comparison between the beams generated by the single-focus antenna and those produced by the bifocal antenna is shown in Fig. 7-7. As can be seen, the bifocal design allows to obtain much closer beams (beam spacing is reduced by a factor of 1.8) with a well-shaped main lobe and similar values for the beamwidth and SLL. The gain of the bifocal beams is also some dB higher than in the monofocal antenna, since the use of an oversized main reflectarray provides a better illumination from all the feeds.

The amplitude distributions of the incident field on the two reflectarrays produced by the extreme feeds (F_1 and F_5) are shown in Fig. 7-8. The module of the incident field is close to -12 dB on the edges of the sub-reflectorarray, while elliptical illuminations are obtained on the main reflectarray with higher levels than those shown in Fig. 7-4 for the monofocal antenna. This is a consequence of applying the bifocal technique with a high degree of beam compression, which spreads the illumination on the main reflectarray along its vertical axis, while the illumination along the horizontal axis is determined by the monofocal design (adjusted for a 1.8 m width of the main reflectarray).

Fig. 7-7 Comparison of the beams generated by the bifocal antenna with $BCR = 1.8$ (solid lines) and the beams generated by the single-focus antenna (dashed lines).

Fig. 7-8 Amplitude (dB) of the incident field on the sub-reflectorarray when the antenna is illuminated from (a) F_1 and (b) F_5 , and on the main reflectarray for illumination from (c) F_1 and (d) F_5 .

Due to the elliptical illumination obtained on the main reflectarray, the bifocal antenna generates elliptical beams, which present a different beamwidth in each of the principal planes. This can be observed in the patterns shown in Fig. 7-6, where the beamwidth at 46 dBi gain for the central beam is $0.67^\circ \times 0.39^\circ$ (azimuth $\times$ elevation). The elliptical shape of the beams can also be checked in the radiation pattern in (u, v) coordinates at 20 GHz of the central beam, shown in Fig. 7-9.

Fig. 7-9 Simulated radiation pattern in (u, v) coordinates for the central beam produced by the bifocal antenna at 20 GHz.

The bifocal antenna provides the required degree of beam spacing compression only in the xz -plane, while it maintains the monofocal characteristic of the original design in the orthogonal plane. To illustrate this fact, two additional beams have been obtained in the azimuth plane at both sides of the central beam, using two 54 mm feed-horns placed adjacent to F_3 in the direction of y -axis (which are named F_{3L} and F_{3R}). The simulated radiation patterns at 20 GHz in the azimuth plane for the three beams are shown in Fig. 7-10. As can be seen, the separation between adjacent beams is around 1.1° , which is almost twice the spacing enforced by the bifocal technique in the offset plane.

The amplitude distributions of the incident field on the main reflectarray produced by the feeds F_{3R} and F_{3L} are shown in Fig. 7-11. The illumination is moved to the lateral edges of the main reflectarray, whose size along the horizontal dimension could be slightly increased in order to maintain low spillover and ensure around -12 dB illumination levels on the edges.

Fig. 7-10 Simulated radiation patterns at 20 GHz in the azimuth plane for the bifocal antenna, considering the central feed (F_3) and two additional feeds adjacent to the central one.

Fig. 7-11 Amplitude (dB) of the incident field on the main-reflectorarray when the antenna is illuminated from (a) F_{3L} and (b) F_{3R} .

The designed bifocal antenna has been used to produce multiple spots at 20 GHz, considering the cluster of horns which is shown in Fig. 7-12(a). This cluster includes the five initial feeds, from F_1 to F_5 , and ten additional feeds placed adjacent to the previous ones in the direction of y -axis. The simulated pattern contours of 40 dBi and 47.5 dBi (which is around -3 dB respect to the maximum gain) for the beams produced by bifocal antenna at 20 GHz are shown in Fig. 7-12(b). As can be seen, the bifocal antenna is able to generate adjacent beams with 0.56° spacing in $v = \text{constant}$ planes. The interleaved beams in $u = \text{constant}$ planes that are required for providing the multi-spot coverage will be generated in the orthogonal polarization, using the capability of reflectarrays to discriminate in polarization, as will be shown in the next section.

Fig. 7-12 Generation of multiple spots: (a) cluster of horns used to illuminate the bifocal antenna, including the initial feeds (from F_1 to F_5), (b) simulated pattern contours of 40 dBi and 47.5 dBi at 20 GHz for the beams produced by the bifocal antenna.

7.2.3 *Bifocal antenna to provide multi-spot coverage in dual polarization*

The discrimination of the orthogonal polarizations will be performed on the main reflectarray, assuming that appropriate reflectarray cells that allow independent control of the phase in each polarization are used for the design of the antenna. Two different phase-shift distributions will be implemented on the main reflectarray, one for each polarization, while the sub-reflectorarray will present the same phase distribution for both polarizations (the one shown in Fig. 7-5(a)). A cluster of dual polarized feed-horns will illuminate the bifocal antenna, so that each feed will produce two adjacent beams in orthogonal polarizations in a 60° lattice with respect to the xz -plane. The proposed

concept is valid for operation in both dual-linear and dual-circular polarization. In the first case, the reflectarray element can be similar to the one used in Chapter 6 for the design of the demonstrator. To generate the beams in dual-circular polarization, either the reflectarray cell shown in [56] or a sequential rotation of the reflectarray elements [59] can be used.

A simple way of obtaining the phase-shift distribution for the orthogonal polarization on the main reflectarray is by adding a progressive phase term to the bifocal phases shown in Fig. 7-5(b). The required phase increment at each reflectarray cell can be calculated by means of the following expression, which provides the phase distribution of the reflected field on the reflectarray surface to generate a collimated beam in the direction (θ_b, φ_b) :

$$\Phi(x_i, y_i) = -k_0 \sin \theta_b (\cos \varphi_b x_i - \sin \varphi_b y_i) \quad (7-1)$$

where x_i and y_i are the coordinates of the element with respect to the geometrical center of the reflectarray. The previous expression has been adapted to work with the (u_b, v_b) coordinates associated to the beam direction:

$$\Phi(x_i, y_i) = -k_0 (u_b x_i - v_b y_i) \quad (7-2)$$

Then, the difference in the required phase between two beams whose maximums are located at coordinates (u_{b1}, v_{b1}) and (u_{b2}, v_{b2}) will be:

$$\Delta\Phi_{2,1}(x_i, y_i) = -k_0 [(u_{b2} - u_{b1}) x_i + (v_{b2} - v_{b1}) y_i] \quad (7-3)$$

Therefore, the increment of phase that must be added to the bifocal phase distribution on the main reflectarray to generate the interleaved beams in the orthogonal polarization can be calculated by using (7-3), and considering the (u, v) coordinates associated to the maximums of the beams produced by the central feed (F_3) in the two polarizations. According to the patterns shown in Fig. 7-12(b), the maximum of the central beam is located at $(u_{b1} = 0.325, v_{b1} = 0)$, while the position of the adjacent beam in the orthogonal polarization to form a 60° lattice with respect to the xz -plane must be $(u_{b2} = 0.332, v_{b2} = 0.1)$. The resulting progressive phase distribution to be added to the phases in the original polarization is presented in Fig. 7-13. Then, Fig. 7-14 shows the final phase-shift distributions for both polarizations that have to be implemented on the main reflectarray.

Fig. 7-13 Increment of phase required for the orthogonal polarization with respect to the initial polarization to produce adjacent beams in a 60° lattice.

Fig. 7-14 Required phase-shift distributions on the main reflectarray at 20 GHz for: (a) the initial polarization, and (b) the orthogonal polarization.

The pattern contours of 40 dBi and 47.5 dBi for the beams produced by the bifocal antenna in both polarizations at 20 GHz are presented in Fig. 7-15, where illumination from the same cluster of feeds (now operating in dual polarization) shown in Fig. 7-12(a) is considered. After adding the beams in the orthogonal polarization, slightly-overlapping elliptical spots are obtained. Thus, the bifocal antenna is able to provide multi-spot coverage for transmission in Ka-band using elliptical beams arranged in a 60° lattice with a minimum end-of-coverage (EOC) gain of 40 dBi, as can be seen in Fig. 7-16. The separation between the maximums of adjacent spots is around 0.56° , which is the required value for current multi-spot applications in Ka-band. The beams are generated by contiguous and non-overlapping feed-horns, thanks to the application of

the bifocal technique in the xz -plane and the discrimination of the orthogonal polarizations at the reflectarray element level.

Fig. 7-15 Pattern contours of 40 dBi and 47.5 dBi at 20 GHz for the beams produced by the bifocal antenna in the two polarizations.

Fig. 7-16 Multi-spot coverage provided by the bifocal antenna.

Therefore, the dual reflectarray configuration with the elliptical main reflectarray (3.55 x 1.81 m) allows to ensure proper illumination from the feeds after applying the bifocal technique, at the same time as using a single oversized main aperture to generate the multi-spot coverage in transmission in Ka-band (although the aperture size is smaller than in other oversized antennas for the same purpose [90], as will be addressed later). Moreover, each feed generates two beams, which implies a 50% saving in the number of feeds with respect to a conventional SFPB system.

The use of dual-frequency reflectarray cells designed to operate both at Tx (20 GHz) and Rx (30 GHz) frequencies in Ka-band will make it possible to implement independent phase distributions at each frequency band. In that case, the same design process (bifocal technique plus operation in dual polarization) can be applied to produce multi-spot coverage for Rx in Ka-band. This will allow to reduce the number of antennas required to provide the coverage (typically four reflectors) to only one moderately-oversized antenna operating both in transmission and reception.

7.2.4 Broadening of the beams

As shown in the previous sections, the beams produced by the bifocal antenna with the elliptical main reflectarray are narrower in elevation than in azimuth, as a consequence of applying the bifocal technique in the xz -plane with a monofocal phase condition in the orthogonal dimension. A preliminary study has been performed to try to broaden the beams in elevation, in order to obtain circular beams with around 0.65° beamwidth for a minimum gain level of 44 dBi. This would improve the minimum EOC gain of the multi-spot coverage shown in Fig. 7-16, which is around 40 dBi.

For this purpose, the phase-shift distribution on the main reflectarray has been modified by adding a quadratic phase term of the form $K \cdot (\Delta x/D)^2$, where Δx represents the position of the reflectarray elements along the vertical axis with respect to the center of the reflectarray, D is the size of the main reflectarray (3.55 m) and K is a constant. The quadratic phase adjustment will produce a certain effect of beam defocusing in the elevation plane, which will result in a broadened main lobe, higher side-lobes levels and lower maximum gain for the beam.

The study has been carried out considering the beam produced by F_1 which points at ($\theta_{b1} = 20.12^\circ$, $\phi_{b1} = 0^\circ$). The simulated radiation patterns at 20 GHz in the xz -plane for

this beam after performing different quadratic phase adjustments are shown in Fig. 7-17(a). Note that $K = 0$ corresponds to the original radiation pattern, without any adjustment in the phase distribution. The figure includes a mask with the desired 0.65° beamwidth at 44 dBi gain. Moreover, a maximum level of side-lobes that is 20 dB below the EOC gain has been considered to ensure a reasonable value of the single-entry C/I (interference produced with the adjacent beam in the same colour).

Fig. 7-17 Beam broadening: (a) simulated radiation patterns at 20 GHz in the xz -plane for the beams produced from F_1 with different quadratic phase adjustments (b) enlarged view of the beams.

As can be seen in Fig. 7-17(b), which shows an enlarged view of Fig. 7-17(a), the defocusing of the main lobe leads to a broadened beam under 44 dBi gain, but the beamwidth associated to that gain level remains lower than 0.65° . Using larger values of K in the quadratic phase adjustment will lead to an important reduction in gain, as well

as to violate the 20 dB requirement of single-entry C/I between adjacent beams of the same color. Therefore, a more sophisticated procedure is required in order to broaden the beams, while trying at the same time not to disturb the bifocal phase distributions that provide the required beam compression. For example, a possible solution would be to apply phase-only synthesis based on the intersection approach technique [132], [133] to broaden the beam and reduce the levels of side-lobes only in one plane.

7.3 Comparison with an oversized shaped reflector

A performance comparison has been carried out between the designed bifocal antenna with the elliptical main reflectarray and a 4.5 m optimized reflector developed within a research activity funded by the European Space Agency (ESA). The information about the performance of the oversized reflector has been extracted from [90], where a 4.5 m diameter shaped reflector has been designed to provide multi-spot coverage for transmission in Ka-band using a large focal length and a SFPB architecture. The highly-oversized aperture is required in order to ensure low spillover with non-overlapping feeds, while the shaping is used to obtain wider beams. The concept of the oversized reflector can be extended to provide multi-spot coverage for both Tx and Rx in Ka-band and has given rise a patent [91].

The results for the comparison of the two antennas are presented in Table 7-2. The maximum and minimum directivity ($D_{\max}$ and $D_{\min}$) in the coverage area, considering a 0.65° beamwidth for the spots, and the peak side-lobe level ($SL_{\max}$) are obtained for the central beam and the extreme beams of the coverage, considering three beams away from the central one in the elevation plane. The simulated radiation patterns at 20 GHz for the seven beams generated by the bifocal antenna in the xz -plane are shown in Fig. 7-18. The peak side-lobe level has been calculated considering the next beam in the same colour (which is not the adjacent beam at 0.56° , but the beam placed at 1.12° separation, as shown in Fig. 7-19).

As can be seen in Table 7-2, the maximum gain provided by the 4.5 m reflector is around 1-2 dB higher than the gain of the spots produced by the bifocal antenna, although this is a consequence of its larger aperture size. Also, the oversized reflector has been optimized (by a specific shaping of the parabolic surface) to fulfill the specification concerning the EOC gain, which is equal to 44.09 dBi in the worst case.

On the other hand, the beams from the bifocal antenna are elliptical, so the minimum directivity in the coverage area is around 40 dBi, associated to a 0.65° beamwidth in the elevation plane. Note that the minimum gain in the azimuth plane of the bifocal beams is also shown in the table, where around 45-46 dBi gain is reached for a 0.65° beamwidth (the beams provided by the bifocal antenna are wider in azimuth than in elevation). Finally, the bifocal antenna provides a better performance in terms of peak SLL for all the beams, which present between 4 and 6 dB lower SLL than the broadened beams from the 4.5 m oversized reflector.

TABLE 7-2
COMPARISON BETWEEN THE OVERSIZED REFLECTOR AND THE BIFOCAL DRA

Beam	D _{max} (dBi)	D _{min} (dBi)		SL _{max} (dBi)
4.5 m reflector				
Central	51.21	47.06		24.05
Worse case	51.21	44.09		27.65
3.55 x 1.8 m reflectarray				
		Elevation	Azimuth	
Central	50.41	41.02	46.31	20.31
Ext – F0	48.92	40.32	45.63	20.66
Ext – F6	50.06	40.16	46.58	21.24

Fig. 7-18 Simulated radiation patterns at 20 GHz in the xz -plane for the bifocal antenna to provide 0.56° separation between adjacent beams.

Fig. 7-19 Simulated radiation patterns in the elevation plane for the central beam and the adjacent beams in the same colour (1.12° separation), with peak side-lobe levels.

The proposed dual reflectarray antenna represents an important reduction in the required size of the aperture to provide multi-spot coverage in Ka-band, since it uses an elliptical reflectarray of 3.55×1.81 m instead of a circular reflector with 4.5 m diameter (whose area is roughly equivalent to the total area of the current state-of-the-art four reflector configuration [82], [83]). Moreover, the application of the bifocal technique provides lower levels of side lobes (i. e., lower interference with the adjacent beams) than in the case of the oversized reflector. Although the oversized reflector has been optimized to fulfill a specific requirement for the EOC gain, the performance of the bifocal antenna can be also improved by means of phase optimization techniques, in order to increase the minimum EOC gain for the coverage. Furthermore, the shaping of the oversized reflector leads to a more complex and expensive fabrication, as it requires of custom metallic moulds that must be manufactured specifically for this mission. On the other hand, the proposed bifocal antenna consists of two flat reflectarray panels, which can be fabricated by conventional photo-etching and processes used in multi-layer printed circuits, and allows for a more efficient packaging and deployment on the satellite.

7.4 Conclusions

A multiple spot beam satellite antenna based on a dual reflectarray configuration with an elliptical flat main reflectarray (3.55×1.81 m) has been proposed to provide multi-spot coverage for transmission in Ka-band. The bifocal technique has been

applied with a high degree of beam spacing compression (by a factor of 1.8) to produce adjacent beams with 0.56° separation in the offset plane, while using a monofocal phase condition in the orthogonal plane, where the beam spacing is around 1.1° . The main reflectarray has been oversized in the same dimension where the beams are compressed (thus resulting in an elliptical reflectarray), so as to ensure low spillover and proper illuminations from the feeds.

The interleaved beams that are required in the azimuth plane for providing the full coverage are generated in the orthogonal polarization, considering that the reflectarray elements make it possible an independent control of the phase-shift introduced in each polarization. The discrimination of the orthogonal polarizations has been implemented on the main reflectarray. The required phase-shift distribution for the orthogonal polarization has been obtained by adding a progressive phase term to the bifocal phases for the initial polarization, in order to produce adjacent beams in a 60° lattice with respect to the xz -plane. This concept is valid for both dual-linear and dual-circular polarizations.

The designed bifocal antenna that operates in dual polarization is able to provide multi-spot coverage for transmission in Ka-band with slightly-overlapping elliptical spots (produced as a result of using an elliptical main reflectarray) in a 60° lattice. The angular spacing between adjacent beams is 0.56° , which fulfills the requirements for current multi-spot applications in Ka-band, and the minimum EOC gain is around 40 dBi. Thanks to the discrimination of the orthogonal polarizations by the same antenna, the number of feeds is reduced by a 50% with respect to a conventional SFPB system (now the feeds must operate in dual polarization).

The minimum EOC gain of the multi-spot coverage can be improved by applying a phase correction technique to broaden the beams in the elevation plane, in order to obtain circular spots. However, the solution to this problem is not trivial. As has been shown, a quadratic phase adjustment does not provide wider beams, unless reducing considerably the maximum gain of the beams and increasing the interference levels with the adjacent beams. Phase-only synthesis based on the intersection approach technique could be applied to broaden the beams and reduce the SLL in one plane, without disturbing the bifocal phases that provide the required degree of beam spacing reduction.

The performance of the bifocal antenna with the elliptical main reflectarray has been compared with that of an oversized shaped reflector shown in [90]. The proposed bifocal antenna is able to provide the multi-spot coverage using a smaller main aperture (elliptical reflectarray of 3.55 x 1.81 m instead of a circular reflector with 4.5 m of diameter), with lower levels of side lobes thanks to the application of the bifocal technique. The oversized reflector has been shaped to provide a larger EOC gain than the bifocal antenna, although this results in higher manufacturing costs. On the other hand, the performance of the bifocal antenna can be improved by applying phase optimization techniques in order to broaden the beams in the elevation plane (thus increasing the EOC gain), while maintaining the same fabrication processes used for printed reflectarrays, without any increment in the manufacturing time and cost.

Finally, the design of a Tx/Rx multiple spot beam satellite antenna can be addressed by the use of appropriate dual-frequency reflectarray cells that will enable independent phasing at Tx and Rx frequencies (20 and 30 GHz) in Ka-band. This will allow for performing a similar design process for the Rx band, applying the bifocal technique to produce adjacent beams with 0.56° spacing in the offset plane, and then, generating the beams in the orthogonal polarization in a 60° lattice. In this case, the proposed bifocal antenna with a moderately-oversized main aperture (an elliptical reflectarray of 3.55 x 1.81 m) would be able to provide multi-spot coverage for both Tx and Rx in Ka-band, leading to an important reduction in the number of antennas required with respect to the four reflector configuration [83], which is the current state of the art for multiple spot beam antennas in satellite systems operating in Ka-band.

Chapter 8

Conclusions and future work

8.1 Conclusions

The motivation of this thesis has been to provide new advances on the design of multi-frequency and multi-beam reflectarray antennas with application to multiple spot beam satellites in Ka-band. In this respect, the thesis can be divided into two main parts. The first part contains the description of a novel reflectarray cell to operate in dual-linear polarization (dual-LP) at two separate frequencies (enabling independent phasing in each polarization and frequency), as well as the design of dual-band reflectarrays to provide independent beams in each polarization at both frequency bands. The second part of the thesis comprises the development of a bifocal design technique for dual reflectarray and dual transmitarray configurations, and its application to the design of multi-beam antennas in Ka-band. The aim of the bifocal technique is twofold, to improve the multi-beam performance of the antenna and to provide a certain degree of reduction on the angular separation between adjacent beams for a multi-spot coverage from a satellite. The main conclusions of the work are summarized below.

The reflectarray element proposed for independent operation in dual-LP at two separate frequencies consists of a two layer configuration with two orthogonal sets of stacked parallel dipoles. Each set is composed of five parallel dipoles on the lower layer, and three additional parallel dipoles which are stacked above the previous ones and are printed on the top of a second dielectric sheet. The geometrical parameters of the cell have been adjusted to operate, first, at Tx frequencies in Ku and Ka bands (12 and 20 GHz), and then, at Tx and Rx frequencies in Ka-band (20 and 30 GHz). The proposed two-layer configuration allows to perform separate design processes for each

reflectarray layer: first, the lengths of the lower dipoles are adjusted to match the required phases at the lower frequency, and then, the lengths of the upper dipoles are adjusted to introduce the required phase-shift at the higher frequency.

A 33 cm reflectarray antenna has been designed with the proposed cell configuration to produce independent beams in each linear polarization at 12 and 20 GHz, and then, a 1.6 m Ka-band reflectarray has been shown to provide adjacent beams in orthogonal linear polarizations both at Tx and Rx bands (20 and 30 GHz). Furthermore, the proposed element has been used to design a 25-cm reflectarray demonstrator for simultaneous operation at Ku (11-13 GHz) and Ka (19-20 GHz) bands in dual polarization (linear or circular), in order to validate both the element and the design procedure. The measurements of the manufactured demonstrator are in good agreement with the simulations in Ku-band, while presenting some discrepancies in Ka-band, due to variations in the electrical properties of the dielectric sheets (this problem can be avoided by an accurate characterization of the materials before performing the antenna design).

The proposed concept for the reflectarray element can be applied to design a reflectarray which is able to fulfill independent requirements at each frequency and/or polarization (for example, generation of a contoured beam in Ku-band and multiple spots in Ka-band), using different feed chains for each mission. The reduced number of layers and the simplicity of the elements will allow for an easier manufacturing and low profile of the resulting antenna. In the case of current satellite systems that operate in Ku and Ka bands, the reuse of the same aperture for two different missions would result in significant savings in the costs, weight and volume of the antenna farm.

A bifocal design procedure has been developed for dual reflectarray antennas in offset configurations, starting from an axially-symmetrical geometry with the two reflectarrays placed in parallel planes. A 2D bifocal design performed in the offset plane is rotated around the antenna symmetry axis, and then, both centered and offset configurations are possible by selecting specific parts of the revolution surfaces. In the case of offset configurations, both reflectarrays can be tilted a certain angle to obtain smoother phase distributions. For this purpose, a novel phase adjustment routine has been implemented to compensate the tilting and maintain the bifocal characteristic of the original design.

A preliminary study on the bifocal technique for the design of multi-beam satellite antennas in Ka-band has been carried out, considering two main cases: reduction of the beam spacing by a factor of 2 (in order to provide adjacent beams with 0.56° angular spacing), and improvement of the multi-beam performance with respect to the equivalent monofocal antenna (in this case, without compressing the beams). The results show that the bifocal technique allows to provide the required 0.56° spacing by using non-overlapping feeds, but at the cost of a lower radiation efficiency of the bifocal antenna (the main reflectarray should be significantly oversized). On the other hand, the bifocal technique can be applied to provide the same beam spacing than in the single-focus case, obtaining a better performance for the beams at the edges of the coverage with satisfactory results for both the gain and the radiation efficiency of the bifocal antenna.

The bifocal technique has been also applied to the design of centered-fed bifocal dual transmitarray configurations. The design with transmitarrays brings some interesting advantages, such as lower sensitivity to surface deformations, the use of centered geometries with a focal ring and the absence of blockage. The design process is also simplified with respect to dual reflectarrays. The results of the simulations show that it is possible to achieve a high degree of beam spacing compression (by a factor of 2), but at the expense of a reduced radiation efficiency, as in the case of the bifocal dual reflectarray antenna.

A general 3D bifocal design technique for dual reflectarray antennas has been developed to overcome the drawbacks of the previous bifocal method, which imposes some geometrical restrictions in the initial positions of the foci and the reflectarrays and cannot be used for the design of offset configurations with a large tilting angle between the two reflectarrays. The new 3D bifocal technique can be used to directly compute the required phase distributions on each reflectarray, without imposing any restrictions in the antenna geometry. The technique is based on an iterative 3D ray-tracing routine that provides a grid of points on the surface of each reflectarray and the values of the partial phase derivatives associated to those points. The partial phase derivatives are interpolated, and then, integrated to obtain the bifocal phases required on each reflectarray. The technique has been validated for an axially-symmetrical configuration by comparison with the previous bifocal method (rotation of a 2D design around the

symmetry axis). The results of the comparison show negligible differences between the two approaches.

A dual reflectarray antenna in an offset compact-range configuration has been designed using the 3D bifocal method to generate multiple beams in transmission in Ka-band (19.7 GHz). The design has been performed considering three different degrees of beam spacing compression: no compression (1.24° of beam spacing), low compression (from 1.24° to 1.12°) and high compression (from 1.24° to 0.56°). The bifocal antenna with no beam compression improves the performance of the extreme beams with respect to the equivalent monofocal antenna. The bifocal antenna with low beam compression allows to obtain closer beams with non-overlapping feeds, at the same time as improving the performance of the extreme beams. Finally, the bifocal antenna with high beam compression provides the required beam spacing (0.56°), but at the cost of reducing the antenna radiation efficiency.

A bifocal dual reflectarray antenna demonstrator with a main reflectarray of 57 x 42 cm has been designed, manufactured and tested, in order to validate the bifocal technique. The prototype has been designed to operate in the 19.2-20.2 GHz band, providing 10 beams with 1.9° of spacing alternating in dual-LP, but the technique can be used to generate adjacent beams in dual-CP by using a sequential rotation technique. The measured patterns show both a good agreement with the simulations and a steady response within the prescribed frequency band. The results of the measurements show the capability of the bifocal technique to reduce the beam spacing (in this case, by a factor of 1.2) and provide a better performance for the extreme beams than the equivalent monofocal antenna.

The smaller beam spacing provided by the bifocal technique will make it possible a reduction in the required antenna size with respect to conventional reflectors for the same beam spacing. Moreover, the fabrication of a bifocal dual reflectarray antenna involves the same conventional photo-etching processes used for printed reflectarrays, as the only difference with the equivalent monofocal antenna are the dimensions of the printed elements. This will allow to improve the performance of current multi-beam satellite antennas (typically designed for a single focus) without increasing the costs of the antenna system. Note that in the case of bifocal dual reflectors, both reflectors are shaped as a consequence of the bifocal design procedure, thus requiring expensive custom moulds that should be manufactured specifically for each mission. The proposed

method can be used for the design of Tx/Rx antennas if appropriate reflectarray cells that enable independent phasing at both frequency bands are employed.

Finally, a bifocal dual reflectarray antenna with an elliptical main reflectarray (3.55 x 1.81 m) has been designed to provide multi-spot coverage at Tx in Ka-band (20 GHz). The bifocal technique is used to produce adjacent beams in the offset plane, reducing the beam spacing by a factor of 2. The interleaved beams in the orthogonal plane for a final 0.56° separation are generated in the orthogonal polarization. The combination of the bifocal technique in one plane and polarization discrimination in the orthogonal plane allows to provide the multi-spot coverage with elliptical beams. The proposed concept can be used to design a Tx/Rx antenna if multi-frequency reflectarray cells, as those presented in the first part of the thesis, are used. This solution represents an improvement with respect to the highly-oversized shaped reflector to provide the multi-spot coverage, as it requires a smaller main aperture.

8.2 Original contributions

- 1) A dual-polarization and dual-frequency reflectarray cell has been proposed and validated for simultaneous operation at two separate frequency bands, allowing independent phasing in each linear polarization at both bands. The reflectarray element consists of two orthogonally-arranged sets of parallel dipoles, distributed on a two-layer configuration. The dipoles on the lower layer will provide the required phases at the lower frequency, while the dipoles on the upper layer will do the same with the phases at the higher frequency. This operating principle is made possible due to the difference in the lengths of the dipoles between the two layers. The proposed element provides smooth phase curves with respect to the dipoles' lengths, covers more than 360° of phase range at both design frequencies, and presents a robust behavior against variations in the angle of incidence.
- 2) A step-by-step design procedure has been implemented and validated for dual-band reflectarray antennas that use the previous cell. The proposed two-layer configuration allows to perform separate design processes for each reflectarray layer: first, the lengths of the lower dipoles are adjusted to match the required

phases at the lower frequency, and then, the lengths of the upper dipoles are adjusted to introduce the required phase-shift at the higher frequency. This leads to a simpler and computationally faster design process. Possible phase errors can be corrected by means of an additional optimization process to be run after the design.

- 3) A 25-cm dual polarized reflectarray demonstrator that operates in the transmit frequencies in Ku (11-13 GHz) and Ka (19-20 GHz) bands has been designed, manufactured and tested with satisfactory results. The demonstrator is able to produce a focused beam in dual linear or dual circular polarization at two separate frequencies, using the aforementioned reflectarray cell. The proposed reflectarray permits an independent optimization of the radiation patterns and position of feed-chains for Ku and Ka bands. The design concept can be used for a transmit satellite antenna to fulfill independent requirements at each band (a contoured beam in Ku-band and multiple beams in Ka-band). The reuse of the same aperture would result in significant savings in the costs, weight and volume of the antenna farm in current satellite systems in Ku and Ka bands.
- 4) A bifocal design technique has been proposed for dual reflectarray antennas in offset configurations. The method starts by considering an axially-symmetrical geometry with the two reflectarrays placed in parallel planes. A 2D ray-tracing algorithm is applied to obtain the required bifocal phase curves of both reflectarrays in the offset plane. These curves can be rotated around the symmetry axis to obtain a 3D solution for the antenna design, and then, both centered and offset configurations are possible by selecting specific parts of the revolution surfaces. In the case of offset configurations, both reflectarrays can be tilted a certain angle to obtain smoother phase distributions, using a novel phase adjustment routine which has been implemented to compensate the tilting and maintain the bifocal characteristic of the original design.
- 5) The bifocal technique has been applied to the design of centered-fed dual transmitarray configurations for the first time. The use of transmitarrays allows for the design of centered geometries with a focal ring, leading to a simpler design process than in the case of dual reflectarrays. These advantages are achieved at the cost of a larger antenna volume. For this reason, different dual

transmitarray configurations have been studied to try to reduce the antenna volume, such as placing the feeds close to the first transmitarray (to integrate both elements on the same system), or reducing the distance between the two transmitarrays (so as to hold them with the same supporting structure).

- 6) To show the capability of the bifocal technique to provide a high degree of beam spacing reduction (by a factor of 2) with respect to the equivalent single-focus antenna for the same feed spacing. The bifocal antenna is able to provide the required spacing of 0.56° between adjacent beams using contiguous and non-overlapping feeds, but at the cost of a reduction in the radiation efficiency. These results have been shown for both dual reflectarray and dual transmitarray configurations.
- 7) To show the capability of the bifocal technique to improve performance of the extreme beams with respect to the equivalent monofocal antenna, when the bifocal technique is applied to provide the same beam spacing than in the monofocal design. In this case, the bifocal antenna presents satisfactory results for the gain and radiation efficiency, at the same time as providing a better shaped main lobe and lower side-lobe levels for most of the beams, which reduces the interference between adjacent beams in the same colour.
- 8) A general 3D bifocal design technique for offset dual reflectarray antennas has been implemented and validated for the first time. The technique allows to obtain the required phase distributions on each reflectarray without imposing any restrictions in the antenna geometry. The technique is based on an iterative 3D ray-tracing routine that provides a grid of points on the surface of each reflectarray and the values of the partial phase derivatives associated to those points. The phase derivatives are interpolated, and then, properly integrated to obtain the bifocal phase functions required on each reflectarray. The initial condition for the phase derivatives along the sub-reflectarray cross section can be determined, in the more general case, from the equivalent monofocal design.
- 9) A bifocal dual reflectarray antenna demonstrator with an offset compact-range configuration has been designed, manufactured and tested for the first time. The antenna has been designed to operate in dual polarization in the 19.2-20.2 GHz

band (Tx in Ka-band). The bifocal demonstrator provides 10 beams with 1.9° of separation alternating in dual-LP (3.8° of spacing between adjacent beams in the same polarization), but the technique can be used to generate adjacent beams in dual-CP by using a sequential rotation technique. The measured patterns show a good agreement with the simulations and a steady response within the prescribed frequency band.

- 10) To demonstrate the capability of the bifocal technique to provide a certain degree of beam spacing reduction, while improving at the same time the performance of the extreme beams. The bifocal antenna demonstrator is able to reduce beam spacing by a factor of 1.2 (from 4.6° to 3.8°) with respect to the equivalent monofocal antenna, with a similar radiation efficiency and without using overlapping feeds. The smaller beam spacing will make it possible a reduction in the required antenna size with respect to conventional reflectors for the same beam spacing. Also, the demonstrator provides better results for the gain (+0.3 dB), SLL (-2 dB) and 3-dB beamwidth (2.6° , similar to the other beams) of the extreme beams. This will allow to improve the performance of multi-beam satellite antennas (typically designed for a single focus) without increasing the costs of the antenna system, since the fabrication of the bifocal dual reflectarray involves the same conventional processes used for printed reflectarrays.
- 11) A bifocal dual reflectarray antenna with an elliptical main reflectarray has been proposed to provide multi-spot coverage for transmission in Ka-band. The bifocal technique is used to produce adjacent beams in the offset plane, reducing the beam spacing by a factor of 2. The main reflectarray has been oversized in the same plane to ensure low spillover. The interleaved beams in the orthogonal plane for providing full multi-spot coverage are generated in the orthogonal polarization. The proposed concept can be used to design a multi-beam satellite antenna to operate at Tx and Rx bands, if appropriate reflectarray cells enable independent phasing at both frequency bands are used. This solution represents an improvement with respect to the highly-oversized shaped reflector, as it requires an elliptical aperture of 3.5 x 1.8 m instead of a circular aperture with around 4.5 diameter.

8.3 Future research lines

The design techniques and the results presented in this thesis for multi-frequency and multi-beam reflectarrays and their capabilities to reduce the number of antennas in geostationary satellites for multi-spot coverage in Ka-band, have open new lines of research, which can be summarized as follows:

- Design of reflectarray antennas to provide adjacent beams in dual-circular polarization. The reflectarray element proposed in Chapter 2 allows independent control of each linear polarization, but most of the current multi-beam satellite antennas in Ka-band operate in dual-CP. The discrimination in dual-CP can be achieved by implementing a sequential rotation of the reflectarray elements, or by the design of a novel reflectarray cell to provide independent phasing in each CP. Both solutions should be further investigated to generate adjacent beams in dual-CP with a single feed. Furthermore, the polarization can be changed in a reflectarray from dual-LP into dual-CP, simply by adding 90° phase-shift to one of the components. This concept may be used to discriminate in dual-LP in the sub-reflectarray and to convert dual-LP into dual-CP on the main reflectarray.
- Design of multi-beam satellite antennas for operation at transmit and receive frequencies in Ka-band. The capabilities of reflectarrays for multi-frequency and multi-beam operation have been addressed in this thesis separately. The use of reflectarray cells that provide independent phasing in each polarization and frequency, combined with the application of the proposed bifocal design technique, allows for the design of a multi-beam satellite antenna to provide multi-spot coverage in transmission (19.7 GHz) and reception (29.5 GHz) in Ka-band. In this case, independent bifocal design processes can be performed for each frequency band, in order to obtain the required phase distributions on each reflectarray
- Extension of the bifocal design technique to dual reflectarray configurations with a parabolic main reflectarray. One main drawback of the proposed dual reflectarray configuration is the high number of 360° cycles that has to be implemented in the main flat reflectarray. The parabolic shape of the main reflectarray will reduce the number of 360° cycles that appear in the phase-shift distribution as a consequence of the large size (around 2 m in diameter). The

bifocal phases required on the paraboloid will be obtained by a 3D ray-tracing procedure similar to the one used with flat reflectarrays. The parabolic shape of the main reflectarray will focus the beam, while the phase-shift introduced by the printed elements will contribute to shape and point the beam in the appropriate direction. In this case, the discrimination in polarization can be in dual-LP on the sub-reflectarray and converted to dual-CP on the parabolic main reflectarray, in order to generate two adjacent beams in orthogonal CP with each feed operating in dual-LP. Note that the feed chain will be simplified, since there is no need of a polarizer.

- Implementation of phase synthesis techniques on a parabolic surface for improving the beam shaping. In some cases, the beam shaping obtained when applying the bifocal technique is not compliant with the requirements, as in the case of the elliptical oversized reflectarray that produces elliptical beams. Another example is when a dual reflectarray antenna is designed to operate at 20 GHz and 30 GHz. In this case, the beams will be narrower at 30 GHz because the electrical aperture is larger. In both cases, an additional phase correction is needed to be compliant with the required beamwidth. A phase-only synthesis technique, as the intersection approach, can be used to optimize the phase distribution on the main reflectarray. The problem is relatively easy in a flat reflectarray, since the starting phase distribution provides a beam shaping close to fulfill the requirements. However, the problem is more difficult when the reflectarray surface is not planar. The extension of the synthesis techniques to curved surfaces will be investigated.
- Investigation of reflectarray antenna configurations capable of generating independent beams for Ku and Ka satellite missions. Reflectarray cells have been demonstrated to operate at two frequencies (12 and 20 GHz). These elements can be used to design a reflectarray antenna to fulfill independent requirements at each frequency and/or polarization. For example, in the case of a satellite transmit antenna, a contoured beam can be generated in Ku-band and at the same time, multiple spot beams can be obtained in Ka-band, considering different feed chains for each mission. The implementation of different missions on the same reflectarray antenna would result in significant savings in the costs,

weight and volume of the antenna farm, especially in the case of telecommunication satellites that operate in Ku and Ka bands. A further investigation can be conducted to design this type of multi mission antenna. Furthermore, reflectarray elements providing independent operation at three frequency bands (12-14, 20 and 30 GHz) will be investigated. These type of multi-frequency reflectarrays may be used to operate in Ku and Ka bands, both in transmission and reception.

- Development of a multi-focal technique for the design of reflectarray and transmitarray antennas. The proposed bifocal technique can be extended to allow the design with a larger number of focal points, following a similar iterative procedure based on 3D ray-tracing. The inclusion of more focal points would provide an improved performance in the case of satellite antennas that are required to produce a high number of spot beams in a SFPB basis. On the other hand, it may require more degrees of freedom for the design, which may lead to more complex antenna configurations (e. g., an additional sub-reflectarray).
- Design of dual transmitarray antennas. Although the application of the bifocal technique to the design of dual transmitarray antennas with a high beam compression ratio presents severe limitations, as shown in Chapter 4, transmitarrays can be used in other applications that do not require beam compression. It would be interesting to study more in detail the capabilities of transmitarray antennas, considering the implementation of suitable transmitarray cells. As a first step, the appropriate transmitarray cells have to be designed, ensuring enough phase variation range in transmission and very low reflection losses. Transmitarray cells capable to introduce independent phasing in each polarization will be investigated. The design of these type of transmitarray cells is very challenging. Once the transmitarray cells are defined, they can be used for the design of single or dual transmitarray configurations. Since the volume of the two transmitarray panels is a severe issue that penalizes the usefulness in real applications, it would be interesting to study the concept of bifocal antennas by synthesizing the phasing on both surfaces of a single panel, in a similar manner as the original application of the bifocal concept to dielectric lenses. The dual-transmitarray configuration can be combined with other capabilities of the

transmitarray cells, such as discrimination in polarization or reconfigurability in a small sized first transmitarray, in order to provide an improved performance with respect to single transmitarray antennas.

8.4 List of publications related to this thesis

The results of the work developed in this thesis have given rise to the following publications in journals, national and international conferences:

8.4.1 Journal papers

- E. Martínez-de-Rioja, J. A. Encinar, M. Barba, R. Florencio, R. R. Boix and V. Losada, “Dual Polarized Reflectarray Transmit Antenna for Operation in Ku- and Ka-Bands with Independent Feeds”, *IEEE Transactions on Antennas and Propagation*, vol. 65, no. 6, pp. 3241-3246, June 2017.
- E. Martínez-de-Rioja, J. A. Encinar, R. Florencio and C. Tienda, “3D Bifocal Design Method for Dual Reflectarray Configurations with Application to Multi-Beam Satellite Antennas in Ka-Band”, submitted to *IEEE Transactions on Antennas and Propagation*.
- D. Martínez-de-Rioja, E. Martínez-de-Rioja, J. A. Encinar, R. Florencio and G. Toso, “Reflectarray to Generate Four Adjacent Beams per Feed for Multi-Spot Satellite Antennas”, submitted to *IEEE Transactions on Antennas and Propagation*.
- Two additional journal papers are being prepared based on the results obtained through the work reported in this thesis.

8.4.2 International conferences

- E. Martínez-de-Rioja, J. A. Encinar, R. Florencio and R. R. Boix, “Dual Polarized Reflectarray Antenna to Generate Independent Beams in Ku and Ka Bands”, in *Proc. 10th European Conference on Antennas and Propagation (EuCAP)*, Davos, Switzerland, April 2016, pp. 1-5.

- E. Martinez-de-Rioja, J. A. Encinar, R. Florencio and R. R. Boix, "Reflectarray in K and Ka Bands with Independent Beams in Each Polarization", in *Proc. IEEE International Symposium on Antennas and Propagation*, Fajardo, Puerto Rico, USA, July 2016, pp. 1199-1200.
- D. Martinez-de-Rioja, E. Martinez-de-Rioja and J. A. Encinar, "Multibeam Reflectarray for Transmit Satellite Antennas in Ka Band using Beam-Squint," in *Proc. IEEE International Symposium on Antennas and Propagation*, Fajardo, Puerto Rico, USA, July 2016, pp. 1421-1422.
- E. Martinez-de-Rioja, J. A. Encinar, R. Florencio and R. R. Boix, "Low-Cost Transmit and Receive Reflectarray Antenna for Satellite Communications in Ka-Band", in *4th Advanced Electromagnetics Symposium (AES)*, Málaga, Spain, July 2016.
- E. Martinez-de-Rioja, J. A. Encinar, A. Pino, B. Gonzalez-Valdes, C. Tienda, S. V. Hum and G. Toso, "Application of Bifocal Concept to Dual Reflectarray Configurations for Multi-Beam Satellite Antennas in Ka-band", in *Proc. 11th European Conference on Antennas and Propagation (EuCAP)*, Paris, France, March 2017, pp. 2427-2430.
- E. Martinez-de-Rioja, J. A. Encinar, C. Geaney, S. V. Hum and A. Pino, "Study of Bifocal Dual Reflectarray Configurations for Multi-Beam Antennas in Ka-band", in *Proc. IEEE International Symposium on Antennas and Propagation*, San Diego, California, USA, July 2017, pp. 1183-1184.
- C. Geaney, J. Sun, S. V. Hum, E. Martinez-de-Rioja and J. A. Encinar, "Synthesis of a Multi-Beam Dual Reflectarray Antenna Using Genetic Algorithms", in *Proc. IEEE International Symposium on Antennas and Propagation*, San Diego, California, USA, July 2017, pp. 1179-1180.
- E. Martinez-de-Rioja, J. A. Encinar, A. Pino, B. Gonzalez-Valdes, S. V. Hum, C. Tienda and G. Toso, "Bifocal Technique Applied to Dual Transmitarray Antennas", accepted in *12th European Conference on Antennas and Propagation (EuCAP)*, London, United Kingdom, April 2018.

- A. Pino, Y. Rodriguez-Vaqueiro, B. Gonzalez-Valdes, O. Rubiños, E. Martinez-de-Rioja, J. A. Encinar and G. Toso, “Design of a Bifocal Dual Reflectarray System with Parabolic Main Surface for a Multifed Space Antenna”, submitted to *IEEE International Symposium on Antennas and Propagation*, Boston, Massachusetts, USA, July 2018.

8.4.3 National conferences

- E. Martinez-de-Rioja, J. A. Encinar, R. Florencio and R. R. Boix, “Diseño de un Reflectarray de Polarización Dual para Generar Haces Independientes en las Bandas Ku y Ka”, *XXX Symposium Nacional de la Unión Científica Internacional de Radio (URSI)*, Pamplona, Spain, September 2015.
- E. Martinez-de-Rioja, J. A. Encinar, R. Florencio and R. R. Boix, “Antena Reflectarray en las Bandas K y Ka con Haces Independientes en Cada Polarización”, *XXXI Symposium Nacional de la Unión Científica Internacional de Radio (URSI)*, Madrid, Spain, September 2016.
- D. Martinez-de-Rioja, E. Martinez-de-Rioja and J. A. Encinar, “Estudio de Reflectarray para Aplicaciones de Satélite en Banda Ka Utilizando Discriminación en Frecuencia”, *XXXI Symposium Nacional de la Unión Científica Internacional de Radio (URSI)*, Madrid, Spain, September 2016.
- E. Martinez-de-Rioja, D. Martinez-de-Rioja and J. A. Encinar, “Antenas Reflectoras Planas en Tecnología Impresa para Radars Meteorológicos”, *XIII Congreso Nacional del Medio Ambiente (CONAMA)*, Madrid, Spain, November 2016.
- E. Martinez-de-Rioja, J. A. Encinar, A. Pino and B. González-Valdés, “New Bifocal Design Method for Dual Reflectarray Configurations with Application to Multiple Beam Antennas in Ka-Band”, *XXXII Symposium Nacional de la Unión Científica Internacional de Radio (URSI)*, Cartagena, Spain, September 2017.

8.5 Framework and research projects related to this thesis

This thesis has been developed at the Departamento de Señales, Sistemas y Radiocomunicaciones of Universidad Politécnica de Madrid. Also, part of the work presented in Chapter 3, concerning the design of multi-beam satellite antennas in Ka-band, has been developed within a research stay at the Edward S. Rogers Sr. Department of Electrical and Computer Engineering of University of Toronto, from August 11th, 2016 to November 11th, 2016, under the supervision of Prof. Sean Victor Hum.

The work presented in this thesis has been supported by several national and international research projects, which are listed below:

- “Reflectarray Antennas with Improved Performances and Design Techniques”, Supporting agency: European Space Agency (ESA-ESTEC), Years: Sept. 2012 – Dec. 2015, Participants: Universidad Politécnica de Madrid (UPM), Universidad de Oviedo, Universidad de Sevilla. Head: José A. Encinar (UPM).
- “New Concepts in Reflectarrays and Transmitarrays for Innovative Antennas and their Experimental Validation”, Supporting agency: Spanish Commission for Science and Technology (CICYT), Years: 2013 to 2016. Head: José A. Encinar (UPM).
- “Multiple Beam Antennas based on Reflectarrays and Transmitarrays”, Supporting agency: European Space Agency (ESA-ESTEC, Contract No. 4000117113/16/NL/AF), Years: Sept. 2016 - Dec. 2018, Participants: Universidad Politécnica de Madrid (UPM), Universidad de Vigo. Head: José A. Encinar (UPM).
- “Multiple-Antenna Advanced Subsystems for Ground and Satellite Wideband Communications”, Supporting agency: Spanish Commission for Science and Technology (CICYT), Years: 2017 to 2020. Head: José A Encinar (UPM).

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