Absorptive frequency-selective structures and their applications in antenna systems of low radar cross section

Huang, Hao

2020

Huang, H. (2020). Absorptive frequency-selective structures and their applications in antenna systems of low radar cross section. Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/145121>

<https://doi.org/10.32657/10356/145121>

This work is licensed under a Creative Commons Attribution-NonCommercial 4.0 International License (CC BY-NC 4.0).

Downloaded on 26 Apr 2023 20:52:14 SGT

Absorptive Frequency-Selective Structures and Their Applications in Antenna Systems of Low Radar Cross Section

Huang Hao

School of Electrical & Electronic Engineering

A thesis submitted to the Nanyang Technological University
in partial fulfillment of the requirements for the degree of
Doctor of Philosophy

2020

Statement of Originality

I hereby certify that the work embodied in this thesis is the result of original research, is free of plagiarised materials, and has not been submitted for a higher degree to any other University or Institution.

15/07/2020

.....

Date

Huang Hao

.....

Huang Hao

Supervisor Declaration Statement

I have reviewed the content and presentation style of this thesis and declare it is free of plagiarism and of sufficient grammatical clarity to be examined. To the best of my knowledge, the research and writing are those of the candidate except as acknowledged in the Author Attribution Statement. I confirm that the investigations were conducted in accord with the ethics policies and integrity standards of Nanyang Technological University and that the research data are presented honestly and without prejudice.

15/07/2020

.....

Date

.....

Prof. Zhongxiang Shen

Authorship Attribution Statement

This thesis contains materials from four papers published in the following peer-reviewed journals and four conferences papers in which I am listed as an author.

Chapter 3 contains material published in

* H. Huang, Z. Shen and A. A. Omar, "3-D absorptive frequency selective reflector for antenna radar cross section reduction," *IEEE Trans. Antennas Propag.*, vol. 65, no. 11, pp. 5908-5917, Nov. 2017.

* H. Huang, A. A. Omar and Z. Shen, "Absorptive frequency-selective reflectors," in *proc. IEEE Int. Conference Computational Electromag. (ICCEM)*, Shanghai, China, Mar. 2019.

The contributions of the co-authors are as follows:

- Prof. Zhongxiang Shen suggested the research topic, provided suggestions, and edited the manuscript drafts.
- I proposed design concept and prepared the manuscript drafts. The structure design, simulations, fabrications, and data analysis were conducted by me.
- Dr. Ahmed Abdelmottaleb Omar assisted in revising the draft.

Chapter 4 contains the material published in

* H. Huang and Z. Shen, "Low-RCS reflectarray with phase controllable absorptive frequency-selective reflector," *IEEE Trans. Antennas Propag.*, vol. 67, no. 1, pp. 190-198, Jan. 2019.

* H. Huang and Z. Shen, "Realization of Low-RCS parabolic reflector antenna using curved 3-D frequency-selective structure," in *proc. IEEE Int. Symp. Antennas Propag.*, Boston, MA, USA, Jul. 2018, pp. 1723-1724.

The contributions of the co-authors are as follows:

- Prof. Zhongxiang Shen provided comments and edited the manuscript drafts.
- I conceived the design concept, performed the structure designs, conducted simulations, and obtained experiment results. The manuscript drafts were prepared by me.

Chapter 5 contains the material published in

* H. Huang, A. A. Omar and Z. Shen, "Low-RCS and beam-steerable dipole array using absorptive frequency-selective reflection structures," *IEEE Trans. Antennas Propag.*, vol. 68, no. 3, pp. 2457-2462, Mar. 2020.

The contributions of the co-authors are as follows:

- Prof Zhongxiang Shen provided the initial idea and revised the manuscript.
- I proposed the design, performed simulations, prepared fabrications and assembly, collected radiation patterns and monostatic RCS, and carried data post-processing and analysis. The manuscript draft was written by me.
- Dr. Ahmed Abdelmottaleb Omar made suggestions on fabrications.

Chapter 6 contains the material published in

* H. Huang and Z. Shen, "Absorptive frequency-selective transmission structure with square-loop hybrid resonator," *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 3212-3215, 2017.

* H. Huang and Z. Shen, "An absorptive frequency-selective reflector based on miniaturized square-loop resonator," *Proc. of 12th European Conference on Antennas and Propagation (EuCAP)*, London, UK, Apr. 2018.

* H. Huang and Z. Shen, "Ultra-Broadband 3-D absorptive frequency-selective transmission structure using commercial absorber," *in proc. IEEE Int. Symp. Antennas Propag.*, Montréal, QC, CA, Jul. 2020.

The contributions of the co-authors are as follows:

- Prof Zhongxiang Shen provided comments and polished the manuscript.
- I proposed the design, performed full-wave simulations and circuit analysis, prepared fabrications and conducted experiments. The manuscript draft was prepared by me.

15/07/2020

.....

Date

.....

Huang Hao

Acknowledgements

Foremost, I wish to express my deepest and enormous gratitude to my supervisor Prof. Shen Zhongxiang, for providing me with this precious opportunity to pursue my doctorate degree at Nanyang Technological University. He not only spares his time to discuss my encountered research problems but also provides constructive feedbacks. Without his continuous instructions and guidance, I would not quickly start my research on low RCS antennas and explore the frontier topics of absorptive frequency-selective structures. This dissertation definitely would not be possible without his invaluable support. His deep insight, radiant optimism, rigorous attitude and unlimited patience have affected me deeply and set a role model for me in the future.

I would like to take the opportunity to express my appreciation to all my collages and my friends in Singapore. It is my honor to work in a talented team and learn from these geniuses. Special thanks go to the former group members of Dr. Shang Yuping, Dr. Ahmed Abdelmottaleb Omar, and Dr. Zuo Le for their assistance and valuable discussions. Many thanks to the technicians in Communication Research Lab and Infocomm Centre for Excellence, Thung Lina, Teng Siew Juan Catherine, and Ng Tengkwee, for their professional support in research.

In the end, my deepest love is devoted to all of my family members. Their unconditional support and endless love refuels me with power and confidence for overcoming the difficulties during my PhD study. Special thanks go to my dearest wife Dr. Luo Hui for her love and encouragement.

“Nobody ever figures out what life is all about, and it doesn’t matter. Explore the world. Nearly everything is really interesting if you go into it deeply enough.”

—Richard P. Feynman

To my dear family

Abstract

Microwave absorbers are widely used to reduce the overall reflection levels of a target or an object. For most of absorbers, they produce reflection bands out of their absorption bands due to the existence of impenetrable ground planes. On one hand, transmission band within the absorption band does not naturally exist in most absorbers. On the other hand, the reflection band of absorbers is fixed once they have been manufactured. These two issues have limited the application of absorbers in antenna systems, especially when the radiation efficiency is the major concern. To overcome these problems, absorptive frequency-selective reflection (AFSR) structures and absorptive frequency-selective transmission (AFST) structures are proposed in this thesis. Frequency responses of absorption-reflection-absorption and absorption-transmission-absorption are introduced, which provide more degrees of freedom in manipulating electromagnetic waves in the frequency domain.

The original design concept of AFSR structure is implemented by integrating multimode resonators inside a periodic unit cell. Its operating principle is interpreted with the aid of propagating modes and equivalent circuit model. Following this, a periodic array of AFSR structure is co-designed with dipole and monopole antennas, in order to achieve RCS reduction at both lower and upper frequency bands with a maintained radiation performance for monopole antenna and an enhanced gain for dipole antenna. Full-wave simulations and experiments are performed on characterizing the radiation and scattering performance of AFSR-backed antennas, which have verified the concept of design and effectiveness of our demonstrations.

The study on AFSR structure is further extended by exploiting its phase response. A phase-controllable AFSR structure is then proposed with maintained amplitude response and variable reflection phase. More specifically, it provides an amplitude response of absorption-reflection-absorption. Meanwhile, the reflected wave fronts can be shaped by spatially varying the distribution of reflection phase, creating new functionalities such as beam focusing, defocusing and steering. This design

concept has been employed in the construction of a parabolic reflector antenna with high directivity and low scattering. Following this, a fully planar version of phase controllable AFSR structure is proposed, which is demonstrated by designing low RCS reflectarray antenna with center feed and offset feed, respectively. Simulations and experiments results are provided, which have verified our design concept.

The RCS reduction of phased array antenna is of high priority because it is one of the major scattering sources in communication platforms. In this thesis, a new approach of designing a low-RCS phased array element is proposed by co-locating a passive AFSR structure with an active radiator inside the same unit cell. The absorption bandwidth for the AFSR structure is widely extended by stacking absorbers into a multi-layer structure. The radiation efficiency and beam steering capability of the proposed array are well maintained. Moreover, the RCS of the proposed AFSR-backed dipole antenna array has a significant reduction compared with the reference PEC-backed array. The proposed AFSR-backed phased array antenna has been demonstrated with full-wave simulations and measurements.

The design concept of AFST structure has been demonstrated with two different approaches, which are based on hybrid resonators and integrated frequency-selective structures, respectively. The former design reduces the consumption of lumped resistors in a unit cell to half through an optimum layout. The latter one does not include any lumped elements and is realized by integrating FSSs with commercial absorbers. Simulation and measurement results are also presented with good agreement, which have shown a low insertion loss transmission band within a wide absorption band. The proposed AFST designs can potentially be applied as antenna superstrates or radomes for RCS reduction without affecting the radiation performance of the antenna.

In summary, this thesis introduces the absorptive frequency-selective structures and demonstrates their applications in antenna systems for RCS reduction. Two specific types of absorptive FSSs are introduced and discussed, which are termed as AFST structure and AFSR structure, according to their frequency responses. The associated applications of AFST structures and AFSR structures are also presented. In the end, this thesis is concluded by making suggestions for future research perspectives. With future research in functional electromagnetic surfaces or structures, more opportunities will be explained and more challenges will be attempted.

Contents

Acknowledgements	ix
Abstract	xiii
List of Figures	xvii
List of Tables	xix
Abbreviations	xxi
1 Introduction	1
1.1 Background and Objectives	1
1.2 Major Contributions of Thesis	4
1.3 Organization of the Thesis	5
2 Literature Review	7
2.1 Antennas of Low Radar Cross Section	7
2.1.1 Radar Cross Section Reduction Techniques	7
2.1.2 Antenna Radar Cross Section Reduction	13
2.2 Absorptive Frequency-Selective Structures	17
2.2.1 Absorptive Frequency-Selective Transmission Structures . .	17
2.2.2 Absorptive Frequency-Selective Reflection Structures	21
3 Low-RCS Monopole and Dipole Antennas	27
3.1 Introduction	27
3.2 Design Concept	27
3.3 Implementation Details	28
3.3.1 Configuration of Unit Cells	29
3.3.2 Propagation Modes	30
3.3.3 Equivalent Circuit Model	32
3.3.4 Dual-Polarized Structure Extension	36
3.3.5 Effect of the Ground Plane Size	38
3.4 Experiment and Results	40
3.4.1 Monopole Antenna	40

3.4.2	Dipole Antenna	44
3.4.3	Discussions on Gain Enhancement	49
3.5	Summary	51
4	Low-RCS Reflector Antennas	53
4.1	Introduction and Motivations	53
4.2	Design Concept	55
4.3	Low-RCS Parabolic Reflector Antenna	57
4.3.1	AFSR Unit Cell Design	57
4.3.2	Curved 3-D AFSR Reflector	59
4.3.3	Radiation and RCS Performance	63
4.4	Low-RCS Reflectarray with Center Feed	64
4.4.1	Planar and Phase-Controllable AFSR Structure	65
4.4.2	Reflection Phase Mapping	72
4.4.3	Experimental Verification	76
4.5	Low-RCS Reflectarray with Offset Feed	83
4.5.1	Reflection Phase Analysis	83
4.5.2	Experimental Verification	86
4.6	Summary	89
5	Low-RCS Phased-Array Antenna	91
5.1	Introduction	91
5.2	Implementation Details	93
5.2.1	AFSR Structure with Commercial Absorber	93
5.2.2	Co-located Array Element and AFSR Structure	95
5.3	Active VSWR Performance	98
5.4	Experiment and Results	100
5.4.1	Fabrication and Measurement Setup	100
5.4.2	Radiation Performance	102
5.4.3	RCS Performance	105
5.5	Summary	108
6	Low-RCS Antenna Radomes	109
6.1	Introduction	109
6.2	Absorptive Frequency-Selective Transmission Structures	110
6.3	AFST Structure Based on Hybrid Resonators	111
6.3.1	Unit Cell Configuration	111
6.3.2	Equivalent Circuit model	114
6.3.3	Normal and Oblique Incidence Performance	117
6.3.4	Experimental Verifications	118
6.4	FSS-Based AFST Structure	121
6.4.1	Absorption Performance	121
6.4.2	Bandpass and Bandstop FSS	123
6.4.3	Integrated AFST Structure	125

6.4.4	Measurements and Results	128
6.5	Summary	130
7	Conclusion and Future Work	133
7.1	Conclusion	133
7.2	Future Works	135
	List of Author's Awards and Publications	139
	Bibliography	141

List of Figures

2.1	Microwave absorbing structures	8
2.2	Low scattering surfaces based on artificial magnetic conductor and polarization conversion structure.	11
2.3	Antenna RCS reduction by using absorptive superstrates.	14
2.4	Antenna RCS reduction by coating lossless metasurfaces	16
2.5	Functionality of absorptive frequency-selective transmission structure	18
2.6	Absorptive frequency-selective transmission structures based on cascading lossless FSS and lossy layers.	19
2.7	Absorptive frequency-selective transmission structures based on 3-D FSS design.	20
2.8	Absorptive frequency-selective reflection structures.	22
2.9	Equivalent circuit model and frequency responses of AFSR structure for cascaded-type and series-type designs.	24
3.1	Frequency responses of AFSR structure	28
3.2	Illustration of integrated AFSR structure with antenna	28
3.3	3D-AFSR structure based on multi-resonators.	29
3.4	Unit cell of vertical stripline resonators with periodic boundary conditions.	31
3.5	Electric field distribution of substrate mode and air mode propagate inside the vertical double-sided stripline structures.	31
3.6	Characteristic impedance and propagation constant of original and approximate air modes	33
3.7	Equivalent circuit model of 3D-AFSR structure for single-polarized incident wave.	34
3.8	Reflection and absorption performance of 3D-AFSR with full-wave simulation and equivalent circuit analysis.	36
3.9	Unit cell and period array of dual-polarized 3D-AFSR structure . .	37
3.10	Reflection coefficients of 3D-AFSR for oblique incident waves	38
3.11	Monostatic RCS of PEC and 3D-AFSR with different sizes	39
3.12	Monopole antenna backed with 7×7 unit cells AFSR array.	41
3.13	Fabricated prototype of AFSR-backed monopole antenna.	42
3.14	Simulated and measured $ S_{11} $ of PEC-backed and AFSR-backed monopole antenna.	43

3.15	Simulated and measured radiation patterns of PEC-backed and AFSR-backed monopole antenna.	43
3.16	Simulated and measured mono-static RCS of PEC-backed and AFSR-backed monopole antenna.	44
3.17	Dipole antenna backed with 7×7 unit cells AFSR array.	45
3.18	Fabricated prototype of AFSR-backed dipole antenna.	45
3.19	Fabricated PEC-backed dipole antenna for reference.	46
3.20	Simulated and measured $ S_{11} $ of PEC-backed and AFSR-backed dipole antenna.	46
3.21	Simulated and measured radiation patterns of PEC-backed and AFSR-backed dipole antenna.	47
3.22	Simulated and measured gain over frequency of PEC-backed and AFSR-backed dipole antennas.	47
3.23	Simulated and measured mono-static RCS of PEC-backed and AFSR-backed dipole antenna.	48
3.24	Current distribution of PEC and AFSR backed dipoles	49
3.25	Current distribution of PEC and AFSR backed monopoles	50
3.26	Reflection phase distribution of PEC and AFSR surface.	50
4.1	Conceptual depiction of a low scattering reflective focusing surface based on phase controllable absorptive frequency-selective reflection structure.	56
4.2	Structure of the AFSR with a dual-polarized bandstop FSS and a commercial absorber.	58
4.3	Simulated reflection coefficients for AFSR structure with TE and TM polarized incident waves under normal and oblique incidence.	59
4.4	Surface geometry of continuous and discrete parabolic reflector	60
4.5	Geometry of two-dimensional parabolic reflector antenna based on curved 3-D bandstop frequency-selective structure.	61
4.6	Radiation patterns of parabolic reflector antenna with curved AFSR structure and PEC at 8.2 GHz.	63
4.7	Monostatic RCS of low RCS parabolic reflector antenna with curved AFSR and conventional reflector antenna.	63
4.8	Simulated bi-static scattering patterns of a low-RCS parabolic reflector antenna and a conventional reflector antenna.	64
4.9	Geometry of the unit cell with a single arc-shaped metallic strip resonator and its simulated reflection and transmission coefficients.	65
4.10	Unit cell structure of a second-order 3-D bandstop FSS and its equivalent circuit model.	66
4.11	Unit cell structure of AFSR by cascading 3-D bandstop FSS over a commercial absorber backed with PEC plane.	69
4.12	Simulated reflection magnitude and phase of low-RCS reflectarray unit cell under the normal incidence.	71
4.13	Reflection phase calculation for each elements of a reflectarray.	73

4.14	Calculated efficiency and reflection phase distribution for proposed reflectarrays.	74
4.15	The prototype of proposed low-RCS reflectarray.	77
4.16	Photographs of fabricated conventional reflectarray and low-RCS reflectarray.	78
4.17	Radiation patterns and gain versus frequency of conventional and low-RCS reflectarrays.	79
4.18	Simulated bi-static scattering patterns for TE polarized wave in xoz plane and TM polarized waves in $yo z$ plane.	81
4.19	Simulated and measured monostatic RCS for conventional and low-RCS center-feed reflectarrays.	82
4.20	Illustrations and coordinates of conventional and low-RCS offset-feed reflectarray antennas.	84
4.21	Reflection phase distribution of offset-feed reflectarray antenna. . .	85
4.22	Fabricated prototypes of the reference and proposed low-RCS reflectarray with an offset feed.	86
4.23	Simulated and measured radiation patterns of offset-feed conventional reflectarray and low-RCS reflectarray antennas.	87
4.24	Simulated and measured gain versus frequency for offset-feed conventional reflectarray and low-RCS reflectarray antennas	88
4.25	Simulated and measured monostatic RCS for conventional and low-RCS center-feed reflectarrays.	88
5.1	Frequency responses of AFSR structure under normal incidence. . .	93
5.2	Frequency responses of AFSR structure under oblique incidence. . .	94
5.3	Frequency responses of enhanced wide-band absorber and 3-D band-stop FSS.	94
5.4	Illustration of AFSR-backed dipole antenna array element and the reference PEC-backed dipole array element.	95
5.5	Simulated $ S_{11} $ and gain of AFSR-backed array element and the reference PEC-backed array element.	97
5.6	Simulated radiation patterns of AFSR-backed array element and the reference PEC-backed array element.	97
5.7	Electric field distribution of low-RCS dipole element in radiating and scattering cases.	98
5.8	Simulated active VSWR of AFSR-backed dipole array at 10 GHz with 45° beam steering in the E-plane and H-plane.	99
5.9	Photograph of the fabricated 8×8 AFSR-backed dipole antenna array with detailed view of assembled parts.	101
5.10	Photograph and measurement setup for monostatic RCS	102
5.11	Simulated and Measured $ S_{11} $ of PEC-backed and AFSR-backed dipole antenna array.	103
5.12	Measured radiation patterns of PEC-backed and AFSR-backed array in E-plane and H-plane at 10 GHz.	104

5.13	Simulated aperture efficiency and measured gain over frequency of AFSR-backed dipole array and the reference PEC-backed dipole array.	104
5.14	Simulated scanned beams in E-plane and H-plane at 10 GHz	105
5.15	Measured monostatic RCS for AFSR-backed dipole array and the reference PEC-backed dipole array.	106
5.16	Simulated bistatic RCS for AFSR-backed dipole array and the reference PEC-backed dipole array.	107
6.1	Functionality of an AFST structure.	110
6.2	Unit cell structure of AFST based on square loop hybrid resonators.	112
6.3	Parametric scan results of different resistor values.	113
6.4	Current distribution of square loop hybrid resonator.	113
6.5	Equivalent circuit model and full-wave simulated results of AFST structure.	115
6.6	Simulated frequency responses of hybrid resonator based AFST structure under the normal and oblique incidence.	117
6.7	Fabricated samples of AFST structure and measurement setup. . .	118
6.8	Simulated and measured frequency response and absorption rate for the fabricated AFST under the normal incidence.	119
6.9	Relative permittivity and permeability of FGM-125 absorber. . . .	122
6.10	Simulated reflection coefficient of single-layer and three-layer cascaded absorbers.	122
6.11	Simulated frequency response of absorber with an open slot. . . .	123
6.12	Simulated transmission and reflection coefficients of bandpass FSS.	124
6.13	Simulated and measured frequency response and absorption rate for the fabricated AFST under the normal incidence.	124
6.14	Simulated performance of AFST structure.	125
6.15	Simulated and measured frequency response and absorption rate for the fabricated AFST under the normal incidence.	126
6.16	Simulated electric field distribution of commercial absorber based AFST structure at different frequencies.	127
6.17	Fabricated sample of commercial absorber based AFST structure. .	128
6.18	Simulated and measured results of commercial absorber based AFST structure.	129

List of Tables

2.1	Comparison of typical microwave absorbers in literature	9
2.2	Comparison of the existing low scattering surface designs in the literature	12
3.1	Physical dimensions of 3D-AFSR unit cells.	30
3.2	Parameters for the equivalent circuit model of the 3D-AFSR structure.	35
3.3	Physical dimensions of the dual-polarized 3D-AFSR unit cells . . .	37
3.4	Summary of AFSR-backed monopole and dipole antennas.	49
4.1	Physical dimensions of AFSR structure for parabolic reflector . . .	58
4.2	Dimensions of substrate height for 1-D array of unit cells	62
4.3	Dimensions and circuit parameters of a second-order bandstop FSS.	67
4.4	Physical dimensions of AFSR structure for reflectarray.	70
4.5	Estimated loss budgets for the reference reflectarray and low-RCS reflectarray.	75
4.6	Comparisons between conventional reflectarray and proposed low-RCS reflectarray.	80
4.7	Comparisons between the conventional and our proposed low-RCS offset-feed reflectarrays.	89
5.1	Physical dimensions of reference and low-RCS dipole array element	96
6.1	Physical dimensions and parameters of AFST unit cell.	111
6.2	Equivalent circuit parameters of AFST unit cell.	116
6.3	Comparison of the existing 2-D AFST design in the literature. . . .	120
6.4	Physical dimensions of Commercial absorber based AFST unit cell .	126

Abbreviations

AFSR	Absorptive Frequency-Selective Reflection
AFST	Absorptive Frequency-Selective Transmission
AMC	Artificial Magnetic Conductor
CAA	Circuit Analog Absorber
EBG	Electronic Bandgap
FSR	Frequency Selective Resorber
FSS	Frequency-Selective Structure
HFSS	High-Frequency Structure Simulator
HIS	High Impedance Surface
PCS	Polarization Conversion Structure
PBC	Periodic Boundary Condition
PEC	Perfect Electric Conductor
PMC	Perfect Magnetic Conductor
PRS	Partially Reflecting Surface
RAM	Radar Absorbing Material
RCS	Radar Cross Section
SLL	Side Lobe Levels
SLHR	Square Loop Hybrid Resonator
VNA	Vector Network Analyzer

Chapter 1

Introduction

1.1 Background and Objectives

The design of a low-observable platform can significantly improve the survivability and reduce the detectability of future carriers, such as aircrafts, missiles, or vessels. As an indispensable part of communication systems, antennas inevitably contribute to a substantial electromagnetic scattering when transmitting or receiving signals. Therefore, antennas of low radar cross section (RCS) have become an important consideration for future stealthy platforms. Although the principles of RCS reduction have been known for many decades [1, 2], some techniques such as modifying shapes and applying radar absorbing materials (RAMs) have been applied in the RCS reduction of targets successfully. Therefore, the intuitive way of designing a low-RCS antenna is to combine these reduction techniques with the existing antenna designs. However, antennas are different from those passive structures, which do not need to transmit or receive signals. Apart from the RCS reduction performance, the radiation performance of antennas at in-band frequencies are also one of the major concerns. Therefore, it is a challenging task to apply these practical RCS reduction techniques to the design of low-RCS antennas.

Some previous research work was introduced on the analysis of antenna's scattering characteristics [3]. The predominant terms of antenna scattering are classified into antenna mode and structural mode. The first term is related to the radiation performance of the antenna and the second one is associated with the physical shape and material composition of the antenna. It is well known that shaping is one of the

important techniques in RCS reduction of a passive structure. However, in terms of antenna, any modified shape or size may affect the radiation performance of antenna. The key point of the research on low-RCS antennas is to find possible ways of reducing the antenna's structural mode while maintaining its antenna mode. In the frequency domain, the scattering of antenna is different at in-band and out-of-band frequencies. The out-of-band scattering of antennas can be reduced by using FSS or absorbers. The in-band scattering is more complicated, which is related to the impedance matching capabilities and the feeding networks. In most cases, the in-band scattering is much lower than the out-of-band ones, as long as these antennas are well matched at the operating frequencies. Therefore, the majority of previous studies are focused on the antenna RCS reduction at the out-of-band frequencies.

The existing solutions to antenna RCS reduction at the out-of-band frequencies are classified into two main types of absorptive and non-absorptive approaches. The idea of absorptive approach is to introduce losses in the antenna structures for scattering reduction. The losses can be implemented with lumped resistors, resistive sheets and absorbing materials. However, the absorption rate for most of lossy materials is a fixed value, once they have been manufactured. Therefore, the operating frequencies of some low-RCS antennas are designed not within the absorption bands of absorbers [4, 5]. Otherwise, the losses from the applied absorbers may result in a degradation of antenna's gain and efficiency. The current absorptive approaches can be improved if the absorption bands can be tailored to fit the operating frequencies of the antennas under consideration. For non-absorptive approaches, new types of metasurfaces such as chessboard surfaces, polarization conversion metasurfaces and coding metasurfaces are introduced in antenna designs to realize a diffusive scattering [6–8]. The limitation of the current non-absorptive approaches is that the aperture efficiency of these low-RCS antennas is very low, which is due to the fact that the applied metasurfaces cannot contribute to the radiation of the antenna at the in-band frequencies. In the meantime, there is a great demand for low-RCS and high gain antennas, because they play a vital role in long distance communication links with an improved signal to noise ratio and a maintained stealthy performance. However, the existing RCS reduction techniques are not capable of solving the considering problem of achieving a high gain and high aperture efficiency for the antenna while maintaining a low RCS. Therefore,

it is essential to develop new techniques to addressing the problem for both RCS reduction and radiation performance enhancement.

It has been discussed that previous designs of low RCS antennas suffer from a number of limitations. In this thesis, we aim to present a state-of-art study on low RCS antennas, providing possible solutions to the existing limitations and proposing new techniques to achieve RCS reduction for a variety of antennas. Two types of absorptive FSSs are proposed, which are termed as AFSR structure and AFST structure. The AFSR structure is characterized with a unique frequency response of absorption-reflection-absorption. The introduced reflection band of AFSR structure will play a role in the gain enhancement of low-RCS antennas. Meanwhile, the lower and upper absorption bands of the AFSR structure will reduce the scattering of antennas. The AFST structure is featured with absorption-transmission-absorption frequency response, which is able to provide a transmission band for the operation of antenna and reduce the antenna RCS at the out-of-band frequencies at the same time.

The objectives of this thesis are summarized as follows.

- To conduct a literature review on RCS reduction techniques and their current applications in low RCS antennas. Some recently reported progresses on AFSR and AFST structures will be reviewed with a detailed analysis and extensive discussions. The primary goal of this thesis is to utilize the unique frequency responses of AFSR and AFST structures as a key to solve the problem of antenna RCS reduction. Meanwhile, some of the criterions such as high gain, high aperture efficiency and high radiation efficiency will be maintained for our low-RCS antennas.
- To propose a solution to RCS reduction of dipole and monopole antennas with maintained or enhanced radiation performance. Explore the possibilities of new design methods for AFSR structures by utilizing the 3-D FSS design concept, which are implemented by vertical microstrip and stripline resonators. The proposed 3-D AFSR structures are intended to be utilized as a ground plane of dipole and monopole antennas for their RCS reduction.
- To achieve RCS reduction for high gain antennas such as reflector antennas. Typical reflector antenna system is characterized with a feeding antenna and a reflector surface. The latter one is designed many times larger than its

operating wavelength for beam collimation in the farfield, contributing to a significant scattering. AFSR structures with controlled reflection phase are intended to solve the problem of RCS reduction for parabolic reflector antennas and reflectarray antennas.

- Low RCS antenna array with steerable radiation beam is essential in radar systems for object detection and target tracking. Our work aims to design a low RCS phased-array element by integrated dipole antenna with AFSR structure. Therefore, the RCS of a high-performance phase array antenna can be minimized at its out-of-band frequencies. Meanwhile its radiation performance and beam steering capability will also be maintained.
- Low RCS radome is an alternative approach of achieving RCS reduction of antennas, while not affecting its radiation performance. We intend to implement a low-loss transmission band within a wide absorption band. The wide absorption bands are realized with lumped resistors and commercial absorbers. The low loss transmission bands can be implemented with hybrid resonators and integrated FSSs, respectively.

1.2 Major Contributions of Thesis

This thesis introduces the design concepts of AFSR structures and AFST structures. Their potential applications in RCS reduction of a variety types of antennas are also demonstrated. Our main contributions can be stated as follows.

- An absorptive frequency-selective reflection structure with an absorption-reflection-absorption frequency response is proposed to achieve RCS reduction of antennas at their out-of-band frequencies. The frequency response of our proposed AFSR structure is superior to conventional structures such as Salisbury screen in terms of absorption bandwidth, reflection band flatness and sharpness of absorption to reflection transition. Our AFSR structure can be readily deployed as the ground plane for a variety of antennas with reduced RCS and enhanced radiation performance.
- The methodology of controlling the phase response at the reflection band of an AFSR structure is proposed in this thesis. Both flat and curved versions

of AFSR structures with manipulated distributions of reflection phases are proposed. This design concept is further demonstrated by reducing the RCS of a parabolic reflector antenna and a reflectarray antenna without sacrificing their radiation efficiencies. The performance of RCS reduction is superior to the existing techniques in terms of reduction level and reduction bandwidth.

- An approach of implementing a low scattering dipole array is proposed. It is demonstrated by co-locating dipole element with AFSR structure inside the same unit cell to achieve enhanced forward radiation and reduced array scattering, simultaneously. The RCS reduction bandwidth is significantly extended by employing cascaded commercial absorbers, Meanwhile, the radiation efficiency of our proposed array is maintained and the beam steering capability is retained.
- Two AFST structures with a frequency response of absorption-transmission-absorption are proposed. Different design methods of AFST structures utilizing hybrid resonators and integrated FSSs are demonstrated. Compared with previous designs in the literature, the hybrid resonator based AFST structure reduces the usage of lumped resistors in a unit cell by 50% with maintained absorption and transmission characteristics. The integrated FSS-based AFST structure moves a step further by proposing a design without using any lumped elements.

1.3 Organization of the Thesis

Chapter 1 introduces the background of this thesis and provides a brief summary of the main contributions and outlines of our research work.

Chapter 2 presents a literature review about the absorptive frequency-selective reflection structures and absorptive frequency-selective transmission structures. The advantages and limitations of the current related research on these structures are discussed. Meanwhile, as the potential applications of AFST and AFSR structures, topics on RCS reduction and low RCS antennas are also introduced.

Chapter 3 proposes the design concept of AFSR structures, which is further implemented by using 3-D resonators. Furthermore, the applications of our 3-D AFSR

structure are demonstrated with RCS reduction of monopole and dipole antennas. It has been shown in both simulations and experiments that RCS reduction with an enhanced or maintained gain enhancement are obtained for the dipole and monopole antennas with our proposed 3-D AFSR structure.

Chapter 4 extends the basic AFSR structure to a phase-controllable one. It has been initially presented with the possibility to manipulate the phase response at the reflection band of an AFSR structure. This possibility is then utilized as a guidance to the design of a low-RCS parabolic reflector antenna and low-RCS reflectarray antennas with normal and offset feedings.

Chapter 5 focuses on the application of AFSR structure in RCS reduction of a high gain and beam steerable antenna. As an example, AFSR unit cell and dipole antenna are co-located into the same unit cell to create a low RCS phased-array element. Experimental results illustrate the effectiveness of AFSR structure in RCS reduction of a phased-array antenna with a maintained radiation performance.

Chapter 6 addresses the issue of creating a transmitting band within a wide absorption band to serve as a low RCS antenna radome. Two different designs of AFST structures based on hybrid resonators and integrated FSSs are proposed, which have shown significant improvement in the transmission and absorption performances compared with conventional methods.

Chapter 7 draws the conclusions for the thesis and proposes potential topics for further investigations.

Chapter 2

Literature Review

2.1 Antennas of Low Radar Cross Section

2.1.1 Radar Cross Section Reduction Techniques

For the purpose of increasing the survivability and reducing the detectability of a target or an object, the control of their radar cross section (RCS) is one of the important design considerations. The RCS of an object or a target can possibly be reduced by modifying its shape or size, reducing the reflectivity of its surface, and changing the direction of scattered power. Different from a passive structure, the RCS reduction of an antenna should also take its radiation performance into considerations. Therefore, it is a challenge to modify the shape or the size of an antenna while maintaining its original radiation performance. In this section, a generalized review of RCS reduction techniques that might be potentially applied in the RCS reduction of antennas is introduced.

RCS reduction based on absorbing materials

The required RCS reduction is possible by applying radar absorbing materials (RAM) to convert electromagnetic energy into heat. The absorption mechanisms of RAM rely on the dielectric loss, magnetic loss and ohmic loss. For many applications, it is very attractive to construct an absorber having small thickness and wide absorption bandwidth. The required absorption response can be practically obtained by using arrays of lumped resistors [9–12] or resistive sheets [13–17],

which are placed at a quarter-wavelength from the perfect electric conductor (PEC) plate. These designs are often referred to as Salisbury screens [18–20] or circuit analog (CA) absorbers [21]. Meanwhile, for magnetic absorbers, the corresponding lossy sheet with magnetic loss can be directly attached to the PEC surface. The difference between ways of placing lossy layers results from the reason that the most efficient way of absorption is to place the lossy layer corresponding to the maximum field strength. For example, the electric field reaches its maximum at a quarter-wavelength from the PEC plate, while the magnetic field reaches the maximum at the surface of its ground plane. Therefore, magnetic absorbers can be made with a thinner profile and still exhibit wider absorption bandwidth.

FIGURE 2.1: Microwave absorbing structures. (a) Multi-layer circuit analog absorber based on lumped resistors [9]. (b) Reflection reduction by using cascaded broadband magnetic absorber with polymer composites [22]. (c) Hybrid absorber with magnetic material and metamaterials for broadband absorption [23]. (d) The layout of wideband metasurface Salisbury screen with optimized reflection phase of ground plane [16]. (e) Illustration of ultra-broadband metamaterial absorber in a pyramidal shape [24]. (f) Conformal and transparent microwave absorber based on indium tin oxide (ITO) films [15].

For both magnetic and non-magnetic absorbers, there is a fundamental limitation on the absorption bandwidth for a given thickness and reflection level, which was derived by Rozanov [25]. This limit defines the relationship of maximum absorption bandwidth, the thickness of absorber, and the angle of incidence. Many efforts have been made to approaching this limit by means of optimizing the design of circuit

analog absorber. Multiple resonances are created and then combined [10, 12] in order to extend the bandwidth of absorption. Apart from that, the reflection phase of a backplane is varied with resonating elements to provided multiple resonances of absorption peaks [16]. In addition, it was further demonstrated that multiple layers of absorbing layers have been applied in the designs [26, 27] providing at least four combined absorption resonances. Many additional possibilities are available, if one integrates magnetic absorbers and circuit analog absorbers with complementary absorption bands. In the case of hybrid absorber designs [23, 28, 29], the absorption responses are taken from both electric and magnetic losses, which can help to expand the absorption bandwidth. Apart from that, some nanomaterials based on the carbon nanotube [30], carbon fibers and other composite material [31] are applied to produce electric losses for the absorption of electromagnetic waves at the microwave frequency.

TABLE 2.1: Comparison of typical microwave absorbers in literature

Design concept	Absorption band (GHz)	Thickness λ_L	Absorption bandwidth ¹	Absorption rate	Reference
Circuit analog absorbers	3.6 - 14	0.097	118%	> 90%	[32]
	2 - 9	0.088	127%	> 90%	[10]
	6 - 30	0.12	133%	> 85%	[16]
	1.51 - 10.1	0.1	148%	> 90%	[12]
	2 - 16.8	0.096	157%	> 90%	[26]
	1 - 13.3	0.1	172%	> 90%	[27]
Magnetic absorbers	1 - 4	0.021	120%	> 90%	[33]
	6.3 - 17.3	0.05	93.2%	> 90%	[34]
	3.19 - 18	0.053	139.7%	> 90%	[35]
	4 - 40	0.05	163.6%	> 90%	[22]
Hybrid absorbers	8.5 - 16	0.057	60%	> 90%	[28]
	4.5 - 18	0.135	120%	> 90%	[23]
	4.2 - 40	0.114	161.9%	> 90%	[29]

¹ The bandwidth is for -10 dB reflectivity.

In summary, the absorption curve for most of microwave absorbers are fixed, once they have been manufactured. For commercial absorbers relying on the dielectric losses and magnetic losses, the modification of their original loss curve is challenging. Therefore, some resonating structures operate in free space might be co-designed with the existing absorber to get the tailored absorption curve that we are desired.

RCS reduction based on metasurfaces

The idea of RCS reduction with shaping is to modify the directivity of scattering patterns and produce the scattering nulls towards the threatening direction. Similar to the shaping techniques, metasurface with specific patterns and configurations have been successfully employed for the RCS reduction of a surface or a structure without changing its original shape. The operating principle of this diffusive scattering metasurfaces is to manipulated the phase or polarizations of the scattering waves. One particular metasurface for controlling the reflection phase is known as high impedance surface (HIS) or artificial magnetic conductors (AMC), which has a great importance in microwave and antenna applications. The design concept of HIS was firstly proposed in [36], which consists of mushroom like structures with periodic array of metallic patches and shorting via holes. The use of such a structure is capable of producing an open-circuit resonance with a high surface impedance $|Z_s|$, leading to a total reflection with 0° reflection phase. In contrast, the reflection phase of a PEC surface is constantly -180° . Therefore, considering the cases of plane wave incidence, the reflection phase difference between PEC and AMC surfaces is characterized by 180° , which can be utilized for a scattering diffusion. There are serval approaches of designing AMC structure in the literature. Most of them are implemented by a periodic array of metallic patches loaded above a PEC layer, which is supported by a layer of dielectric substrate. The first design of PEC and AMC subarrays arranged in chessboard configurations for RCS reduction was reported in [37]. In fact, the RCS reduction bandwidth of the original chessboard design is restricted by serval factors including narrow phase range and angular sensitivity, which are associated with the resonance of AMC cells. Furthermore, improvements are made based on two AMC cells with a tailored frequency dispersion to provide wideband scattering reduction [32, 38–44]. It is noted that it is also possible to further broaden the reduction bandwidth by performing optimization algorithms on the arrangement of cells and varying the height of scatterers [45].

In the above mentioned AMC surface for RCS reduction, the thickness and bandwidth are restricted. The states of polarization are also an important parameter of electromagnetic waves. In order to manipulate the polarization states of reflected or transmitted waves, devices named polarization conversion structure (PCS) are designed. One particular reflective PCS is to achieve 90° polarization rotation for

FIGURE 2.2: Low scattering surfaces based on artificial magnetic conductor (AMC) and polarization conversion metasurface (PCM). (a) Chessboard structure consists of AMC and PEC elements [37]. (b) Broadband chessboard structure with two different AMC cells [42]. (c) Unit cell of chessboard structure based on two cross shapes [39]. (d) Photograph of fabricated low scattering PCM and its corresponding coding sequence [46]. (e) Schematic unit cell of PCM for RCS reduction. Double heads arrow are arranged rotationally to generate destructive phase cancellation between the neighboring cells. [47].

linearly polarized (LP) waves. These devices consist of metallic patterns and metallic ground plane, which are printed on the opposite sides of a dielectric substrate. The thickness of a single-layer reflective PCS can be reduced to less than $\lambda/10$ and the polarization rotation bandwidth is capable of exceeding 100%, as it was proposed in [48]. One distinct feature of PCS cells is that their rotating angles $\Delta\varphi$ will add to a reflection phase of $2\Delta\varphi$, which is well known as the geometry phase or Pancharatnam-Berry phase [49–51]. Furthermore, this linked relationship acts as a simple and powerful tool to obtain arbitrary reflection phase within the 360° range. For example, the requirement of 180° phase difference can be equivalently fulfilled by rotating 90° or mirroring the reference PCS cells, as shown in Figure 2.2(d) and (e). By arranging the original and rotated PCS cells in a chessboard configuration, the RCS of these surfaces can be reduced [46, 47, 52–55]. The advantage of using PCM for RCS reduction is that the design procedure of unit cells can be simplified

through rotating the existing design. Moreover, compared with AMC cells, PCM cells are more advantageous in a wider reduction bandwidth and a thinner profile.

In summary, in terms of RCS reduction performance, the diffusive scattering metasurfaces are more attractive than the microwave absorbers in their profile and weight. However, the bi-static RCS reduction performance for the diffusive scattering metasurface is not as good as microwave absorbers, which can be improved by using hybrid techniques.

RCS reduction based on hybrid techniques

TABLE 2.2: Comparison of the existing low scattering surface designs in the literature

Design concept	Frequency band (GHz)	Thickness λ_L^1	Reduction bandwidth ²	Arrangement of cells	Reference
AMC Cells	14.5 - 21.8	0.06	41%	Chessboard	[39]
	4.1 - 7.6	0.086	60%	Chessboard	[32]
	7 - 13	0.093	60%	3-bit coding	[40]
	9.4 - 23.3	0.07	85%	Chessboard	[41]
	3.8 - 10	0.08	91%	Chessboard	[42]
	3.8 - 10.7	0.08	95%	Chessboard	[43]
	5.5 - 32.3	0.11	141.8%	4-bit coding	[45]
PCM Cells	17 - 42	0.085	84.8%	1-bit coding	[52]
	7.9 - 20.8	0.08	89.8%	3-bit coding	[53]
	6.1 - 17.8	0.12	98%	Chessboard	[54]
	8.2 - 25.6	0.065	102%	2-bit coding	[55]
	5.1 - 22.1	0.054	125%	4-bit coding	[46]
	9 - 40	0.075	126.5%	2-bit coding	[47]
Hybrid Cells	10.5 - 18	0.23	52.6%	Chessboard	[56]
	13 - 31.5	0.09	83.1%	4-bit coding	[57]
	6.5 - 20	0.08	101.8%	2-bit coding	[58]
	3.15 - 18	0.06	140%	Randomly coding	[59]

¹ The wavelength is calculated based on the lowest operating frequency.

² The bandwidth is for 10 dB reflection reduction.

An alternative approach to the development of low scattering surface is to combine the above mentioned techniques including AMC, PCM, and resistive coatings into the same unit cell [56–59]. A similar idea has been demonstrated with a hybrid unit cells comprising three different scatters and generating an attractive reduction bandwidth with a low profile [59]. In fact, the hybrid low scattering surface can

also be implemented by combining mechanisms of polarization conversion and absorption [57]. This allows the functionalities of absorption and diffusion to operate inside the same unit cell in order to achieve a broadband scattering reduction.

The low scattering surfaces discussed above are summarized in Table 2.2 showing that a variety of design concepts can be implemented for the scattering control of a planar surface. Therefore, functional metasurfaces are not the only suitable to scattering reduction in the microwave regime, but also able to provide possibilities that conventional absorbers are not able to obtain.

In summary, the demonstrated RCS reduction techniques serve as a basis for low RCS antenna designs. As a special case of RCS reduction, the design of a low RCS antenna should also take its radiation performance into the considerations. Therefore, the selection of absorbing material should be very carefully and additional structures might be required to avoid the losses at the operating band of antennas. In addition, the placement of low RCS metasurface should also be well selected in order not to affect the radiation performance of antennas.

2.1.2 Antenna Radar Cross Section Reduction

The scattering cross section of antennas has been investigated since the early 1950s. After the Second World War, efforts were devoted to characterizing and categorizing the radar echoing of the existing military vehicles and naval vessels. During the progress of these studies, it is identified that antennas have contributed to a prominent scattering cross section. Therefore, researchers began to think the possible solutions to reduce the scattering cross sections of antennas, while maintaining their radiation performance. This has started the study on RCS reduction of antennas. In this section, the advantages and limitations of existing low RCS antenna designs will be presented by absorptive and non-absorptive techniques.

Absorbers in RCS reduction of antenna

The first effort to reduce antenna RCS is using radar absorbing materials (RAM). The goal is to cover antennas with some lossy materials to reduce their scattering cross sections. The design examples of absorptive superstates are illustrated in Figure 2.3. In [60], a resistive sheet was applied at the edges of an antenna to reduce

the reflections. An alternative method of reducing RCS of a slot array antenna is to cover their radiation apertures with resistive sheets [5] or lumped resistors [61]. In fact, in order to maintain the radiation performance, the operating frequencies of applied absorbers and antennas are located at separate bands [4, 5, 62]. An alternative approach of reducing antenna scattering is to use lossy coating at the vacant area of the antennas [63–65]. The idea of using absorptive coating was further developed into a low scattering antenna radome [66, 67], by introducing a passband within the absorption band. The presented low RCS circularly polarized patch array in [67] consists of two layers. The co-located patch element and the absorptive superstrate are shown in Figure 2.3(c). Slot resonators are designed to bypass the losses, which allows the superstrate to be totally transparent at the radiation band.

FIGURE 2.3: Antenna RCS reduction by using absorptive superstrates. (a) RCS reduction of array with resistive sheet [60]. (b) Microstrip feed slot antenna array with the lossy superstrate [5]. (c) Co-located unit cell of circular polarized patch antenna and frequency-selective rasorber structure [67].

In summary, the presented designs of low RCS antenna are based on utilizing absorbers as the superstrate of antennas. However, at the antenna's operating band, the lossy superstrate does not contribute to the radiation of antenna. This may limit its applications when high gain and high aperture efficiency are required for a low RCS antenna.

Metasurfaces in RCS reduction of antenna

Recently, interest in designing functional metasurfaces has grown considerably after a large number of lossless metasurfaces with low scattering were developed. The requirement of RCS reduction for planar or curved surfaces arises from some special applications such as an aircraft or an airborne platform, where the aerodynamic

characteristics are required. Therefore, the restrictions on shape and profile have posted a challenge in conventional RCS reduction methods. One possible approach of antenna RCS reduction is to use artificial magnetic conductor (AMC) structure. In [68], AMC cell and PEC cell were designed with 180° phase difference, which are loaded around the patch antenna in a chessboard configuration for RCS reduction. The advantage of using AMC structures is their low profile and easy to be conformal with the target antenna. The quantity and arrangement of AMC cells can be further optimized in [6, 69–71], in order to obtain a better performance of scattering reduction. The 3-D scattering pattern of slot array antenna in [70] with and without diffusive metasurface coatings are compared in Figure 2.4(f). It is observed that a significant scattering reduction is obtained by the diffusive coating surfaces. A similar idea has been demonstrated with polarization conversion metasurface (PCM) to reduce the RCS of patch antenna [7, 8, 72, 73]. The proposed PCM unit cells are capable of achieving linear polarization rotation. The operating bandwidth of polarization conversion is further expanded by designing multiple resonators inside the same unit cell. These unit cells are grouped into different sub-arrays with reduced RCS through randomly rotated polarization states. A comparison example of 3-D scattering pattern in [8] for a reference patch antenna with and without PCM is demonstrated in Figure 2.4(d) and (e).

In recent years, achieving high gain antenna with a compact structure has been a major concern for antenna designers. One typical solution is the Fabry-Perot (F-P) cavity antenna, which is featured with high gain and a low profile. The advantage of F-P antenna is that it does not require feeding networks and external feeds, leading to a simplified structure. The design concept of a low-RCS F-P antenna was firstly proposed in [74]. In the presented F-P antenna with reduced RCS, the partially reflecting surface (PRS) is modified from one layer into a two-layer structure. Two different metallic patterns are designed at both sides of PRS, serving functionalities of gain enhancement and incident wave absorption, simultaneously. Alternatively, the lossy side of the PRS can be replaced with a chessboard surface [75], coding metasurface [76], and polarization conversion surface [77] to reduce the RCS. The design of low RCS F-P antenna is further improved by introducing tunable reflection phase cells at the ground plane [78], achieving the frequency-reconfigurable functionality.

FIGURE 2.4: Antenna RCS reduction by coating lossless metasurfaces. (a) Low RCS microstrip patch antenna based on holographic metssaurfaces [6]. (b) Circularly polarized patch array enhanced gain and reduced RCS by using polarization conversion EBG structure [7]. (c) RCS reduction of Patch antenna with polarization conversion metasurface (PCM) [72]. (d) 3-D scattering pattern of original patch antenna [8]. (e) Suppressed scattering pattern of patch antenna with PCM [8]. (f) Ridged waveguide slot array with reduced in-band RCS by using metasurface bars [70].

In summary, the combination of existing RCS reduction techniques into the existing design of antennas has demonstrated a low RCS performance in the literature. However, problems are remain exists for the requirements of high gain, high radiation efficiency and low RCS. The main limitation of the metasurface is that they are not able to enhance the radiation performance of antennas at the in-band frequency. The aperture efficiency of antennas are decreased when combined with the metasurface for RCS reduction. In addition, aperture antennas such as parabolic reflector antennas and reflectarray, the solution to a low scattering reflector with maintained in-band reflection property has not been proposed. Meanwhile, low scattering antenna with beam steering performance is also highly desired in a communication systems with a stealthy requirement. Furthermore, new functionalities and designs of low RCS antenna can be demonstrated by integrating absorptive FSS into the design of low RCS antennas.

2.2 Absorptive Frequency-Selective Structures

For many types of lossless FSS, their frequency responses are limited to providing only transmission and reflection. Apart from these frequency responses, absorption is also an important functionality. Therefore, the frequency responses of transmission and reflection can possibly be combined with absorption response to achieve new functionalities. Based on this concept, the AFST structures and AFSR structures are proposed with unique frequency responses of absorption-transmission-absorption and absorption-reflection-absorption. Meanwhile, by introducing the out-of-band absorption performance, new applications of AFST and AFSR structures will be demonstrated in the RCS reduction of antennas.

2.2.1 Absorptive Frequency-Selective Transmission Structures

As described in the previous sub-section, Bandpass FSS relies on the induced electric current on an aperture to generate a passband for the incident waves. Therefore, at the in-band frequencies of bandpass FSS, the incident wave is able to pass through with low insertion loss. While at the out-of-band frequencies, the incident waves are totally reflected, contributing to an increased level of scattering. However, in some circumstances, it is required that the out-of-band scattering from an antenna can be controlled while the in-band radiation of antenna is maintained. Based on this requirement, the conceptual idea of AFST structure was firstly proposed in [79]. The idea of an AFST structure is to serve as a frequency-selective radome to reduce the scattering of an antenna.

In the meantime, structures with similar performance of AFST were termed with different names in the literature, such as frequency selective rasorber (FSR) [80–98], frequency selective radome [99, 100] and absorptive/transmissive FSS [101–104]. The early implementations of AFST structure were coming from the modified design of Salisbury screen absorbers [18, 19, 105], which are originally a lossy sheet located above a conducting plate. In order to allow incident wave to pass through, it is possible to cut slots on the ground plane of circuit analog absorbers. For this type of AFST designs [99, 101, 102, 106], the insertion losses from the resistive sheets are not considered because the transmission band and absorption

FIGURE 2.5: Functionality of absorptive frequency-selective transmission structure [80]. (a) Original scattering of an object without AFST structure. (b) Reduction of scattering while maintaining in-band transmission with AFST or FSR structure.

band are separated with each other. However, the transmission band is restricted and located at either lower or upper bands of the absorption band. The significant characteristic of AFST structure that distinguishes from other types of FSS is the frequency response of absorption-transmission-absorption.

Based on the structural view, two categories of AFST design methods can be distinguished, which are based on cascading 2-D FSS structure [80, 82–85, 107–116] and utilizing 3-D FSS design [94–98, 117–121].

2-D AFST structure

The possibility of the transmission band is located within the absorption band should also be considered. In this case, the losses from the absorption layer will lead to an increase of insertion loss for AFST structures [101, 122]. Therefore, several techniques were required to minimize the insertion losses of AFST structures at their passband. In this case, open-circuit resonance and short circuit resonance are utilized to provide different ways of bypassing methods. For example, a transmission window was designed within the absorption band of circuit analog absorber in [82]. To control the absorption at the passband, resistors are in shunt with a pair of lumped LC resonators. The lossy components are then bypassed when the shunt resonators are in short-circuit resonance, as presented in [82]. However, the short circuit resonance approach has some disadvantages. Each lumped resistor

FIGURE 2.6: Absorptive frequency-selective transmission structures based on cascading lossless FSS and lossy layers. (a) Unit cell of AFST with cascaded two layers structure [82]. (b) Unit cell of AFST based on interdigital resonator [85]. (c) Unit cell of dual-polarized and dual-band AFST structure [86]. (d) Unit cell of AFST based on split-loop resonator [84]. (e) Hybrid resonator based AFST unit cell [83]. (f) Unit cell of dual-band absorber [87].

has to be paired with a shunt LC resonators, which will lead to a large quantity of lumped elements required per unit cell. An alternative approach based on open circuit resonance was then proposed in the literature. It can be demonstrated that the current flowing through the lossy elements can be truncated if the resonator is in open-circuit resonance. Therefore, the losses from the absorption layer is bypassed. The implementations of such required open-circuit resonators can be interdigital resonators in [85, 110, 114], loop resonators in [80, 83, 84], printed LC resonators in [87, 102, 123] and quarter-wavelength stub resonators in [116, 124]. It is noted that AFST structure can also be specially designed to have multiple transmission bands. The examples of dual-band AFST structures were reported in [86, 87].

In summary, for most of 2-D AFST designs, their design concept is based on cascading a lossy layer with a bandpass FSS layer. The absorption response is realized by utilizing lumped resistors, while various resonators are designed to bypass the loss at the transmission band. The layout of lumped elements and resonators for current designs can be further optimized with a reduced amount of lumped elements.

3-D AFST structure

FIGURE 2.7: Absorptive frequency-selective transmission structures based on 3-D FSS unit cell. (a) The schematic of 3-D frequency-selective absorber by loading an array of lumped resistors on bandpass FSS [98]. (b) Unit cell of AFST structure with improved insertion loss [118]. (c) Dual-polarized FSR based on multi-mode resonators [97]. (d) Unit cell of FSR using planar slotline structure [96]. (e) Geometry of 3-D FSR unit cell with extended upper absorption bandwidth [95]. (f) Unit cell of FSR structure by cascading 2-D lossy layer with 3-D bandstop FSS [94].

An alternative approach to design AFST structure is to combine 3-D bandpass FSS with absorbers inside the same unit cell. Probably, the first design based on this idea was reported in [98], where an array of lumped resistors was loaded at the microstrip line based bandpass FSS. For this specific design, bypassing techniques are not required because the lossless and lossy resonators are operating at two separate bands, respectively. A similar idea was later proposed in [118], where the circuit analog absorbers and bandpass FSS are co-operated inside the unit cell. The losses from absorbers can be bypassed using a lumped LC resonators at the passband. Therefore, incident waves are able to be selectively absorbed or passed through, which depend on the resonate frequencies of these resonators. Furthermore, this design concept was extended from a single-polarized structure to a dual-polarized one in [97]. Apart from that, a variety of other AFST structures

based on the concept of 3-D FSS were proposed, and they are summarized in Figure 2.7. Their operating principles rely on constructing independent lossless and lossy cavities. More specifically, transmission line resonators in 3-D FSS are introduced, which are characterized by supporting the propagation of multiple-modes [12, 125, 126]. In addition, the physical implementation of these transmission line resonators can be chosen from various forms, such as parallel-plate waveguide in [94, 95], slot line structure in [96], and microstrip line structure in [98]. A strong point of 3-D FSS based AFST structure is the capability of isolating the lossless and lossy cavities naturally, which leads to a simplified overall design. A weak point is that the profile of AFST structures is nearly half a wavelength at the passband frequency. Therefore, an AFST structure based on ferrite absorber and slow wave propagation structure was proposed in [119, 127]. This work utilizes a previously developed techniques of folded substrates [128] to reduce the profile of AFST structure for specific applications around the megahertz range.

In summary, the 3-D AFST structures are characterized by two independent transmission path and absorption path. The current design of absorption path can be improved by replacing lumped resistors with commercial absorbers in future designs for better absorption performance and power handling capacity.

2.2.2 Absorptive Frequency-Selective Reflection Structures

2-D and 3-D AFSR structure

Recently, research interest in AFSR has grown considerably with a number of new structures and devices developed in the microwave regime, which aims at reducing the RCS of antennas while maintaining its good radiation performance. For example, the earlier report of a device with AFSR-like frequency response was in [4], where a microstrip antenna is located on the top of a bandstop FSS and a flat layer of radar absorbing material (RAM). However, it only exhibits a one-sided absorption band, which limits its practical applications. An AFSR structure is characterized with a frequency response of absorption-reflection-absorption, which has two absorption bands located at the lower and upper sides of the reflection band. With the circuit analog absorber reported in [129], the absorption bandwidth was 105% by optimizing the induced current distribution. Based on this

In summary, the 3-D AFSR structure have the advantages of independent control of absorption and reflection responses, compared with the 2-D AFSR structure. In addition, the presented AFSR structure in the literature has not been demonstrated in the RCS reduction of antennas. Future research might be working on the applications of AFSR structure in the RCS reduction of antennas.

Design guidelines for AFSR structure

The understanding of its operating principle is critical for the design and optimizations of AFSR structures. Following the review of AFSR structures, their design methods can be summarized into two main categories of cascaded type [4, 131] and series type [118, 130, 132, 133, 135], which are named after the connection method between the absorber and resonator [131]. The design guidelines and their corresponding equivalent circuit model will be discussed in this section.

Each type of design has its own advantages and disadvantages. For the cascaded-type, its advantage is the separate design of absorption and reflection responses, which can greatly simplify the AFSR design. Moreover, once the absorber is chosen, more resonators [136] are possible to be integrated within the unit cell without degrading the absorption performance. Meanwhile, for the series type, the absorbers and resonators are designed to operate collectively, which is slightly more complicated than the cascaded-type. The absorption performance can be implemented with absorbing materials or lumped resistors. Most of the cascaded-type AFSR structures are utilizing absorbing materials because they are more simple and reliable. It is worth mentioning that they have better power handling capacity than the lumped resistors, which is more suitable for high power applications. However, absorbing materials also have the shortcomings of bulky and heavy, especially for magnetic absorbers [137–139]. The advantage of using lumped resistors for absorption is lightweight and easy to design. However, they are expensive, especially when large amount of high frequency resistors are needed. Moreover, the parasitic effect of lumped elements is not easy to predict due to themselves and the soldering process involved.

The design methods and operating principle of AFSR structures can analyzed with equivalent circuit models for gaining physical insight into their operating principles. The cascaded-type design and series-type are summarized here. The cascaded-type is based on a microwave absorbing material. The incident wave is

FIGURE 2.9: Equivalent circuit model and frequency responses of AFSR structure for cascaded-type and series-type designs. (a) Equivalent circuit model for cascaded-type design. (b) Reflection coefficients of cascaded-type design. (c) Equivalent circuit model for series-type design. (d) Reflection and absorption coefficients of series-type design.

stopped from penetrating into the absorbers by cascading an additional layer of bandstop FSS. At the desired reflection band, this bandstop FSS creates a short-circuit resonance to reflect the incident waves. At other frequencies, the bandstop FSS is almost transparent. Therefore, the incident wave can pass through the FSS and then is substantially absorbed by the absorber with very low reflection. A typical equivalent circuit model for the implementation of this cascaded-type AFSR structure is shown in Figure 2.9(a) and (b). It is seen that the series resonance of the LC resonator generates a reflection band within the absorption band. Meanwhile, the bandwidth of reflection is determined by the quality factor of LC resonators. Moreover, more resonators can be included in the cascaded-type AFSR design to expand the reflection bandwidth or create multiple reflection bands.

The series-type AFSR is based on a circuit analog absorber, where the absorbers and resonators are connected in series. Therefore, the induced current flowing through the resistor can be truncated by the resonators. Therefore, a reflection

band is created within the absorption band by introducing an open-circuit resonance to cut off the current path. The characteristic of series-type AFSR design and its equivalent circuit are shown in Figure 2.9(c) and (d). The absorption path is cut-off when the resonators are in open-circuit resonance, which can be obtained with a pair of parallel LC resonators. It is observed that the lossy resistors contribute to two absorption bands that are located at lower and higher frequencies, corresponding to f_L and f_U , respectively. Meanwhile, the shunt resonator I provides an open-circuit resonance at f_1 , which provides the first reflection band. The reflection bandwidth is further extended by creating a series resonance at frequency f_2 from Resonator II.

The series type AFSR structure is commonly implemented by connecting lumped resistor and resonators in series. The advantage of utilizing lumped resistor is that it is lightweight and the performance of absorption is easy to control. However, this the series type AFSR structure has a limited power handling capability. Therefore, it is more suitable to be integrated with low power platforms. Meanwhile, the cascaded AFSR structure is implemented by cascading resonators with absorbers. The limitation of power handling capability can be overcome by utilizing this cascaded structure.

To conclude this chapter, we have reviewed the history and the current trend of RCS reduction techniques. Two mainstream techniques of absorptive approach and non-absorptive approach are discussed. These general RCS reduction techniques not only deliver the understanding of physical principles, but also help in the design of low RCS antennas. In addition, this review has also covered recently developed new EM structures of AFST and AFSR. To review all functionalities of AFST and AFSR structures mentioned in the available literature is possible but hardly instructive. Here, we have selectively chosen those typical designs which correspond to the most important ideas that could be potential useful in our low RCS antenna designs. We hope that the present review will help to establish a linking bridge connecting the past and future of low RCS antenna designs.

Chapter 3

Low-RCS Monopole and Dipole Antennas

3.1 Introduction

It is well-known that most of microwave absorbers naturally produce reflections outside of its absorption band, mainly due to the existence of an impenetrable conducting plane. However, the idea of exploring possibilities to realize a reflection band or several reflection bands within a wide absorption band produces more freedom to manipulate the electromagnetic waves in the frequency domain. For a perfect electric conductor surface, it reflects incident waves of all frequencies and polarizations. The absorptive frequency selective reflection (AFSR) structure distinguishes from conventional microwave absorber and perfect electric conductor by introducing frequency-dependent loss inside it, which offers more potentials for applications in antenna systems.

3.2 Design Concept

As an illustrative example, Figure 3.1 shows the reflection and absorption performance of AFSR structure in the frequency domain. It is observed that a reflection band is located between two absorption bands. The exhibited reflection in the

FIGURE 3.1: Frequency responses of AFSR structure

FIGURE 3.2: Illustration of integrated AFSR structure with antenna

middle band is intended to be utilized as an antenna's ground plane, which provides necessary reflection at antenna's operating band. Meanwhile, two absorption bands are located at the upper and lower sides of the reflection band, providing a substantial absorption for the incident waves. Figure 3.2 illustrates the operating principle of the antenna-absorber system with this AFSR structure. Within the operation band of the antenna, AFSR reflects antenna's backward radiation, resulting in a nearly maintained or even enhanced radiation performance. While the out-of-band incident signals are absorbed by the AFSR, leading to a substantial reduction of RCS.

3.3 Implementation Details

Although many structures may provide the desired frequency selective reflection response. However, they are not able to achieve a two-sided wide absorption bands.

FIGURE 3.3: 3D-AFSR structure based on multi-resonators.

To realize the frequency response of absorption-reflection-absorption, a 2D-periodic array of multimode resonators may be used to construct the unit cell of AFSR structure. It may consist of three resonators, one is lossless resonator to provide a reflection band. The other two resonators are lossy ones, which operate at the lower and upper absorption bands, respectively. These resonators may be implemented with transmission line (TL) resonators and the frequency of resonance is controlled through the equivalent electric length of the resonators. Furthermore, detailed descriptions will be provided in the later sections, including steps of design and analysis.

3.3.1 Configuration of Unit Cells

Figure 3.3 shows the perspective view of a 3D-AFSR unit cell. It is observed that the AFSR unit cell contains vertical substrate, metallic strips, lumped resistors and PEC backplanes. It is noted that a horizontal metallic strip is soldered with a lumped resistor, which is located at the front side of the unit cell. Meanwhile, the back side of unit cell is covered with a PEC backplane, which is a common practice in circuit analog absorber to achieve good absorption performance. The lossy resonator is made up of both PEC backplane and lumped resistors. The desired lossless resonators are created with vertical metallic strips, which are attached to the dielectric substrates. The physical dimensions of 3D-AFSR unit cells is shown in Table 3.1. The utilized dielectric substrate material for the vertical metallic strips is RO4230 with a thickness of 1.524 mm and a dielectric constant $\epsilon_r = 3$. The resonant frequencies of lossless and lossy resonators are determined by the

physical length of l and h , respectively. For our given unit cell, the substrate thickness is thin and t/p is a relatively small value. Meanwhile, the substrate thickness t is comparable to the width of metallic strip w . Therefore, the fringing field for the air mode inside the dielectric can be neglected [140]. Therefore, the substrate mode is still supported and can be modeled as quasi-TEM mode like a microstrip line.

TABLE 3.1: Physical dimensions of 3D-AFSR unit cells.

Parameter	p	h	w	t	l
Dimensions (mm)	12	20	1	0.76	10.5

The basic structure of AFSR is a two-dimensional periodic array of vertical striplines resonators with periodic boundary condition (PBC). It is observed in Figure 3.4 that two metal strips are placed above and beneath two dielectric substrates, respectively. Due to the symmetry characteristic of PBC condition, the vertical strip can be mirrored as a double-sided parallel-strip line, assuming that the thickness of the strip is negligible. From another perspective, the PBC boundary condition between two strips can be further replaced with a virtual PEC boundary, while without changing the electric field distribution across the boundary. Therefore, a periodic unit cell of vertical striplines with periods p along the x -axis and y -axis can also be represented by a different boundary condition. The alternative boundary condition is defined by PEC boundary in the x -direction and PMC boundary in the y -direction. It is known that the unit cell in Figure 3.5 supports two basic quasi-TEM modes in Regions I and II [125], which can be implemented by either double-sided parallel-strip lines or two back-to-back microstrip lines.

3.3.2 Propagation Modes

The periodic array of striplines supports the propagation of two quasi-TEM modes, which can be exploited to design the multi-mode resonators in 3D-AFSR. Considering the incident wave with y -polarized electric field, the two versions of implementing of the unit cell are basically the same in terms of electric field distribution, which are illustrated in Figure 3.4. However, the reason for choosing double-sided parallel-strip line is that it supports a third mode for dual-polarized incident waves,

FIGURE 3.4: Unit cell of vertical stripline resonators with periodic boundary conditions.

FIGURE 3.5: Electric field distribution of substrate mode and air mode propagate inside the vertical double-sided stripline structures.

which means it can be easily expanded to a dual-polarized structure [136]. Meanwhile, the microstrip unit cell only operates with single-polarized incident waves, which supports two quasi-TEM modes in Regions I and II, respectively.

To understand the operation and field distribution of two quasi-TEM modes, the screenshot of electric field propagating inside a vertical stripline unit cell is plotted in Figure 3.5. It is observed that the electric field is mainly concentrated within two regions. The first region is located between the metallic strips and substrate, where the electric field is mainly confined within the dielectric substrate. This first mode is characterized with substrate mode for identification. The second region is defined as air mode, where electric field is mainly confined within the two metallic strips.

It is noted that the substrate thickness and width of metallic strip will affect the distribution of substrate mode and air mode. If the metallic strip width is the same as the period of unit cell, these two modes will be strictly confined in their respective regions. However, with the decreasing of the metallic strip width, electric fields propagating in the air region will leak some power into the substrate region. A comparison is made in Figure 3.6 between the original air mode and approximated air mode. The original air mode is the case with metallic strip width w equal to the period of unit cell p . For the approximate case, $t/p = 0.083$, $t/w = 1.32$, and the thickness of substrate $t = 1$ mm. It is seen that the propagation constant and characteristic impedance are almost the same before and after approximation, which has proved that the method of approximation is valid. It is also noted that the approximation cannot obtain accurate results for air mode propagating between two thin strips. A quasi-static approach is utilized in [141], which will give us an accurate result. This approximation method provides a simple and quick way to describe two modes with simplicity at the design stage.

3.3.3 Equivalent Circuit Model

As discussed in the previous section, the periodic array of striplines supports the propagation of two quasi-TEM modes. A microstrip line supports the propagation of substrate mode and a partially filled parallel-plate waveguide allows the propagation of air mode. Therefore, the required circuit resonators for 3D-AFSR structure are constructed by using these multimode resonators. It is noted that

the physical length of vertical strip is l , which is equivalent to the electric length of θ_1 at the operating frequency of 9.2 GHz. The distance between the lumped resistor and PEC back plane is h , which is equivalent to the electric length θ_2 at the same operating frequency. The electrical length θ_1 has a relationship of $\theta_1 = 2\theta_2$, in order to obtain the same reflection band at f_M for both lossless and lossy resonators.

FIGURE 3.6: Characteristic impedance and propagation constant of original and approximate air modes

Since the primary objective of our design is to produce a reflection band within a wide absorption band. The reflection band is created by using a lossless stripline resonator with an electric length of $\lambda/2$. The absorption band is related to the lossy resonator with an electric lengths of $\lambda/4$ and $3\lambda/4$, corresponding to its fundamental and harmonic resonances. For the ease of description, the absorption band is defined with frequencies f_L and f_U , and the reflection band is at f_M . The relationship between the reflection and absorption band is $f_L < f_M < f_U$, which is represented by a frequency response of absorption-reflection-absorption. At the reflection band f_M , a total reflection is expected with $|S_{11}| = 1$. At the lower and upper absorption frequencies f_L and f_U , the desired reflection coefficient reduction is more than 10 dB.

The equivalent circuit model of the proposed single-polarized 3D-AFSR structure is shown in Figure 3.7. The characteristic impedance of the substrate mode is Z_1 . According to the approximation made in the mode analysis, we assume that the characteristic impedance of the air mode Z_2 is equal to the impedance of partially

FIGURE 3.7: Equivalent circuit model of 3D-AFSR structure for single-polarized incident wave.

filled dielectric parallel plate waveguide. To gain insight into the operating principle of 3D-AFSR, three cases are discussed in detail.

1) At the lower frequency of f_L , $\theta_1 = 2\theta_2 = \pi/2$. It is known that a quarter wavelength transmission line section transforms an open circuit to a short circuit and vice versa [142]. Therefore, the short-circuited air mode resonator is equivalent to an open circuit. On the other hand, the open-circuited substrate mode resonators are equivalent to a short circuit. The total input impedance of Z_{in} will be represented by a resistive value. In this case, the ideal reflection coefficient $|S_{11}| = 0$, which means a substantial absorption. The absorption rate is calculated as $1 - |S_{11}|^2$, with $|S_{21}| = 0$ due to the existence of impenetrable PEC backplanes.

2) At the middle frequency of f_M , $\theta_1 = 2\theta_2 = \pi$. It is also known that the input impedance is equal to the load impedance for a half wavelength transmission line [141]. The lumped resistor is bypassed by a shunt short-circuited substrate mode and air mode resonator, due to the fact that the electric length between the resistor and PEC back plane is $\lambda/2$ and λ , respectively. In this case, the total impedance of Z_{in} when looking into Port 1 is close to zero, which produces a perfect reflection response with $|S_{11}| = 1$.

3) At the higher frequency of f_U , $\theta_1 = 2\theta_2 = 3\pi/2$. Both the electrical lengths of air mode resonator and substrate mode resonator are three-quarter wavelength. This case is the harmonic resonance of Case 1 which leads to an absorption at the higher frequency. The input impedance of Z_{in} remains to be a resistive value, which leads to a substantial absorption.

TABLE 3.2: Parameters for the equivalent circuit model of the 3D-AFSR structure.

Parameters	Values
Z_1	73Ω
Z_2	360Ω
C_d	$0.04 pF$
C_g	$0.003 pF$
θ_1	180° at 9.2 GHz
θ_2	180° at 15 GHz

The parameters for the equivalent circuit model of the 3D-AFSR structure is provided in Table 3.2. The calculation of characteristic impedances (Z_1 and Z_2) and electric lengths (θ_1 and θ_2) are obtained from their physical dimensions by using the equations of microstrip line and parallel plate waveguide, as described in [142]. The edge capacitance between the striplines [143] is calculated as

$$C_d = \epsilon_r \frac{w}{t} \frac{\tan \Delta\theta}{Z_0 \omega} \quad (3.1)$$

ϵ_r is the dielectric constant of the substrate, t is the substrate thickness, w is the width of strip, and ω is the angular frequency. $\tan \Delta\theta$ is further expressed by

$$\tan \Delta\theta = \frac{2p}{\lambda} \left(\frac{p-2t}{p} \ln \frac{p}{p-2t} + \frac{t}{p} \ln \frac{p}{t} \right) + S\left(\frac{2p}{\lambda}\right) - S\left(\frac{2t}{\lambda}\right) - S\left(\frac{2(p-2t)}{\lambda}\right) \quad (3.2)$$

where p is the period of unit cell and

$$S(x) = \sum_{n=1}^{\infty} [\sin^{-1}(x/n) - x/n] \quad (3.3)$$

The equivalent inductance of the horizontal metallic strip L_s is determined by its width w and physical length D , which is expressed by

$$L_s = \mu_0 \frac{D}{2\pi} \log \left(1 / \sin \left(\frac{\pi w}{2D} \right) \right) \quad (3.4)$$

A comparison is made between the full-wave simulation results using a commercial software CST and equivalent circuit analysis, as shown in Figure 3.8. The reflection coefficients are simulated under the normal incidence. The single-polarized 3D-AFSR structure exhibits 112.3% fractional bandwidth (from 1.6 GHz to 5.7 GHz) for the lower absorption band, and 35.3% bandwidth (from 9.1 GHz to 13

GHz) for the upper absorption band, respectively. The total profile of this 3D-AFSR is only $\lambda/10$ at its lowest absorption frequency. Meanwhile, a flat reflection is observed band from 7 GHz to 8 GHz. Furthermore, the two resonant dips realized by capacitive impedance matching are also well-predicted in the results equivalent circuit model. Good agreement have been achieved between results from EM simulation and equivalent circuit analysis, which have verified the validity of circuit model.

FIGURE 3.8: Reflection and absorption performance of 3D-AFSR with full-wave simulation and equivalent circuit analysis.

It is noticed that the single-polarized 3D-AFSR structure has a frequency response similar to Salisbury screen, which is characterized by the reflection band occurring after every half a wavelength. However, our structure is superior to Salisbury screen in terms of absorption bandwidth, reflection band flatness, and sharpness of transition from absorption to reflection. The reason is that our structure has introduced a series resistor and capacitor network for the reflection and absorption of incident waves. The proposed structure is slightly different by creating two resonant dips, which is due to the impedance of resonator changing from a capacitive one into an inductive one when the frequency increases.

3.3.4 Dual-Polarized Structure Extension

Since the double-side striplines are utilized for the implementation of single-polarized 3D-AFSR structure, it is easy to expand this design to a dual-polarized one by creating a symmetrical structure for both polarizations. In addition, four lumped

FIGURE 3.9: Unit cell and period array of dual-polarized 3D-AFSR structure

resistors are soldered on a metallic rectangular ring and they are effective for two orthogonal polarizations. Another array of vertical strip line sheets is added, which is based on the single-polarized 3D-AFSR version. The illustration of dual-polarized 3D-AFSR structure is plotted in Figure 3.9. It is noted that the structure is symmetrical and insensitive to the polarization of incident plane waves. The electrical length of substrate mode resonator can be controlled by strip length l and effective dielectric constant ϵ_{eff} of the substrate. The effective dielectric constant ϵ_{eff} can be reduced by drilling holes on the original substrates [144], which is given by

$$\epsilon_{eff} = \epsilon_r - \frac{\pi R^2}{D_x D_y} (\epsilon_r - 1) \quad (3.5)$$

where ϵ_r is the dielectric constant of the substrate and R is the diameter of the drilled hole. D_x and D_y are the distance between the adjacent holes. The reason for drilling holes on the dielectric substrates is to reduce the reflection from the substrates.

TABLE 3.3: Physical dimensions of the dual-polarized 3D-AFSR unit cells

Parameter	p	h	w	t	l	D
Dimensions (mm)	12	24	1	0.76	18	0.8

The simulated reflection coefficient for both TE (E-field in the y -polarized) and TM (E-field in the x -polarized) incident waves are shown in Figure 3.10. Under the normal incidence, the reflection coefficient of this dual-polarized AFSR is less than 10 dB from 1.4 GHz to 4.1 GHz and from 6.7 GHz to 9.5 GHz. Moreover, the insertion loss is smaller than 0.28 dB from 4.8 GHz to 5.6 GHz, which indicates that a relatively flat reflection band has been obtained. For TE-polarized obliquely incident wave, the reflection coefficient of 3D-AFSR structure changes with the increasing angle of oblique incidence. While for TM-polarized incident wave, 3D-AFSR's frequency response is nearly unchanged with the varying incident angles.

FIGURE 3.10: Reflection coefficients of 3D-AFSR for oblique incident waves

3.3.5 Effect of the Ground Plane Size

It is known that if the periodic array has sufficient number of unit cells, the frequency response of this finite array will be nearly maintained with a unit cell in an infinite array. However, in practice, it is impossible to create a periodic array with infinite number of unit cells. Meanwhile, for practical applications using AFSR structure as antenna ground plane, we also need to understand the relationship between the ground plane size and RCS reduction performance. Therefore, an analysis has been conducted by comparing the monostatic RCS of PEC and 3D-AFSR structure with different sizes, as shown in Figure 3.11. For a structure with the 5×5 unit cells structure, the period array has a length of λ . It is seen that the monostatic RCS reduction is more significant at the higher band due to a smaller wavelength at higher frequency. However, the monostatic RCS of AFSR is higher than PEC at the middle band, which is due to the effect of edge diffraction. For a larger structure with 15×15 unit cells, the length of the periodic array is

(a) 5×5 unit cells (b) 7×7 unit cells (c) 15×15 unit cells

FIGURE 3.11: Monostatic RCS of PEC and 3D-AFSR with different sizes

3λ at the center frequency of 5 GHz. It is observed that good RCS reduction has been achieved with more than 10 dBsm at both lower and upper bands. However, a large number of unit cells will increase the complexity and cost of the fabricated structure. Therefore, a compromise has been made between RCS reduction and size of the periodic array. It is found that 7×7 unit cells with an array size of

$1.4\lambda \times 1.4\lambda$ is a balanced choice for the implementation of our proposed AFSR structure. In terms of its performance, it has a 10 dBsm RCS reduction from 1.8 GHz to 4.1 GHz and 6.8 GHz to 9.1 GHz for a periodic array of 7×7 unit cells. In addition, it also has a flat reflection band from 4.7 GHz to 5.9 GHz. It is considered as a suitable candidate for an antenna's ground plane or reflector.

3.4 Experiment and Results

To verify the effectiveness of our AFSR structure in the RCS reduction of antennas, we choose monopole antenna and dipole antenna as examples for demonstration. To make a fair comparison, the operating frequency and ground plane sizes of PEC-backed and AFSR-backed antennas are the same. It is also noted that the monopole and dipole antennas in our design are using the same AFSR structure as the ground plane, which contains 7×7 unit cells and has a size of $84 \text{ mm} \times 84 \text{ mm}$. The radiation patterns and RCS are simulated with CST and then measured in the microwave anechoic chamber.

3.4.1 Monopole Antenna

The geometrical model of a monopole antenna above a 3D-AFSR structure is shown in Figure 3.12. The monopole antenna is located in the center of a 7×7 AFSR array. A coaxial line is utilized to penetrate the AFSR structure from the bottom to the top, in order to feed the monopole antenna. The length of the printed monopole is $h_1 = 13 \text{ mm}$ and the height of periodic AFSR array is $h_2 = 24 \text{ mm}$. The prototype of 2-D periodic array of AFSR structure contains 49 unit cells in both x and y directions. The fabricated prototype of AFSR-backed monopole antenna is shown in Figure 3.13. It is seen that the monopole antenna is a printed metallic strip, which is fabricated on a Rogers 3003 substrate with a relative dielectric constant of 3 and a thickness of 0.76 mm. To better understand the assembly process, these 3D unit cells are assembled by two types of PCB boards with predesigned notches for interlocking them together. The width of the notches is lightly larger than the thickness of substrates for a perfect fit-in. The designed PCB boards are fabricated on a Rogers 4230 substrate with a thickness of 1.524 mm. The top layer of an AFSR structure contains four lumped resistors in each unit cell.

FIGURE 3.12: Monopole antenna backed with 7×7 unit cells AFSR array.

These lumped resistor are soldered on a metallic ring, which is fabricated on a Rogers 5870 substrate with a relative dielectric constant of 2.2 and a thickness of 0.381 mm. It is noted that the required resistance of lumped resistor is 250Ω and it is difficult to find an exact one. Therefore, we use a thick film resistor ERJ-3EKF2490V manufactured by Panasonic, which has a resistance of 249Ω for the implementation. Compared with PEC-backed monopole, the profile of AFSR-backed monopole is slightly higher due to the existence of AFSR resonators. The $|S_{11}|$ of PEC-backed and AFSR-backed monopole antennas are measured with a vector network analyzer (VNA). Their radiation patterns are directly obtained from the antenna measurement system. The scattering cross sections are measured with a pair of horn antennas. One horn antenna is utilized as the transmitting antenna and the other is used as the receiving antenna. These horn antennas are connected to a VNA for post processing and extraction of scattered signals.

Figure 3.14 shows the comparison of simulated and measured $|S_{11}|$ of the proposed AFSR-backed monopole and PEC-backed monopole. For PEC-backed monopole, it has a $|S_{11}| \leq -10$ dB from 5.0 GHz to 6.0 GHz. For AFSR-backed monopole, the bandwidth is slightly narrow, with $|S_{11}| \leq -10$ dB from 5.3 GHz to 5.6 GHz. This narrow bandwidth is attributed to the resonant ground plane and the uneven current distribution on the top surface of AFSR tends to have a narrower bandwidth. The measured results are in good agreement with simulated results, and the small difference is attributed to the fabrication and measurement errors. It is also highlighted that the frequency range of ASFR-backed monopole is located within 3D-ASFR's reflection band, which is from 5.0 GHz to 6.0 GHz. The polarization of radiated electric field for monopole antenna is perpendicular to the top surface of 3D-AFSR, which can be treated as an oblique incidence. Since the reflection band of AFSR structure is stable for both normal and oblique incidences, at the reflection band of our AFSR structure, it can be equivalent to a PEC plate, which serves as the ground plane for the monopole antenna.

FIGURE 3.13: Fabricated prototype of AFSR-backed monopole antenna.

The radiation patterns of PEC-backed monopole and AFSR-backed monopole in the xoz -plane are shown in Figure 3.15. It is seen that good agreement has been achieved between simulated and measured patterns. For PEC-backed monopole, the measured maximum gain occurs at -45° with a gain of 3.5 dB. The measured maximum gain of AFSR-backed monopole is 4.4 dB at -45° . The reason for this 0.9 dB gain enhancement of the AFSR-backed monopole is due to an extended current

distribution on AFSR's surface, which will be detailed in the discussion in Section 3.4.3. The shape of radiation patterns remains nearly unchanged for PEC-backed and AFSR-backed monopole antennas.

FIGURE 3.14: Simulated and measured $|S_{11}|$ of PEC-backed and AFSR-backed monopole antenna.

FIGURE 3.15: Simulated and measured radiation patterns of PEC-backed and AFSR-backed monopole antenna.

The mono-static RCS results of both PEC-backed monopole and AFSR-backed monopole antennas are investigated by terminating the antennas with 50Ω resistors. A good agreement has been achieved between simulated and measured results, although a small variation may exist, which is caused by the errors during the process of fabrication and measurement procedures. For the in-band RCS from

5 GHz to 6 GHz, the AFSR-backed monopole has RCS similar to a PEC-backed monopole. Within this reflection band, the AFSR functions as a ground plane to maintain the radiation performance of the monopole antenna. For the out-of-band from 2 GHz to 5 GHz and from 5 GHz to 10 GHz, the 3D-AFSR serves as a wide-band absorber with two-sided absorption bands. Considering the performance of measured RCSR, more than 10 dB RCS reduction has been achieved from 2.3 GHz to 4.5 GHz and from 6.5 GHz to 10 GHz, representing fractional bandwidths of 64.7% and 42.2%, with regard to the center frequencies of 3.4 GHz and 8.25 GHz, respectively.

FIGURE 3.16: Simulated and measured mono-static RCS of PEC-backed and AFSR-backed monopole antenna.

3.4.2 Dipole Antenna

In order to verify the effectiveness of 3D-AFSR on antenna RCS reduction, another example is considered by using the horizontal electric dipole. The radiation performance and RCS results are compared between AFSR-backed dipole and PEC-backed dipole. The geometrical model of a printed dipole antenna integrated with AFSR structure is shown in Figure 3.17. The conventional half-wavelength dipole antenna is mounted at the center of the 7×7 AFSR array. The profile of AFSR array has a height $h_1 = 24$ mm, which is the same as the previously constructed one for low-RCS monopole antenna. The dipole antenna consists of two parts: feed balun and two thin arms, which are both printed on a Rogers 5870

substrate with a thickness of 0.381 mm. The arm of the dipole has a length of $l = 13.2$ mm and a width of 1 mm. To ensure a balanced feeding, the two arms of the dipole are fed with a tapered microstrip to stripline transition. The transition has a height of $h_2 = 14$ mm, which is the distance between dipole and the top surface of our AFSR structure. Figure 3.18 shows the fabricated AFSR structure integrated with a dipole antenna. In this case, AFSR structure is the same as the previously discussed monopole case, except the fact that the a dipole antenna of horizontal polarization is mounted. To make a fair comparison, all variables are well controlled including using the same dimensions for the ground plane size and dipole antenna size.

FIGURE 3.17: Dipole antenna backed with 7×7 unit cells AFSR array.

FIGURE 3.18: Fabricated prototype of AFSR-backed dipole antenna.

The distance between dipole and AFSR structure is a key factor, which will affect the radiation performance. Due to the fact that the AFSR can be regarded as PEC at its reflection band, a distance of nearly $1/4$ wavelengths is chosen to obtain

an in-phase superposition of antenna's radiated waves and AFSR reflected waves. Therefore, the directivity of dipole antenna is enhanced with AFSR structure at its in-band frequency. Meanwhile, at the out-of-band frequency of the dipole antenna, the AFSR structure is lossy and absorbs the incident waves. Compared with PEC-backed dipole, the ground plane reflects incident waves of all frequencies, leading to a large RCS. To make a fair comparison between these two antennas, their ground planes are of the same size. In Figure 3.19, the photo of the reference PEC-backed dipole antenna is fabricated and then measured as a benchmark.

FIGURE 3.19: Fabricated PEC-backed dipole antenna for reference.

FIGURE 3.20: Simulated and measured $|S_{11}|$ of PEC-backed and AFSR-backed dipole antenna.

Figure 3.20 shows the comparison of simulated and measured $|S_{11}|$ of the proposed AFSR-backed dipole and PEC-backed dipole. For the PEC-backed dipole, the measured $|S_{11}| \leq -10\text{dB}$ starts from 5.0 GHz to 6.2 GHz, which agrees well with simulation results. For our 3-D AFSR-backed dipole, the measured $|S_{11}|$ is slightly narrower but maintains less than -10 dB from 5.1 GHz to 5.9 GHz, which is close to the simulation results. The bandwidth of the AFSR-backed dipole is reduced, compared with the original one. The reason is because AFSR structure has a narrower bandwidth, which limits the overall bandwidth of the integrated antenna

and AFSR structure. This case is similar to a reflectarray antenna, which has a wideband horn antenna as a feed, but its actual bandwidth is limited by the resonant frequency of reflectarray elements.

FIGURE 3.21: Simulated and measured radiation patterns of PEC-backed and AFSR-backed dipole antenna.

FIGURE 3.22: Simulated and measured gain over frequency of PEC-backed and AFSR-backed dipole antennas.

The radiation patterns of both AFSR-backed dipole and PEC-backed dipole are shown in Figure 3.21. It is observed that the AFSR-backed dipole has 8 dB back lobe level higher than PEC-backed dipole. Apart from that, it is seen that two side lobes occurred at the $\pm 90^\circ$ in the H-plane patterns of AFSR-backed dipole. These back lobe and side lobes are attributed to a larger edge diffraction produced

FIGURE 3.23: Simulated and measured mono-static RCS of PEC-backed and AFSR-backed dipole antenna.

by the dipole mounted on an AFSR structure. To suppress the edge diffraction, a corrugated ground plane at the edges of 3-D AFSR may be introduced in the future design. The simulated and measured gain results of these two dipoles are plotted in Figure 3.22. It can be observed that the maximum gain of AFSR-backed dipole is 9.5 dB, while the PEC-backed dipole is 7.2 dB. For the PEC-backed dipole, although the aperture size is increasing with frequency, the gain has decreased due a higher dipole profile with a wider beam width. It is interesting to find that the AFSR-backed dipole has a higher gain than PEC-backed dipole within the operating bandwidth. Moreover, the beam width of its radiation pattern is narrower for the AFSR-backed dipole, leading to a 2.3 dB gain enhancement.

Scattering performance of both AFSR-backed dipole and PEC-backed dipole is tested by terminating these two antennas with $50\ \Omega$ lumped loads. It is mentioned that the lowest operating frequency for two horn antennas utilized in the RCS measurement system is 2 GHz, which limits the lowest measurement frequency. It is clear to see that measured results agree well with simulated ones. The measured results have some ripples around the simulated curve, which may be caused by the mutual coupling between transmitting antenna and receiving antenna in far-field test system. It can be seen in Figure 3.23 that PEC-backed dipole and AFSR-backed dipole have a similar RCS from 5.0 to 5.9 GHz, which indicates a reflective ground plane within the antenna's operating frequency. Moreover, two absorption bands are observed at both sides of the reflection band, which shows an absorptive

ground plane to reduce antenna's out-of-band RCS. It is seen that the measured RCSR is more than 10 dB from 2 GHz to 4.2 GHz and 6.6 GHz to 10 GHz, representing a fractional bandwidth of 71% and 41% at their center frequencies of 3.1 GHz and 8.3 GHz, respectively. At the end, the measured performance of 3D-AFSR backed antennas is summarized in Table 3.4. Compared with AFSR structure for monopole and dipole, a satisfactory RCS reduction has been achieved for both lower and upper bands without sacrificing the radiation efficiency of these antennas.

TABLE 3.4: Summary of AFSR-backed monopole and dipole antennas.

Antenna type	Lower RCS reduction bandwidth	Reflection bandwidth	Upper RCS reduction bandwidth	Gain enhancement
Monopole antenna	64.7 %	18.1 %	42.2 %	0.9 dB
Dipole antenna	71 %	14.5 %	41 %	2.3 dB

3.4.3 Discussions on Gain Enhancement

The advantage of integrating AFSR structure with monopole and dipole antennas is that RCS reduction and gain enhancement can be obtained, simultaneously. To gain a deeper understanding of this enhanced gain, the current distributions on both the PEC surface and the top surface of 3D-AFSR are plotted in Figure 3.24 and Figure 3.25, for AFSR-backed dipole and monopole antenna, respectively.

FIGURE 3.24: Current distribution of PEC and AFSR backed dipoles

FIGURE 3.25: Current distribution of PEC and AFSR backed monopoles

FIGURE 3.26: Reflection phase distribution of PEC and AFSR surface.

The effective area of current on the ground plane are marked with dotted line for a clear demonstration. It is observed that the current distribution for the PEC-backed dipole is more dense than the AFSR-backed dipole, which are concentrated within a circular region of the ground plane. Meanwhile, the current distribution for AFSR-backed dipole is more sparse, which mainly flows on the metallic grid of the AFSR structure. It is observed that the effective current on the ground plane takes a larger region, which explains the 2.3 dB gain enhancement and narrower beam width. For the case of the AFSR-backed monopole antennas, a similar way of analysis is utilized. It is seen that the region of current flowing on the top surface of AFSR structure is slightly larger than the PEC ground plane. Therefore, the gain enhancement for the monopole antenna is 0.9 dB, which is not very obvious. In addition, the phase distribution of the reflected electric field is investigated and then plotted in Figure 3.26. It is observed that PEC surface has a uniform phase distribution of -180° . Meanwhile, the reflection phase of AFSR surface has a

tapered distribution, which is similar to the phase distribution of a curved reflector. Therefore, the reflection phase distribution of the AFSR surface may also explain the gain enhancement of AFSR-backed dipole antenna.

3.5 Summary

This chapter has introduced the design concept of absorptive frequency selective reflector that can exhibit a flat reflection band and two wide absorption bands. The proposed AFSR structure is implemented by using multiple resonators in a 3-D unit cell, which operate collaboratively to show an absorption-reflection-absorption response. Equivalent circuit model and design guidelines have been provided to explain its operating principle. The potential of this new structure is verified by two examples of monopole and dipole antennas in reducing their out-of-band RCS. It has been demonstrated that this AFSR structure can be utilized as antennas' ground plane to achieve a significant RCS reduction without sacrificing their radiation efficiency. Moreover, gain enhancement is also realized for the designed low RCS monopole and dipole antennas.

Chapter 4

Low-RCS Reflector Antennas

4.1 Introduction and Motivations

High gain antennas are the key to increasing the signal to noise ratio for long-distance communications such as satellite-terrestrial links, deep-space networks, and radars. Previously, in order to increase the gain of an antenna, the straightforward solution was to enlarge the aperture size of an antenna. This method will also increase the scattering area of the antenna, resulting in an elevated RCS. Consequently, it is a challenge to design a high gain antenna with high radiation efficiency and low-RCS. Many investigations only focus on the RCS reduction of antennas, as briefed in the literature review of Chapter 2. Among them, some common practices are utilized, such as shaping the structure of antennas [145], applying radar absorbing materials (RAMs) [4], using antenna radome [99], and adopting cancellation techniques [146]. However, there are tradeoffs among the RCS reduction, degraded radiation performance of the antenna, and increased complexity of the system.

To date, several attempts have been demonstrated to achieve a high gain and low-RCS antenna. One solution is focused on reducing the RCS of antenna arrays [2]. It is based on designing superstrate layers to absorb the impinging waves [61], [5] or to deviate the scattered waves outside the operating frequency of antennas [64, 147, 148]. However, it may happen in some cases that the radiation performance of the antenna is degraded due to the effect of the lossy superstrate. Moreover, deviating structural scattering of antenna through coating lossless metasurface is

not effective for bistatic radars. Another solution to obtaining a high gain and compact structure is the Fabry-Perot (FP) cavity antenna [149]. The scattering of this antenna is mainly attributed to the partially reflecting surface (PRS), which is designed for a highly directive beam. Similar to the reported RCS reduction of an antenna array, the scattering from the PRS layer can also be reduced with lossy layers [150], [151], or diffused by metasurface coatings [76], [75]. However, one of the drawbacks of the FP cavity antenna is its inherently narrow bandwidth of operation, which is due to the resonance nature on which its operating principle is based.

It is known that the directivity and front-to-back ratio of an antenna can be increased when it is placed over a large ground plane. In this case, the RCS reduction of this antenna can be achieved by replacing the ground plane with a frequency-selective surface (FSS) [152–154]. The FSS reflects waves at in-band frequency of the antenna and transmits incident waves at out-of-band frequencies. However, the FSS-backed antenna is not suitable to be mounted on a platform because the back side of FSS has to be a free space environment. To overcome this drawback, we have introduced the absorptive frequency-selective reflection structures (AFSRs) in Chapter 3 with a frequency response of absorption-reflection-absorption. In the previously discussed examples, the operation principle of AFSR is relied on using 3-D stripline resonator to create a reflection band within the absorption band. Therefore, the AFSR structure provides a perfect electric conductor (PEC) within the operating frequency of an antenna, while it also behaves as an absorber at the out-of-band frequencies. As a result, the RCS of antennas with AFSR ground can be significantly reduced without sacrificing their radiation performance. Nevertheless, most of the proposed applications for the AFSR structure were only focused on utilizing its amplitude response. To the best of our knowledge, there are very few studies on exploiting the phase response at the reflection band. The aim of presented work in this chapter is to move a step further by designing a unit cell to control the amplitude and phase responses of an AFSR unit cell, simultaneously. More specifically, the reflection phase of an AFSR structure within its reflection band can be manipulated. Following this, the application of this phase-controllable AFSR structure is demonstrated with a low-RCS parabolic reflector antenna and a reflectarray.

4.2 Design Concept

The design concept of our proposed phase controllable absorptive frequency-selective reflector is shown in Figure 4.1. For an incident electromagnetic wave of frequency f_0 , it behaves as a PEC plane with total reflection. Meanwhile, the wavefront of reflected waves can be arbitrarily controlled by a specific pattern of reflection phase distribution. For the lower frequency band $f_0 - \Delta f$ or the upper frequency band $f_0 + \Delta f$, it acts as an absorber with good absorption performance. The proposed phase controlled AFSR structure is demonstrated with a two-step design. The first step is integrating a microwave absorber with a bandstop FSS. It is known that a good microwave absorber exhibits wide absorption band and large absorption ratio, which can be utilized as a suitable base for AFSR structure to provide an excellent absorption performance. Meanwhile, a bandstop FSS can stop incident waves from penetrating through it into the absorbers and generate an absorption-reflection-absorption response. It is noted that more degree of freedom is obtained with a separate design of bandstop FSS and absorbers. The second step is to manipulate the wavefront of the reflected waves. Theoretically, the phase of electromagnetic wave can be accumulated through the propagation path or compensated by a planar patch at its resonance. Since the reflection phase of each AFSR structure is controlled independently, it is capable of transforming a spherical wave into a planar wave and vice versa. The main difference from a conventional reflector antenna is that our AFSR one is able to keep a low scattering cross section.

One typical application of our proposed phase-controlled AFSR structure is to reduce the RCS of a parabolic reflector antenna or a reflectarray while maintaining its radiation performance. It is well known that the scattering of a reflectarray mainly consists of the reflector scattering and the feeding antenna scattering [21], [1]. In this case, only reflector scattering is taken into consideration because it takes the major part of antenna scattering at its out-of-band frequencies. For a conventional reflectarray backed by a metallic ground plane, all incident energy will be scattered if the loss is not taken into consideration. However, among all back-scattered components, only the re-radiated parts with adequately compensated reflection phase can form an in-phase wavefront in the far-field with a highly directive beam. The rest of the back-scattered components from non-resonant patches and ground plane will contribute to an elevated scattering level. Therefore, the key to achieving a low scattering parabolic reflector antenna or a reflectarray is to reduce its structural

FIGURE 4.1: Conceptual depiction of a low scattering reflective focusing surface based on phase controllable absorptive frequency-selective reflection structure.

cross sections while maintaining its good radiation performance. More specifically, the structural scattering from the antenna is reduced by introducing two absorption bands at its out-of-band frequencies. Meanwhile, at its in-band frequencies, the reflection phase from the bandstop FSS cascaded on a commercial absorber can also be flexibly manipulated for beam focusing.

4.3 Low-RCS Parabolic Reflector Antenna

Parabolic reflectors are the most commonly used high gain antennas for long distance communications with low-noise receiving systems. The shaping of a paraboloidal surface has the capabilities of realizing co-phased radiated wave by placing a feeding antenna at the location of its focus. The spherical wavefront of the feed is compensated into a planar one through the propagation path delay with a paraboloidal surface [155], [156]. When a wave is incident upon a parabolic reflector, all energy is scattered back due to the total reflection characteristic of an ideal electric conductor. Considering the back-scattered waves, only those components with appropriately compensated phase delay will be focused at the focal point and then received by the feed antenna. While the rest components are either reflected backwards or scattered to other directions, which will lead to an elevated level of backscattering. Therefore, the major scattering of a parabolic reflector antenna can be classified into reflector scattering and feed scattering. The feed scattering is related to the radiation performance of parabolic reflector at in-band frequencies. The structural scattering from reflector is dominant at the out-of-band frequencies. Therefore, the reduction of reflector's scattering can be realized while maintaining the radiation performance.

In this section, we aim at reducing the reflector scattering part by replacing parabolic-shape PEC reflector with a curved AFSR structure, which consists of 3-D bandstop FSS backed with an absorber. The radiation performance of antenna can be maintained, while its RCS can be reduced significantly. At the antenna's in-band frequencies, the curved AFSR structure can still reflect and then focus the incident waves. At the antenna's out-of-band frequencies, the bandstop FSS is almost transparent, and the incident waves can pass through it and be dissipated by the absorbers.

4.3.1 AFSR Unit Cell Design

The structure of AFSR unit cell is shown in Figure 4.2 and it consists of two parts. The upper part is a second-order bandstop FSS based on 3-D unit cells and the bottom part is a commercial absorber (Eccosorb FGM-125) backed with PEC. The second-order frequency response is produced by two quarter-wavelength

stripline resonators, which are shorted by via holes at the end of each strips. The substrate material used in this design is Rogers RT5880, which has a permittivity of 2.2 and thickness of 3.054 mm. The bandstop structure is similar to the structure presented in [157], except a metallic line is inserted between double-sided striplines. For the structure arranged with the same height, a virtual PEC boundary is formed naturally between two symmetrical strips. While this symmetry no longer exists when the unit cells are arranged at different heights. In order to maintain the frequency response of unit cell, the inserted vertical metallic strip serves a role of common ground for the asymmetric stripline resonators at its both sides. It is noted that the 3-D unit cell structure is symmetrical. Therefore, it is insensitive to the polarization of incident waves.

FIGURE 4.2: Structure of the AFSR with a dual-polarized bandstop FSS and a commercial absorber.

TABLE 4.1: Physical dimensions of AFSR structure for parabolic reflector

Parameter	h_1	h_2	h_3	d	p	t	D
Dimensions (mm)	16	6.15	2.5	3.2	12	3.048	0.8

The simulated reflection coefficients of the proposed unit cell under the normal and oblique incidences are shown in Figure 4.3. The 3-D bandstop FSS resonates at around 8.2 GHz, which rejects the signals passing through this structure at this frequency. The reflection response is flattened due to the effect of a second-order resonance, which is generated by cascading two short-ended resonators in the vertical direction. At other frequencies, the bandstop structure is almost transparent,

which allows signals to pass through and absorbed by the backed commercial absorber. The simulated results for both TE and TM polarizations are given for various angles of incidence. It is observed that the reflection is less than -10 dB from 2.1 to 6.1 GHz and 8.8 to 13.2 GHz for both polarizations, respectively. The frequency response of absorption-reflection-absorption is stable up to 45° oblique incident angles.

FIGURE 4.3: Simulated reflection coefficients for AFSR structure with TE and TM polarized incident waves under normal and oblique incidence.

4.3.2 Curved 3-D AFSR Reflector

Based on the principle of geometry optics, it has been shown that a parallel beam incident upon the parabolic reflector will converge at the focal point. In the same

manner, if a point source is located at the focus of parabola, the reflected rays will be transformed into a parallel beam. The surface of a paraboloidal reflector can be formed by rotating a parabola curve about the axis. Since the parabola curve is a continuous one, the generated reflector is also continuous with a smooth curvature. It is feasible in theory that geometry optics property will be maintained if a continuous parabolic surface is discretized into subwavelength sized reflectors, as shown in Figure 3.8. Conceptually, the surface of parabolic can be designed into a curved one or a staircase approximation of a continuous curvature to achieve reflection phase compensation based on their geometry. It is expected that the radiation performance of a continuous reflector and a discrete reflector will be nearly unchanged if the steps of discretization are small enough. Therefore, the PEC blocks for discrete reflector can also be replaced by our AFSR structure to produce an absorption-reflection-absorption response.

FIGURE 4.4: Surface geometry of continuous and discrete parabolic reflector

The parabola surface can be described by three parameters, the focal length F , the diameter D , and the distance between focus and reflector H . These parameters are related with each other based on the parabola equation

$$H = F - \frac{D^2}{4F} \quad (4.1)$$

The continuous paraboloid surface is formed by rotating the parabola curve around its axis. For a discrete paraboloid surface, the variation of its diameter is divided into n blocks and m blocks, in two orthogonal directions, respectively. Each block has a period of p . The distance H between the focal point to the discretized block is calculated as

$$H = F - \frac{(mp)^2 + (np)^2}{4F} \quad (4.2)$$

Therefore, the height variation of paraboloid surface is modified into a discontinuous one, as shown in Figure 4.4. The radiation performance of a discrete reflector will be maintained as a continuous one, if we choose a small step of p .

FIGURE 4.5: Geometry of two-dimensional parabolic reflector antenna based on curved 3-D bandstop frequency-selective structure.

For the design of a high-gain reflector antenna, the reflection phase of each unit cells should be gradually varied to produce a collimated beam, when it is illuminated with an external feed source. To produce such a phase shift, it can be implemented by varying at least one parameter of the unit cell. However, for the consideration of RCS reduction, changes in any dimension of the AFSR unit cell

will alter the frequency response. Here, instead of varying the dimensions of unit cell, spatial phase delay is used to compensate the desired phase. From a global view in Figure 4.5, the unit cells of AFSR structure are arranged with different heights and forming a parabolic surface to focus the beam. Locally, due to the gradual variation between unit cells, the frequency response of each unit cell is maintained.

TABLE 4.2: Dimensions of substrate height for 1-D array of unit cells

Parameters	Values (mm)
L_1	16.20
L_2	16.83
L_3	17.87
L_4	19.33
L_5	21.20
L_6	23.49
L_7	26.20
L_8	29.32
L_9	32.86
L_{10}	36.82

To clearly illustrate the gradual changes in the height of unit cell, Figure 4.5 plots a cross-sectional view of 1-D array of unit cells. Compared with traditional parabolic reflector antenna, the continuous parabolic curve is divided into several discrete steps. Each step is implemented by the proposed 3-D bandstop FSS unit cell. These unit cells are elevated with different heights to produce the required spatial phase delay. The dimensions of the substrate height for different unit cells are provided in Table 4.2. The minimum height of the curved AFSR reflector is 16.20 cm, which serves as the baseline of reflection phase for the entire reflector. The maximum height of the element is 36.82 cm at the edge of reflector. It is worth mentioning that this discrete 2-D parabolic surface is further implemented by interlocking horizontal and vertical substrate strips. Each strip has pre-designed curvature to form the surface of parabola after they are assembled together. The assembled parabolic reflector antenna is fed by a pyramidal horn with an aperture dimension of 57.2 mm $\times$ 44.1 mm. The diameter of the parabolic reflector is 255 mm, which consists of 349 unit cells. The focal length is 180 mm, which is calculated by the setting horn antenna at the phase center of the reflector.

4.3.3 Radiation and RCS Performance

FIGURE 4.6: Radiation patterns of parabolic reflector antenna with curved AFSR structure and PEC at 8.2 GHz.

FIGURE 4.7: Monostatic RCS of low RCS parabolic reflector antenna with curved AFSR and conventional reflector antenna.

The operating principles of the proposed low-RCS parabolic reflector antenna are verified by comparing it with a traditional parabolic reflector antenna. Figure 4.6 shows the radiation patterns of the proposed low-RCS parabolic reflector antenna and its counterparts. Both antennas have achieved good radiation performance, with simulated gain of 25.2 dB and 24.4 dB, respectively. The 3 dB beamwidth for the low-RCS one and conventional PEC one is 5° and 4° , respectively. The side lobe level (SLL) for the low-RCS parabolic reflector antenna and the traditional version are -19.5 dB and -20.3 dB, respectively. Compared with the PEC parabolic reflector antenna, it is seen that the radiation performance is nearly maintained

for the low-RCS parabolic reflector using curved AFSR structure. In Figure 4.7, compared with traditional parabolic reflector, the monostatic RCS of the proposed one has been reduced by about 10 dB at the out-of-band frequencies. The bi-static scattering patterns of two antennas in the backscattering direction are compared at three different frequencies. The scattering main lobe and side lobes of the low-RCS parabolic reflector are significantly smaller than the conventional one within the angular range from -30° to 30° at 4 GHz. At frequency of 6 and 9 GHz, due to the absorption effect of the curved AFSR reflector, the angular range of reduced scattering pattern is extended from -90° to 90° .

FIGURE 4.8: Simulated bi-static scattering patterns of a low-RCS parabolic reflector antenna and a conventional reflector antenna.

4.4 Low-RCS Reflectarray with Center Feed

The idea of using AFSR structure to reduce the RCS of large aperture radiators such as parabolic reflector antenna is introduced in Section 4.3. For some applications, the installation space is restricted and the profile of reflectors is limited to a lower one. It is more favorable to use a planar reflector instead of a curved one. Therefore, in this section, we focus on designing a planar AFSR structure to control the reflection amplitude and phase responses simultaneously. The potential

application of this phase-controlled AFSR unit cell is demonstrated with a high gain and low RCS reflectarray. Compared with the demonstrated parabolic reflector antenna, low-RCS reflectarray is more challenging since the reflecting surface is a planar one rather than a curved one.

4.4.1 Planar and Phase-Controllable AFSR Structure

The first step is to create a reflection band within a wide absorption band by designing a bandstop frequency-selective structure (FSS). The basic design of our bandstop FSS is originated from a thin arc-shaped metallic strip printed in the vertical direction (z -direction), which is equivalent to a series LC resonator for an incident electric field parallel to the strip. The equivalent inductance is induced from the metallic strip and the capacitance is formed by the gap between adjacent strips. As shown in Figure 4.9, a transmission zero is introduced at the resonant frequency of the strip resonator. The physical length of the metallic strip resonator is slightly shorter than half a wavelength at the resonant frequency f_c , which is approximately calculated as

$$f_c \approx \frac{c}{2l_s\sqrt{\varepsilon_{eff}}} \quad (4.3)$$

where c is the speed of electromagnetic wave in free space, ε_{eff} is the effective relative permittivity of the substrate material, and l_s is the physical length of the strip resonator.

FIGURE 4.9: Geometry of the unit cell with a single arc-shaped metallic strip resonator and its simulated reflection and transmission coefficients.

(a) Geometry and its equivalent circuit.

(b) Full-wave and circuit simulation results of bandstop FSS

(c) Coupling coefficient k of two resonators versus g .

FIGURE 4.10: Unit cell structure of a second-order 3-D bandstop FSS and its equivalent circuit model.

To obtain a relatively flat reflection band, two transmission zeros are created by coupling two arc-shaped metallic strip resonators face-to-face, as shown in Figure 4.10. The generated lower and higher transmission zeros are denoted as f_L and

f_H , respectively. The coupling between two resonators is described by a coupling coefficient k , which is calculated as

$$k = \frac{f_H^2 - f_L^2}{f_H^2 + f_L^2} \quad (4.4)$$

Since the size of two arc-shaped strip resonators is fixed, the coupling coefficient k is determined by the gap g between two resonators, as shown in Figure 4.10.

TABLE 4.3: Dimensions and circuit parameters of a second-order bandstop FSS.

Parameters	Values
p	18 mm
t	1.59 mm
g	3.5 mm
w	0.4 mm
R	5.9 mm
h_1	3.3 mm
h_2	14.5 mm
θ	161°
L_1	19.85 nH
L_2	19.27 nH
C_1	0.0335 pF
C_2	0.0243 pF
Z_0	377 Ω

The coupling coefficient k decreases as the gap g increases, leading to two transmission zeros getting closer, which plays an important role in determining the locations of transmission zeros at two different frequencies. The equivalent circuit model for the cascaded arc-shaped strip resonators is shown in Figure 4.10. For its simplicity, the effect of the dielectric in the vertical direction is not considered and the metal thickness is assumed to be zero. Some minor difference exists between the full-wave simulation and circuit model results, which may be attributed to the approximations made in simplified circuit model. The unit cell is equivalent to a pair of series LC resonators, spaced by an impedance inverter. Resonator I consists of L_1 and C_1 and Resonator II consists of L_2 and C_2 , respectively. Two transmission zeros are attributed by the series resonance of two resonators, which are calculated as

$$f_L = \frac{1}{2\pi\sqrt{L_1 C_1}} \quad (4.5)$$

$$f_H = \frac{1}{2\pi\sqrt{L_2 C_2}} \quad (4.6)$$

Resonator I behaves as a capacitor when $f < f_L$, while it is equivalent to an inductor when $f > f_L$. Similarly, for Resonator II, it is capacitive when $f < f_H$, while it is inductive when $f > f_H$. The resonator impedances for Z_{r1} and Z_{r2} are calculated as

$$Z_{r1} = j\omega L_1 \left(1 - \frac{1}{\omega^2 L_1 C_1} \right) \quad (4.7)$$

$$Z_{r2} = j\omega L_2 \left(1 - \frac{1}{\omega^2 L_2 C_2} \right) \quad (4.8)$$

For a frequency below f_L , Z_{r1} and Z_{r2} for Resonator I and Resonator II are both capacitive. While viewing from Port 1, the capacitor Z_{r2} is transformed into an inductor Z_{in} after quarter-wavelength transmission line. The capacitive part from Z_{r1} and the inductive part from Z_{in} form a parallel LC resonator, which leads to a transmission pole f_{L0} at the resonant frequency. For a frequency above f_H , another transmission pole f_{H0} can be analyzed by using the same way as the previous one. Two transmission poles f_{L0} and f_{H0} of the proposed bandstop FSS are calculated as

$$f_{L0} = \frac{1}{2\pi} \sqrt{\frac{M - \sqrt{M^2 - 4L_1 L_2 / C_1 C_2}}{2L_1 L_2}} \quad (4.9)$$

$$f_{H0} = \frac{1}{2\pi} \sqrt{\frac{M + \sqrt{M^2 - 4L_1 L_2 / C_1 C_2}}{2L_1 L_2}} \quad (4.10)$$

$$M = \frac{L_1 C_1 + L_2 C_2}{C_1 C_2} + Z_0^2 \quad (4.11)$$

Unlike the previous design of 3-D bandstop FSS [157], which is based on transmission line resonators from a multi-mode cavities structure with multi-propagation paths. Our proposed one here is based on metallic strips with multiple transmission zeros and transmission poles, which can be equivalent to LC resonators. The design process of our proposed bandstop FSS mainly relies on full-wave simulation and the equivalent circuit model only serves as an aid to understand the operating principle of this FSS. The main advantage of using arc-shaped resonators is because the adjustment of resonant frequency f_c , and the coupling coefficient k is not constrained by the period of the unit cell. For example, the resonant frequency of the resonator can be lowered by increasing the radius r or central angle θ of the arc. The extended dimension of strips is within the vertical plane (y - z plane). Therefore, the resonant frequency of metallic strip resonator and coupling between two resonators can be designed flexibly without changing the period of the unit

cell. Moreover, the layout of strip resonators in the vertical plane also maximally reduces the overlapping area with the phase shifting layer. As a result, the inter-layer coupling between the phase shifting layer and bandstop structure has been reduced for this type of bandstop FSS. In Table 4.3, the physical dimensions and parameters for the equivalent circuit model are provided.

(c) Simulated $|S_{11}|$ and absorptivity under the normal and oblique incidence

FIGURE 4.11: Unit cell structure of AFSR by cascading 3-D bandstop FSS over a commercial absorber backed with PEC plane.

To validate the analysis, a dual-polarized bandstop FSS with two transmission zeros and two transmission poles is designed, as shown in Figure 4.11. To reduce the polarization sensitivity of the unit cell, the resonators are arranged in both x -direction and y -direction for dual polarizations. The simulated frequency response and the dimensions of our proposed 3-D bandstop FSS are also provided. The

period of the unit cell is 18 mm. The arc-shaped strip resonator is designed on a RT5880 substrate with a dielectric constant of 2.2 and a thickness of 3.175 mm. The physical dimensions of AFSR structure based on 3-D bandstop FSS and commercial absorber are listed in Table 4.4. It is observed that the 3-D bandstop FSS has two reflection poles at the frequency of 6.3 GHz and 7.5 GHz, corresponding to a fractional bandwidth of 13.2% for less than 1 dB reflection loss. By cascading the proposed 3-D bandstop FSS with a wideband absorber from Eccosorb FGM-125, the frequency response of absorption-reflection-absorption can then be obtained. Figure 4.11 shows the resultant AFSR structure after integration. For the normal incidence, it is observed that it has a reflection band with less than 1 dB reflection loss from 6.0 GHz to 7.6 GHz. It is also seen that two absorption bands exist at the lower and upper reflection bands, with an absorptivity of more than 80%, corresponding to a fractional bandwidth of 53.6% and 19.4%, respectively. Furthermore, it is also seen that the reflection and absorption responses are stable for an oblique incidence angle up to 45°.

TABLE 4.4: Physical dimensions of AFSR structure for reflectarray.

Parameters	Values
p	18 mm
t	1.59 mm
g	3.5 mm
w	0.4 mm
r	5.9 mm
h_1	3.3 mm
h_2	14.5 mm

To control the reflection phase within the reflection band, metallic patches of varying size are designed to load above the AFSR structure, as shown in Figure 4.12. This unit cell is characterized by using a periodic boundary with an infinite array setting. At the reflection band of bandstop FSS, the power escaped from the initially excited patch resonator is larger than the power dissipated internally. This case can be used for the design of the phase shifting layer. By varying the patch size, the reflection phase of each unit cell also changes accordingly. While at the lower and upper absorption bands of the AFSR structure, the incident electromagnetic wave can partially pass through the phase shifting layer. After transmitting through the 3-D bandstop FSS at its passband, the incident electromagnetic wave is finally absorbed by the absorbing materials with no or little reflection. Since the unit cells of the phase shifting layer are rearranged in a specific pattern for

(a) Schematic model of AFSR unit cell with top-loaded patches

(b) Simulated $|S_{11}|$ and $\angle S_{11}$ with different size of patches. (c) Simulated $|S_{11}|$ and $\angle S_{11}$ with size variable patch at 7 GHz.

FIGURE 4.12: Simulated reflection magnitude and phase of low-RCS reflectarray unit cell under the normal incidence. (The shadow region marks the suitable operating frequency range for low-RCS reflectarray)

the co-phased electrical field, the absorption rate of each cell also follows a similar distribution. The overall scattering from our proposed phase shifting layer backed with the AFSR structure is therefore significantly reduced, when compared with the same array backed with a PEC plane.

The designed phase shifting layer for the AFSR structure with size-varying patches is shown in Figure 4.12. It is demonstrated that a reflection band is maintained at 7 GHz with lower and upper absorption bands located at its both sides. It is observed that the absorptivity of two absorption bands is slightly reduced when increasing the size of the patch. This is expected because the larger size of patches in a unit cell comes with more reflection. In this case, the lower absorption band is less affected than the upper absorption band, which may be attributed to the fact that the length of elements is smaller in terms of the wavelength at lower frequencies. Through a quick parametric scan of the patch size, it is found that the reflection loss is less than 1 dB at 7 GHz and it has a reflection phase range of about 250° . It is understood that the reflection amplitude and phase responses are related to the structure and material properties below the patch [158]. In this case, the grounded dielectric substrate is replaced with a bandstop FSS, resulting in a reduced quality factor of the resonant structure and reduced element phase shift ranges.

4.4.2 Reflection Phase Mapping

Two reflectarrays are designed with a detailed comparison of their radiation and RCS reduction characteristics. The first one is based on our proposed low scattering reflectarray. The second one is based on conventional reflectarray elements, which is served as a benchmark for assessing the performance of our low-RCS reflectarray. Both antennas are designed for operating at 7 GHz with the same aperture size and feed antenna for the convenience of comparison.

The required reflection phase of a reflectarray element is designed to compensate the spatial phase delay of a planar reflecting surface. Therefore, for a specific reflection phase distribution, the beam of the reflected wave can be designed to point towards a specific direction. Based on the phase analysis in [159, 160], the required reflection phase ψ_i for each reflectarray element with a broadside beam is

FIGURE 4.13: Reflection phase calculation for each elements of a reflectarray.

calculated as

$$\psi_i = k_0(R_i - \vec{r}_i \cdot \vec{r}_0) + \psi_0 \quad (4.12)$$

where k_0 is the wavenumber in free space. R_i is the distance between the phase center and the element. $\vec{r}_i$ is the vector pointing from the central element to the i th element. $\vec{r}_0$ is the unit vector along the direction of the collimated beam. ψ_0 indicates a relative reflection phase for reference. As shown in Figure 4.13, a staircase approximation of circle is applied in the calculation of reflection phase for each element of reflectarray.

To obtain an optimal aperture efficiency of reflectarray, two crucial factors of tapering efficiency η_{taper} and spillover efficiency η_{sp} are considered. In the classic design of reflectarray, horn antennas are utilized as the illuminators to feed the reflectarray. However, the radiation pattern of horn antenna is not an omnidirectional one, which is characterized by an amplitude taper with a $\cos^n$ -like function. The equations of taper efficiency η_{taper} for the non-uniform illumination is given by

$$\eta_{taper} = \frac{1}{\pi} \cot^2 \left(\frac{\theta_0}{2} \right) \frac{\left| \int_0^{\theta_0} \int_0^{2\pi} \sqrt{G_{feed}(\theta, \varphi)} \tan \left(\frac{\theta}{2} \right) d\theta d\varphi \right|^2}{\int_0^{\theta_0} \int_0^{2\pi} G_{feed}(\theta, \varphi) \sin(\theta) d\theta d\varphi} \quad (4.13)$$

FIGURE 4.14: Calculated efficiency and reflection phase distribution for proposed reflectarrays.

Meanwhile, the spill-over efficiency equation η_{spill} is given by

$$\eta_{spill} = \frac{\int_0^{\theta_0} \int_0^{2\pi} G_{feed}(\theta, \varphi) \sin(\theta) d\theta d\varphi}{\int_0^{\pi} \int_0^{2\pi} G_{feed}(\theta, \varphi) \sin(\theta) d\theta d\varphi} \quad (4.14)$$

where $G_{feed}(\theta, \varphi)$ is the radiation pattern of the feed antenna, θ is the polar angle

and φ is the azimuthal angle. It is considered that a higher value of n for $\cos^n$ -like radiation pattern of the feeder will increase the taper and spillover efficiency. Meanwhile, the total aperture efficiency contains the product of two efficiencies ($\eta_{total} = \eta_{taper} \cdot \eta_{spillover}$), resulting in an increase of total efficiency. In addition, the radiation pattern of the feed $G_{feed}(\theta, \varphi)$ can be expressed by

$$G_{feed}(\theta, \varphi) = \begin{cases} G_0 \cos^n(\theta), & \text{if } 0 \leq \theta \leq \pi/2 \\ 0, & \text{if } \pi/2 < \theta \leq \pi \end{cases} \quad (4.15)$$

The feed antenna is a linearly polarized pyramidal horn with a gain of 12.4 dB at 7 GHz. The radiation pattern of the feed antenna is tapered by 10 dB at $\pm 30^\circ$ at a frequency of 7 GHz, which corresponds to the tapered illumination level from the center to the edge of the reflecting plane. The aperture size of the horn antenna is 84 mm $\times$ 64 mm. Meanwhile, the feed horn is placed in line with the center of the reflectarray. The optimal subtended angle θ for the feed antenna and the reflecting plane is 30° . As shown in Figure 4.14, a tradeoff is made between the spillover efficiency and taper efficiency to obtain the maximum value of their product. The optimal focal length to diameter ratio (f/D) is chosen as 0.76, where f is the focal length and D is the diameter of the reflecting plane. In an actual case, aperture efficiency is lower than this optimal value because of the phase error, feed blockage, and loss from other sources.

TABLE 4.5: Estimated loss budgets for the reference reflectarray and low-RCS reflectarray.

Type of loss	Conventional reflectarray	Proposed low-RCS reflectarray
Spillover loss	0.55 dB	0.55 dB
Taper loss	0.55 dB	0.55 dB
Dielectric loss	0.23 dB	0.23 dB
Phase error loss	0.41 dB	0.51 dB
Feed blockage loss	0.41 dB	0.41 dB
Max. aperture directivity	26.5 dBi	26.5 dBi
Calculated directivity	24.3 dBi	24.2 dBi

To predict the performance of two reflectarrays, the loss factors that contribute to the overall aperture efficiency are estimated. Because the feed antenna is in line with the center of the reflector, the feed blockage causes a gain reduction of 0.41 dB. The f/D ratio of 0.76 is chosen as a tradeoff between the taper efficiency

and spillover efficiency, leading to a reduced gain by 1.1 dB. The dielectric loss of substrate has reduced the reflected power by 0.23 dB for the conventional and proposed low-RCS reflectarray. A detailed summary of calculated constituents of antenna efficiencies and loss factors are listed in Table 4.5. The calculated directivity is larger than the simulated directivity because other loss factors exist, which are not taken into account. In summary, the total calculated antenna loss is 2.15 dB and 2.25 dB, for conventional reflectarray and proposed low-RCS reflectarray, respectively. The errors of the reflection phase are also investigated by perturbing the patch size of reflectarray elements. It is found that the gain fluctuation of our proposed low RCS reflectarray is less than 1.5 dB within its operational bandwidth.

Both reflectarrays have an equal aperture diameter of 288 mm, which consists of 148 square-patch elements. The conventional one is designed with the size variable patches on a grounded substrate. The other one, shown in Figure 4.15, is our proposed low-RCS reflectarray. Both reflectarray elements have a period of 18 mm and are equal to 0.42λ to avoid grating lobes, where λ is the free-space wavelength at 7 GHz. Both the conventional reflectarray and the phase shifting layer of our proposed low-RCS reflectarray are fabricated on RT5880 substrate with a thickness of 1.575 mm and dielectric constant of 2.2. For low-RCS reflectarray, the top layer consists of square-patches of variable sizes to control the reflection phase. AFSR structure is cascaded below the phase shifting layer, which consists of 3-D bandstop FSS, commercial absorber FGM-125 and PEC ground plane. It may be noted that the physical size of 3-D bandstop FSS layer is larger than the size of the phase shifting layer. The additional bandstop unit cells at the edge of a periodic array are employed to provide an infinite array environment setup for the inner unit cells.

4.4.3 Experimental Verification

In order to verify the design concept of phase controllable absorptive frequency-selective reflector, both the conventional reflectarray and our proposed low-RCS reflectarray are fabricated, as shown in Figure 4.16. Two reflecting surfaces are mounted on an aluminum plate for the convenience of measurement. The feeding antenna is placed above the reflecting surface with an appropriate distance to ensure 10 dB illumination amplitude taper from the center to the edge of reflectarray elements. The actual distance between the horn aperture and the reflecting surfaces

FIGURE 4.15: The prototype of proposed low-RCS reflectarray.

is 220 mm and the diameter of the reflecting surface is 288 mm. To eliminate the deviation of the focus position, the feeding horn antenna is tightly fixed with four supporting arms made of fiber glass, ensuring a sufficient mechanical strength. Meanwhile, air holes are drilled at the supporting arms to minimize the effect of blockage. The dielectric substrates are fabricated through the printed circuit board technology. For the reference reflectarray, the designed pattern is etched on a single-layer substrate. For our proposed low-RCS reflectarray, the phase shifting layer, 3-D bandstop FSS and absorbing layer are fabricated separately and then bonded together with screws and glues. For the assembly of 3-D bandstop FSS, the egg-crate structure is constructed by printing the designed resonator patterns on

the substrates and then interlocking the horizontal and vertical substrates together with pre-designed notches.

FIGURE 4.16: Photographs of fabricated conventional reflectarray and low-RCS reflectarray.

To validate the design of conventional and proposed low-RCS reflectarrays, their radiation performance is verified through simulation and experiments. The simulated maximum directivity for the conventional reflectarray and proposed low-RCS reflectarray is 23.8 dBi and 23.5 dBi, which are corresponding to aperture efficiencies of 53.7% and 50.1%, respectively.

Furthermore, the radiation patterns and realized gain of two reflectarrays are measured in an anechoic chamber. The comparisons between simulated and measured radiation patterns of the reference reflectarray and proposed low-RCS reflectarray in the E-plane and H-plane are shown in Figure 4.17. It is seen that the maximum measured realized gain are 23.6 dB and 23.3 dB for the conventional reflectarray and

(a) E-plane radiation pattern at 7.1 GHz

(b) H-plane radiation pattern at 7.1 GHz

(c) Gain versus frequency for conventional and low-RCS reflectarray

FIGURE 4.17: Radiation patterns and gain versus frequency of conventional and low-RCS reflectarrays.

our proposed low-RCS reflectarray, corresponding to antenna efficiencies of 51.2% and 47.8%, respectively. The simulated and measured side-lobe levels (SLLs) are generally above the expected level of 15 dB for both reflectarrays under test. It is seen that the measured radiation patterns are slightly unsymmetrical because the feeding cable of the horn antenna is placed in front of the reflectarray during measurements. The simulated and measured results of antenna gain versus frequency for the conventional reflectarray and our proposed low-RCS reflectarray are also shown in Figure 4.17. The measured in-band gain variation within less than 1 dB has bandwidth of 6.9% and 5.6% for the conventional reflectarray and proposed low-RCS reflectarray, respectively. The gain variations are mainly attributed to the phase errors of resonant elements with a very narrow bandwidth for beam collimation. The difference between the simulated and measured results in the antenna gain is due to some errors in the process of fabrication and measurement.

TABLE 4.6: Comparisons between conventional reflectarray and proposed low-RCS reflectarray.

Antenna Parameters	Conventional reflectarray	Proposed low-RCS reflectarray
Aperture diameter	288 mm	288 mm
Center frequency	7.1 GHz	7.1 GHz
Simulated directivity	23.8 dBi	23.5 dBi
Measured gain	23.6 dB	23.3 dB
Aperture efficiency	53.7%	50.1%
1 dB gain bandwidth	6.9%	5.6%
Lower RCS reduction bandwidth	for reference	76.9% (10 dB)
Upper RCS reduction bandwidth	for reference	17.1% (8 dB)

The RCS reduction performance is evaluated by comparing our proposed low-RCS reflectarray with the conventional reflectarray. The simulated bi-static scattering of both antennas is plotted in Figure 4.18 at three different frequencies for both TE polarization in the xoz plane and TM polarization in the $yo z$ plane. As seen from the upper half plane of the scattering patterns for both TE and TM polarizations, the magnitude of backscattering for our proposed low-RCS reflectarray is smaller than the conventional reflectarray. The angular range of RCS reduction is from -15° to 15° at the out-of-band frequencies of reflectarrays. Moreover, the bi-static RCS reduction performance works for both polarizations, which is mainly due to the symmetry of the structure. For the lower half plane of scattering patterns, it

(a) Scattering patterns at 3 GHz

(b) Scattering patterns at 4 GHz

(c) Scattering patterns at 9 GHz

FIGURE 4.18: Simulated bi-static scattering patterns for TE polarized wave in xoz plane and TM polarized waves in $yo z$ plane.

FIGURE 4.19: Simulated and measured monostatic RCS for conventional and low-RCS center-feed reflectarrays.

is seen that the magnitude of scattering is almost the same for both conventional reflectarray and proposed low-RCS reflectarray, due to the existence of the same PEC backplane. It is also observed that, for the backward scattering, our proposed low-RCS reflectarray achieves approximately 10 dB scattering reduction at 3 GHz and 4 GHz. In addition, it is also seen that more than 8 dB reduction of backscattering is achieved at 9 GHz for TE and TM polarized waves.

The monostatic RCS results of our proposed low-RCS reflectarray and the conventional reflectarray are also investigated by measurement in an anechoic chamber. It is noted that both feed antennas of reflectarrays are terminated with matched resistive loads during the measurement. Both the transmitting and receiving horn antennas have an electric field polarization in the y -direction. As shown in Figure 4.19, for the lower band, the measured monostatic RCS has more than 10 dB reduction from 2 to 4.5 GHz, corresponding to a fractional bandwidth of 76.9%. For the upper band, the measured RCS has more than 8 dB reduction from 9.1 to 10.8 GHz, corresponding to 17.1% fractional bandwidth. At out-of-band frequencies, most of the scattering is contributed by the reflecting surface and scattering from the feed can be neglected. At in-band frequencies, the monostatic RCS of reflectarray is inherently low. This phenomenon can be explained with the aid of the reciprocal theorem. During the measurement of RCS for reflectarrays, the incident plane wave is focused back to the feed antenna after illuminated on the reflecting plane. While most of the focused power will be received by the feed and then absorbed through a lossy termination. Therefore, it is observed that the scattering

level at in-band frequencies is significantly lower than other frequencies. The measured results of two antennas are summarized and listed in Table 4.6. It has shown that our proposed low-RCS reflectarray has achieved both highly directive beam and RCS reduction at the same time without sacrificing its radiation efficiency.

4.5 Low-RCS Reflectarray with Offset Feed

A low-RCS reflectarray consists of a periodic array of phase controllable AFSR unit cells is able to produce a collimated beam at in-band frequency and maintain a low scattering profile at the out-of-band frequency. The operating principles of our proposed center-feed low-RCS reflectarray was discussed in the previous section. Based on the assumption that the elements of reflectarray are operating in farfield range of the feed antenna, the orientation of horn antenna can be either a center-feed or an offset-feed one. In this case, the wave incident on the reflectarray element with normal or oblique incidence can be considered as a planar wave locally. Therefore, the designed phase controllable AFSR unit cell can also be utilized for RCS reduction of an offset-feed reflectarray. Compared with a center-feed reflectarray, an offset-feed reflectarray has the advantage of less blockage from the feed and direction of beam can be tilted to the desired direction for various applications.

4.5.1 Reflection Phase Analysis

The radiated wave of a feed horn antenna is considered as a spherical wave. In order to convert the spherical wave into a planar wave, the reflected phase of each reflectarray element is precisely calculated for compensation. The phase delay is proportional to the distance between the reflectarray element and the phase center of feed horn antenna. By individually adjusting the reflection phase of each elements, the phase distribution of the reflected field is compensated, which generates a pencil beam pointing at the given direction in the farfield. To describe the relative spatial relationship between feeding horn, reflection plane and beam direction, a Cartesian coordinate is set in Figure 4.20. The origin of the coordinates is set at the center of the reflecting plane. The feed horn antenna is located at $(0, -D, H)$, which is tilted from the normal direction. $\vec{r}_f$ is the position vector

pointing from the axis of horn antenna to the center of reflecting plane. The direction of the main beam is pointing at $\vec{u}_b$, which is normal to the reflecting surface in our design. Meanwhile, the elements of reflectarray are distributed on the xoy plane, with a position vector $\vec{r}_n$ pointing from the center of reflecting plane to the n th element of reflectarray. The reflection phase φ_n for each element is calculated as

$$\varphi_n = k \left(|\vec{r}_n - \vec{r}_f| - \vec{r}_n \cdot \hat{u}_b \right) + \Delta\varphi_0 + 2n\pi \quad (4.16)$$

where k is the wave number in free space and $\Delta\varphi_0$ is the constant reference phase. n is the integer to keep φ_n within the range of 360° .

(a) Conventional offset-feed reflectarray

(b) Low-RCS offset-feed reflectarray

FIGURE 4.20: Illustrations and coordinates of conventional and low-RCS offset-feed reflectarray antennas.

The offset-feed reflectarray consists of 148 elements arranged in a circular shaped pattern. To minimize the feed blockage, the feeding horn is placed at edge of reflecting plane with a distance of $D = 108$ mm from the projection of the horn aperture to the center of the reflecting plane. The height H between the phase center of the horn antenna and the reflecting plane is 165 mm. It is mentioned that H is an important parameter, which will affect the radiation performance of reflectarray. If H is a small value, the aperture efficiency will decrease due to the increased tilted angle of the feeding antenna. If H is too large, the side lobe of the antenna will also increase. Therefore, considering the trade off between side lobe level and aperture efficiency, an optimized value is chosen. The titled angle for the feeding horn antenna is around 33° . The period of reflectarray unit cell is 18 mm,

which is 0.43λ in terms of the operating frequency. The reflection phase for each reflecting elements based on equation (4.16) is calculated in Figure 4.21. In this case, the reflecting plane is a circular one, which is divided into discrete reflection phase distribution with a staircase approximation. Since the reflection phase range obtained by conventional and low-RCS reflectarray element is less than 360° , the implemented reflection phase exceed this range is represented by its closest value.

FIGURE 4.21: Reflection phase distribution of offset-feed reflectarray antenna.

The required reflection phase shift is obtained by varying the length of square patch for each reflectarray element. Since each patch is at its resonance, the reflection phase and patch length are not in a linear relationship. However, the correlations between patch length and reflection phase are summarized in Figure 4.12. For the conventional reflectarray, it is a single layer structure with square patch placed above the ground plane. The unit cell of low-RCS reflectarray element consists of two layers. The top layer is the phase shifting layer, which is made up of square patches with varying sizes. The bottom layer is the AFSR structure, which is placed below the phase shifting layer to provide a reflection band within the absorption band. The detailed implementation of the AFSR unit cell is based on the design presented in Section 4.4. It is noted that the patch can still be resonant at the reflection band of AFSR structure. The reflection phase control is obtained by varying the resonant dimensions of the AFSR-backed patches. Due to the narrow

bandwidth of resonant patches and AFSR structures, the reflection phase range is less than 360° and the phase variation versus the patch length is non-linear.

4.5.2 Experimental Verification

Two offset-feed reflectarray antennas are designed using square-patches of varying sizes. The first reflectarray is a conventional one with a normal grounded dielectric substrate. The second reflectarray is of low RCS, which utilizes size-varying patches loaded above an absorptive frequency-selective reflection structure (AFSR). Both reflectarrays are designed with the same aperture size and same feeding techniques. The substrate material for two reflectarrays is RT5880 with a thickness of 1.575 mm. Meanwhile, the operating frequency of two reflectarrays is designed at 7.2 GHz. To validate the concept of design, two reflectarrays are fabricated and assembled, which are shown in Figure 4.22. The conventional reflecting surface and low RCS reflecting surface are mounted on an aluminum platform for the ease of measurement. The feeding horn antenna is fixed on a teflon stand and the tilted angle of horn antenna is 33° relative to the normal direction of reflecting plane.

(a) Reference offset-feed reflectarray

(b) Fabricated low-RCS reflectarray

FIGURE 4.22: Fabricated prototypes of the reference and proposed low-RCS reflectarray with an offset feed.

To demonstrate the radiation and scattering performance of the proposed conventional offset reflectarray and low-RCS offset reflectarray, both simulations using

(a) E-plane pattern at 7.2 GHz

(b) H-plane pattern at 7.2 GHz

FIGURE 4.23: Simulated and measured radiation patterns of offset-feed conventional reflectarray and low-RCS reflectarray antennas.

commercial software and experiments in a microwave anechoic chamber are performed. The comparison between simulated and measured antenna radiation patterns in E-plane and H-plane are shown in Figure 4.23. The directivity of radiation pattern at 7.2 GHz is 23.1 dBi and 22.9 dBi for conventional offset-feed reflectarray and low-RCS offset-feed one, corresponding to an aperture efficiency of 45.7% and 43.6%, respectively. The measured realized gain is 22.9 dB and 22.7 dB at 7.2 GHz for conventional offset one and low-RCS offset reflectarray, respectively. The simulated side lobe levels (SLLs) of radiation pattern are above 15 dB, which are

15.9 dB for conventional offset-feed one and 17.8 dB for low-RCS offset-feed reflectarray, respectively. The H-plane radiation pattern is symmetrical and achieves a side lobe level of more than 20 dB. The simulated and measured gain versus frequency is plotted in Figure 4.24. It is shown that the measured antenna gain within 1 dB variation has a bandwidth of 4.1% and 4.5% for conventional offset one and low-RCS offset reflectarray, respectively.

FIGURE 4.24: Simulated and measured gain versus frequency for offset-feed conventional reflectarray and low-RCS reflectarray antennas

FIGURE 4.25: Simulated and measured monostatic RCS for conventional and low-RCS center-feed reflectarrays.

To validate the scattering performance of conventional offset reflectarray and low-RCS reflectarray, the monostatic RCS value of these two antennas are simulated

and measured, as shown in Figure 4.25. Since the conventional offset reflectarray consists of size-varying patches backed with a metallic ground plane, at the out-of-band frequency of conventional reflectarray, the incident plane wave will be all scattered back, leading to an increased level of RCS. While for our proposed low-RCS offset one, the incident electromagnetic wave can partially pass through the phase shifting layer and then transmit through 3-D bandstop FSS at its passband. Finally, the incident wave is substantially absorbed by the commercial absorber at the back, which reduces the monostatic RCS significantly compared with the conventional offset reflectarray. The simulated results of monostatic RCS have shown that 73.2% and 26.6% fractional bandwidth of monostatic RCS reduction of 10 dB and 8 dB for lower and upper bands, respectively. The measured monostatic RCS has shown a good agreement with the simulated one, which has verified the design of our proposed low-RCS reflectarray. The comparisons between the conventional offset-feed reflectarray and our low-RCS offset-feed reflectarray are summarized in Table 6.3.

TABLE 4.7: Comparisons between the conventional and our proposed low-RCS offset-feed reflectarrays.

Antenna Parameters	Conventional offset-feed reflectarray	Proposed low-RCS offset-feed reflectarray
Aperture diameter	288 mm	288 mm
Center frequency	7.2 GHz	7.2 GHz
Simulated directivity	23.1 dBi	22.9 dBi
Measured gain	22.9 dB	22.7 dB
Aperture efficiency	45.7%	43.6%
1 dB gain bandwidth	4.1%	4.5%
Lower RCS reduction bandwidth	for reference	73.2% (10 dB)
Upper RCS reduction bandwidth	for reference	26.6% (8 dB)

4.6 Summary

A phase controllable absorptive frequency-selective structure has been proposed in this chapter to manipulate the reflection phase of 3-D bandstop FSS backed with an absorbing material. The design concept is demonstrated by designing a curved and a planar reflective focusing structure with low-RCS. The frequency response of

absorption-reflection-absorption is obtained by cascading 3-D bandstop FSS with commercial absorbers. The reflection phase is controlled by varying the height of AFSR structure or introducing an additional layer of size-varying patches. The performance of conventional and low-RCS parabolic reflectarrays are compared. Two reflectarrays with center-feed are fabricated: a conventional reflectarray and our proposed low-RCS reflectarray. The radiation pattern and realized gain of each reflectarray have been demonstrated by the measurement results. In summary, the directivity, bandwidth and aperture efficiency of our proposed low-RCS reflectarray is comparable to the conventional reflectarray. Meanwhile, its RCS has achieved significant reduction at lower and upper bands. An explanation for the lower monostatic RCS at in-band frequencies is offered. Measured and simulated results are in good agreement, which verifies the design concept. In addition, the application range of our proposed phase controllable AFSR structure is expanded with the demonstration of a low-RCS offset-feed reflectarray antenna. Two samples of offset-feed reflectarray with conventional design and low-RCS design are fabricated and then tested.

Chapter 5

Low-RCS Phased-Array Antenna

5.1 Introduction

In Chapter 4, we proposed an approach to manipulate the reflection phase of an absorptive frequency-selective reflection structure (AFSR) to reduce the RCS of reflector antennas, including parabolic reflector antenna, center-feed reflectarrays, and offset-feed reflectarray. Nevertheless, the maximum aperture efficiency of reflector antennas is restricted due to the external feed and the total profile of the antenna is increased as well. Therefore, we hope to consider co-locating AFSR structure with the elements of phased-array antennas to obtain a beam steerable antenna with high gain and low scattering in this chapter.

Phased array antennas are essential in radar systems for object detection and target tracking [161]. In other applications, for instance, software defined radio [162] and multi-functional platforms [163], steerable and reconfigurable patterns are of particular interest. In operation, a phased array antenna is capable of electronically steering a highly directive radiation beam. Meanwhile, it also inevitably appears with a remarkable level of scattering [1], especially at its out-of-band frequencies. Considering the requirement of a stealth platform, it is also highly desirable that a phased-array antenna also features a low RCS. The major challenge is to preserve or to enhance the in-band radiating performance of antenna array, while reducing the array's RCS at out-of-band frequencies. On the other hand, frequency selective-surfaces (FSSs) are traditionally co-designed with antennas as radome [164] or ground plane [154] to realize a scattering reduction of these radiators. However,

utilizing FSS as either an antenna's superstrate or substrate is only effective for monostatic RCS reduction. Recently, the phase [32, 37, 165] and polarization [73, 166, 167] of the scattered waves can be arbitrarily manipulated by properly designed metasurfaces to achieve an optimal redistribution of scattered power in the space domain through algorithm-based optimization methods [168]. Nevertheless, these randomly distributed scattering elements are not capable of enhancing the radiation of an antenna.

Since the directivity of an antenna and its RCS are closely linked together through the effective radiating aperture. It is, therefore, a challenge to design a large radiating aperture with high aperture efficiency, while maintaining a low RCS. One solution is to modify a conventional Fabry-Perot (F-P) cavity antenna by coating a resistive layer [74] or an optimized scattering diffusive surface [169]. However, the bandwidth of F-P cavity antenna is inherently narrow due to its resonance mechanism and the radiated beam is not able to scan within a wide angle. Other feasible attempts include utilizing polarization conversion metamaterials (PCMs) as the ground plane of a tightly coupled dipole antenna array (TCDA) [170]. However, the reported array exhibits RCS reduction only in the upper frequency band and the reduction level is sensitive to the polarization and incident angle of the incoming wave. Moreover, other techniques such as randomly rotated elements [171], self-complementary grid [147], and dielectric superstrate [172] are only effective for the in-band RCS reduction of antenna arrays. To the best of our knowledge, an active phased array capable of beam scanning and wideband RCS reduction at both lower and upper bands have not been reported in the literature.

In this chapter, an absorptive frequency-selective structure with a frequency response of absorption-reflection-absorption is co-located with a dipole array. The elements of dipole array and AFSR structure are collaboratively sharing the same unit cell, while the reflection band of AFSR structure covers the operating band of the dipole antenna, providing an enhanced radiation in the forward direction. Meanwhile, the structural scattering of the array is significantly reduced due to the backed wide-band absorbers. The design details and simulated results of AFSR-backed dipole array are summarized. The fabrication and experimental results of our low-RCS phased array are also presented.

5.2 Implementation Details

The AFSR structure based on commercial absorber and bandstop FSS is a convenient way to achieve the frequency response of absorption-reflection-absorption. Some techniques are employed to enhance the absorption performance as well as to expand the reflection bandwidth. This enhanced AFSR structure can be utilized for the RCS reduction of a phased-array antenna.

5.2.1 AFSR Structure with Commercial Absorber

By integrating the enhanced absorber with a bandstop FSS, the frequency response of the AFSR structure is obtained and shown in Figure 5.3. For an incident wave with E-field oriented in the x -direction, it is observed that the reflection loss is less than 1 dB with 20% reflection bandwidth, while more than 90% absorption rate is achieved at its lower and upper bands, corresponding to two separate absorption bands with 83% and 40% bandwidth. Meanwhile, for an incident wave with E-field oriented in the y -direction, more than 10 dB reflection reduction has achieved over 160% the fractional bandwidth.

FIGURE 5.1: Frequency responses of AFSR structure under normal incidence.

The original absorption performance for a commercial absorber, FGM-125 from Eccosorb, is at least 10 dB reflection reduction from 2 to 10 GHz. However, the absorption bandwidth can be further expanded by cascading a few layers into a pyramidal shape to provide with better impedance matching and absorption.

Therefore, by utilizing a three-layer cascaded structure, the absorption bandwidth is extended to 160% with more than 90% absorption rate, compared with the original one. To produce a reflection band within the absorption band, a second-order bandstop FSS is designed with two transmission zeros at 10.1 GHz and 10.3 GHz, respectively.

FIGURE 5.2: Frequency responses of AFSR structure under oblique incidence.

FIGURE 5.3: Frequency responses of enhanced wide-band absorber and 3-D bandstop FSS.

The performance of the AFSR structure under the oblique incidence is also studied in Figure 5.1 and Figure 5.2. It is observed that the frequency response of the AFSR structure is stable under the TM polarized incident waves, with incident H-field oriented along the y -direction, respectively. For the TE incident waves with incident E-field along the x -direction, the frequency response has a slight variation with the increasing angle of incidence, which is attributed to the change

of mutual coupling between the two resonators. For TE incident waves with E-field along the y -direction, the absorption performance is nearly unchanged with over 90% absorption rate from 2 to 18 GHz. This design illustrates that a flat reflection band within an ultrawide absorption band can be obtained by combining 3-D bandstop FSS and pyramidal absorbers.

5.2.2 Co-located Array Element and AFSR Structure

An integrated unit cell of a dipole antenna with AFSR structure is illustrated in Figure 5.4, which consists of a single-polarized printed dipole antenna, 3-D bandstop FSS, and a pyramidal absorber with three layers of sheet absorbers. As a benchmark, conventional dipole array backed with a perfect electric conductor (PEC) is also designed. The dimensions of the AFSR-backed and PEC-backed dipole array elements are given in Table 5.1.

FIGURE 5.4: Illustration of AFSR-backed dipole antenna array element and the reference PEC-backed dipole array element.

For the single-polarized 3-D bandstop FSS, two ‘I’ shaped short-circuit resonators are cascaded together in the vertical direction with two resonating frequencies f_{r1} and f_{r2} , under the incident plane wave of x -polarization, as indicated in Figure 5.4. The length of resonator L_f determines the fundamental resonant frequency and it is utilized here for producing the short-circuit resonance. The distance h_{f1} affects the coupling between these two resonators, which influences the reflection bandwidth. The advantage of using two cascaded resonators is to create a sharp transition from absorption band to reflection band with a high absorption selectivity. Moreover, these two separate reflection poles at f_{r1} and f_{r2} are merged together and generate a relatively flat reflection band with doubled bandwidth, compared with the case of a single resonator. In addition, a small gap h_{f2} is reserved between the bandstop FSS and absorbers. Although eliminating this gap may reduce the profile of low-RCS dipole array, it will introduce additional loss to the resonators because of the effect from the closely loaded absorbers. To obtain an enhanced forward radiation, the separation between the dipole element and 3-D bandstop FSS is about a quarter wavelength of the operating frequency. The two arms of a dipole element are connected by a two-wire transmission line tapered from a microstrip line, which is then connected to the bottom feedlines for the convenience of impedance matching.

TABLE 5.1: Physical dimensions of reference and low-RCS dipole array element

Parameters	Length (mm)	Parameters	Length (mm)
p	16	L_f	6.6
h_1	1	L_a	6.5
h_2	5.4	L_1	3.5
h_3	3.3	L_2	6
h_4	14	W_f	3.2
h_a	4	W_a	0.5
h_{f1}	6.5	h_s	4.5
h_{f2}	4.25	L_s	5.5
W_s	0.5		

With an appropriate selection of dimensions, this AFSR-backed array element has a simulated impedance bandwidth of 15% in an infinite array environment, with the input reflection coefficient less than -10 dB from 9.5 GHz to 10.5 GHz, as shown in Figure 5.5. Compared with PEC-backed dipole, the operating bandwidth is almost maintained. The radiation pattern of AFSR-backed array element has a stable gain of about 5.4 dBi, a stable beamwidth of 82° and a front-to-back ratio of about 17 dB, as shown in Figure 5.6. Therefore, the integration of the

FIGURE 5.5: Simulated $|S_{11}|$ and gain of AFSR-backed array element and the reference PEC-backed array element.

FIGURE 5.6: Simulated radiation patterns of AFSR-backed array element and the reference PEC-backed array element.

AFSR structure with dipole antenna can enhance the antenna gain and suppress the backward radiation, compared with the standard dipole antenna with a bi-directional radiation pattern. The simulated directivity of the benchmark PEC-backed dipole element in an infinite array configuration is 5.42 dBi and the aperture efficiency for a single element is 97.5%. For the AFSR-backed dipole element, the realized gain is maintained from 9.5 GHz to 10.5 GHz, which corresponds to the reflection band of the AFSR structure. It is also noted that the realized gain of the AFSR-backed dipole element quickly drops at the absorption band of AFSR structure. From the simulated radiation patterns, it is also observed that the

radiation performance of our proposed AFSR-backed array can nearly be retained compared with the reference PEC-backed array.

FIGURE 5.7: Electric field distribution of low-RCS dipole element in radiating and scattering cases.

The effect of the AFSR structure working with dipole array element is illustrated in Figure 5.7. The single dipole element feeds through a vertical microstrip line and works as a driven element. Meanwhile, the AFSR structure performs as a parasitic element to provide an in-phase interference with the driven element. It is observed that the majority of power is transformed from a guided wave into a radiated wave. It is observed that the intensity of electric field is small at the absorber region, representing a low power dissipation at the radiation state of the antenna. The AFSR-backed dipole array element is absorptive for incident plane waves outside the reflection band of the 3-D FSS. It is noted that high intensity electric field exists in the absorber region, which represents substantial power loss with good absorption rate.

5.3 Active VSWR Performance

In the last section, we assume that the element of AFSR-backed dipole array is in an infinite array environment. Therefore, a unit cell model is created and simulated

for fast characterization of the array's radiation performance. For the simulation of a unit cell model, the four side walls of an array element are set with master-slave boundaries and the top wall is applied with absorbing boundary for radiated waves. The next step is to construct a finite array of 8×8 elements to test its radiation performance.

FIGURE 5.8: Simulated active VSWR of AFSR-backed dipole array at 10 GHz with 45° beam steering in the E-plane and H-plane.

To increase the simulation efficiency of large-sized array, a finite array domain decomposition method (DDM) tool box provided by HFSS is utilized in our design. Since the unit cells of array are exactly the same for all array elements, the generated mesh for single element is copied for all other elements, which reduces the requirement of time and mesh sizes for finite array analysis. The entire network data for all active ports are stored, which facilitates the analysis of mutual coupling between elements and obtain active input impedance. In Figure 5.8, a good active impedance matching at the central frequency of 10 GHz is achieved for our 8×8 array prototype at scanning conditions, with active VSWR < 3 for 45° beam scanning in the E-plane and H-plane, respectively. Based on the simulation results of 8×8 array using finite array domain decomposition tools, the excitations of all array elements can be edited with independent assignment of amplitude and phase. Therefore, a list of excitation phases is pre-calculated and then assigned based on the required beam steering angle. Steerable beams with a gain of 22.7 dBi at the broadside direction and 20 dBi at 45° scanning are obtained.

5.4 Experiment and Results

5.4.1 Fabrication and Measurement Setup

A prototype of 8×8 AFSR-backed dipole array is constructed and shown in Figure 5.9. The total array consists of four different parts of dipole array, 3-D bandstop FSS, wideband absorber and feeding network. The proposed AFSR-backed array has an active region of 163 cm^2 and it is constructed on a ground plane of size $16.8 \text{ cm} \times 16.8 \text{ cm}$. To avoid the edge termination effect, additional AFSR unit cells are added at the outermost layer. Meanwhile, the active radiation region of our proposed low scattering array is 163 cm^2 . The pyramidal shaped absorbers are made from a commercial absorber of FGMU-125, which is manufactured by Eccosorb. It has been customized into small pieces and then stacked into three staircases to improve the absorption bandwidth. Owing to the pyramidal shape and the usage of commercial absorber with both electric and magnetic losses, an ultrawide absorption band with 160% bandwidth and a lower overall profile of $0.06\lambda_L$ are obtained at the same time. The total profile of the absorber part is significantly reduced, compared with conventional approaches using circuit analog absorbers. It is noted that cylindrical holes and slots are cut in the absorbing layer and the metallic ground layer, which allow the vertical antenna substrates to be installed. Since these openings are relatively small in wavelength, they should not affect the absorption performance. The dipole array is etched on both sides of an RT5880 substrate with $\epsilon_r = 2.2$ and thickness of 0.787 mm, while the resonators of 3-D bandstop FSS are fabricated on the same material with a thickness of 1.57 mm. The AFSR structure and antenna array are connected together with pre-designed interlocking notches, which are easy to assemble and the structure is quite stable under external forces.

The measurements of radiation performance for our proposed array and the reference array are performed in a microwave anechoic chamber, as shown in Figure 5.10. The tested samples are fixed on a rotator, which is then driven by an electric motor with pre-installed measurement programs. The power levels of receiving signals are recorded and then transformed into the measured gain based on pre-calibrated results of a standard horn antenna (ETS-Lindgren's Model 3115). It is also well noted that the measurement of monostatic RCS of our proposed array is conducted by using two identical broadband horn antennas (LB-20245-SF) from AINFO for

transmitting and receiving signals. These two horn antennas are placed at the same location and then connected to a vector network analyzer (VNA) for signal post-processing. The in-built function of a time domain analysis from an Agilent E8362B VNA is utilized to eliminate the influence from the mutual coupling effects between horn antennas and to extract the echoes from the tested objects. The scattering cross sections of both fabricated samples of our proposed low-RCS phased array and the reference dipole array are measured and then compared.

FIGURE 5.9: Photograph of the fabricated 8x8 AFSR-backed dipole antenna array with detailed view of assembled parts.

(a) Schematic drawing of monostatic RCS measurement.

(b) Photograph of measurement in an anechoic chamber.

FIGURE 5.10: Photograph and measurement setup for monostatic RCS

5.4.2 Radiation Performance

The radiation performance of our proposed AFSR-backed array including reflection coefficients, radiation patterns and gain are simulated in a commercial software of CST Microwave Studio, and then measured in an anechoic chamber. Figure 5.11 illustrate the simulated and measured reflection coefficients of PEC-backed array and our proposed AFSR-backed array. The simulated impedance bandwidth is

20%, with $|S_{11}| < -10\text{dB}$ from 9 to 11 GHz for AFSR-backed array. The measured impedance bandwidth is about 22%, with $|S_{11}| < -10\text{dB}$ from 8.8 to 11 GHz.

(a) Measured $|S_{11}|$ of PEC-backed array (b) Measured $|S_{11}|$ of AFSR-backed array

FIGURE 5.11: Simulated and Measured $|S_{11}|$ of PEC-backed and AFSR-backed dipole antenna array.

A good agreement has been obtained between the simulated and measured results. It is noted that the impedance bandwidth is larger than the gain bandwidth for the AFSR-backed array. The possible reason is due to the loss from the feeding network and absorbers, which might reduce the Q factor and slightly increase the impedance bandwidth. However, the gain bandwidth truly reflects the operating frequency range of our proposed AFSR-backed array. The simulated and measured radiation patterns in both E-plane (y - z plane) and H-plane (x - z plane) of the proposed AFSR-backed dipole antenna array are shown in Figure 5.12, along with those for the reference PEC-backed dipole antenna array.

As shown in Figure 5.13, the simulated E-plane radiation pattern of AFSR-backed array is very stable over the operating frequency range. It has a stable measured gain of 21.5 dBi and beamwidth of 12.5° , which is almost identical to the reference PEC-backed array. The radiation pattern in the H-plane has also been verified theoretically and experimentally. It is confirmed that good radiation performance is achieved at 10 GHz for our proposed AFSR-backed array. Furthermore, the simulated gain of AFSR-backed array is found to be stable from 9.5 to 10.5 GHz, which is an encouraging result. More importantly, the gain of AFSR-backed array is maintained compared with the reference PEC-backed array. The simulated aperture efficiency of AFSR-backed array is more than 70% over the operating frequency range. It should be mentioned that some power is inevitably dissipated

FIGURE 5.12: Measured radiation patterns of PEC-backed and AFSR-backed array in E-plane and H-plane at 10 GHz.

FIGURE 5.13: Simulated aperture efficiency and measured gain over frequency of AFSR-backed dipole array and the reference PEC-backed dipole array.

due to the fact that the vertical feeding lines are penetrated through the absorbers, which slightly decreases the aperture efficiency of AFSR-backed array. The isolation between co-polarization and cross-polarization is over 20 dB over the operating frequency band.

FIGURE 5.14: Simulated scanned beams in E-plane and H-plane at 10 GHz

The beam-steering performance of our AFSR-backed array is highly attractive as shown in Figure 5.14. A two-dimensional scan has been performed in both E-plane and H-plane. The radiation patterns are able to continuously scan from 0° to 45° with a fixed step of 15° . More importantly, the scanned patterns are stable and no grating lobes are observed during the beam scanning over the entire operating frequency band.

5.4.3 RCS Performance

The monostatic RCS of our AFSR-backed dipole array and the reference PEC-backed dipole array is investigated. It should be mentioned that both two dipole arrays are terminated with $50\ \Omega$ matched loads under this situation. The designed polarization of the array is defined as the co-polarization, while its orthogonal one is defined as the cross-polarization.

As shown in Figure 5.15, the simulated 10 dB two-sided out-of-band RCS reduction for co-polarization has been achieved from 3 GHz to 8.6 GHz for the lower side and from 13.3 GHz to 18 GHz for the upper side, corresponding to fractional

(a) Monostatic RCS for co-polarized incident wave.

(b) Monostatic RCS for cross-polarized incident wave.

FIGURE 5.15: Measured monostatic RCS for AFSR-backed dipole array and the reference PEC-backed dipole array.

bandwidth of 96.5% and 30%, respectively. The reduced RCS at in-band frequency is attributed to the termination of the resistive load, which dissipates the signal received by the antenna. The in-band monostatic RCS of our proposed array is nearly unchanged, compared with the reference antenna array. The results of RCS reduction for the cross-polarization are encouraging, as shown in Figure 5.15. The AFSR-backed dipole array has 10 dB RCS reduction from 3 to 18 GHz, corresponding to a fractional bandwidth of 142%. It is confirmed both theoretically and experimentally that our proposed AFSR-backed array can achieve a significant RCS reduction for both co-polarized and cross-polarized incident plane waves.

FIGURE 5.16: Simulated bistatic RCS for AFSR-backed dipole array and the reference PEC-backed dipole array.

The bistatic RCS of the proposed AFSR-backed array is investigated at two sampled frequencies of 4 GHz and 14 GHz, for both co-polarization and cross-polarization. The results of bistatic RCS reduction results are attractive as well. It is observed in Figure 5.16(a) and (b) that 15.4 dB of backscattering reduction is obtained at 0° and 6 dB reduction is achieved at the $+60^\circ$ direction in the azimuth plane for co-polarized incident plane wave. Similar to the case of 4 GHz, a remarkable scattering reduction is observed from -60° to $+60^\circ$ at a higher frequency of 14 GHz for co-polarized incident wave. More importantly, due to the well-controlled scattering for the antenna mode, the scattering patterns of AFSR-backed array for co-polarized incident wave is much smaller than that of the reference PEC-backed

array. The scattering patterns for cross-polarized incident wave are dominated by the structure mode of the array. Owing to the absorption of AFSR structure, the scattering pattern of AFSR-backed array is smaller, compared with the PEC-backed array. As shown in Figure 5.16(c) and (d), the scattering cross section of AFSR-backed array for cross-polarized incident waves exhibits a noticeable reduction of 15 dB and 15.2 dB over the angular range from -60° to $+60^\circ$ at 4 GHz and 14 GHz, respectively. In summary, our proposed AFSR-backed dipole array exhibits a significant scattering cross section reduction compared to the reference PEC-backed array.

5.5 Summary

In this section, a beam-steerable dipole array has been integrated with absorptive frequency-selective reflection structure (AFSR) to obtain low scattering and enhanced radiation performance. The methodology of co-locating dipole antenna with AFSR structure has been presented. Two prototypes of AFSR-backed dipole array and reference PEC-backed array are simulated, fabricated and then measured. The radiation and scattering performances of AFSR-backed dipole array are verified both theoretically and experimentally. It has been demonstrated that 15% operating bandwidth with more than 70% aperture efficiency has been obtained by our proposed AFSR-backed array. The in-band RCS of the proposed array is almost unchanged. Meanwhile, a wideband out-of-band RCS reduction is obtained with fractional bandwidth of 96.5% and 30% for lower and upper absorption bands, when the incident wave is co-polarized. In addition, a significant RCS reduction has been achieved for cross-polarized incident wave with 142% fractional bandwidth, comparing with the reference array.

Chapter 6

Low-RCS Antenna Radomes

6.1 Introduction

Antenna radome is a protective enclosure covered above the surface of an antenna. An ideal radome should be designed without degrading the antenna's radiation performance. Therefore, a good radome is almost transparent to the incident waves at antenna's operating frequency. Meanwhile, the out-of-band incident waves can be reflected or absorbed to reduce the effect of spurious signals. Frequency-selective surfaces is a suitable candidate to achieve frequency filtering for antenna radome. However, it inevitably reflects all incident waves out of the operating bands, resulting in large scattering cross sections. For the consideration of RCS reduction of communication platform, the scattering of an antenna can possibly be reduced by using a low-RCS antenna radome. The concept of absorptive frequency-selective transmission structure (AFST) or frequency-selective rasorber (FSR) has been attracting a growing attention recently due to their unique frequency response of absorption-transmission-absorption. In particular, an AFST structure is able to provide a low loss transmission band for the antenna's proper operation, while absorbing all incident waves outside the transmission band, simultaneously.

The state-of-the-art performance of AFST is to create an absorption-transmission-absorption response. In this chapter, two designs of AFST structure are proposed.

One is based on hybrid resonators, which significantly reduce the number of required lumped resistors with our design. Another one is a new approach to designing an AFST structure based on commercial absorbers, which serves as a guideline and can be applied to any other AFST design readily.

6.2 Absorptive Frequency-Selective Transmission Structures

Apart from reflection and transmission responses, absorption within a certain range of frequency is also an important functionality of a surface or structure. For many existing designs of FSS, the reachable performance is limited by the losses inside the unit cell. A thin layer capable of absorbing all incident electromagnetic waves is demanded from microwave to optical frequencies. Historically, absorbers are utilized as non-reflective coating to cover an object for backscattering reduction. Therefore, most of classical absorbers are implemented by attaching a conducting plates at the back of a lossy material. Absorption of incident waves is substantially obtained by power dissipation within the lossy material after one or several reflections at the conducting backplane. However, the incident wave can not transmit through the absorbers due to the existence of the backplane.

FIGURE 6.1: Functionality of an AFST structure.

The aim of our design is to create a transmission band within a wide absorption band, which is characterized by a frequency response of absorption-transmission-absorption. The operating principle of an AFST structure illustrated in Figure 6.1,

incident waves at the lower frequency f_1 and the upper frequency f_3 are absorbed by the structure. Meanwhile, incident waves at the middle frequency f_2 are capable of passing through the structure with a very low insertion loss.

6.3 AFST Structure Based on Hybrid Resonators

6.3.1 Unit Cell Configuration

For a circuit analog absorber, it is characterized with a two-layer design of an absorption layer and a conducting backplane. The absorption of incident waves is contributed by the induced current flowing through the lumped resistor. Meanwhile, the losses within the lumped resistors can be bypassed or truncated by creating a short or open circuit resonance. Therefore, the absorption frequency can be controlled by the resonant frequency of resonators. Since a conducting backplane reflects the incident waves of all polarizations, slots are cut on the conducting plane to create a passband. The proposed unit cell of our AFST is illustrated in Figure 6.2. It consists of two layers of printed circuit boards (PCBs), which are separated by an air spacer. The first layer is intended for absorption of incident waves. It is seen that a square-loop hybrid resonator is connected with metallic strips. Lumped resistors are soldered between these metallic strips to provide the desired absorption at certain frequencies. The second layer is a bandpass FSS, which is implemented by cutting slots on a conducting plate.

TABLE 6.1: Physical dimensions and parameters of AFST unit cell.

Parameters	Values	Parameters	Length (mm)
p	24 mm	l_w	0.5 mm
l_1	7.3 mm	g_1	0.25 mm
l_2	9.8 mm	g_2	0.8 mm
l_3	8.88 mm	ε_r	3
l_4	1.52 mm	R_a	294 Ω
t	1 mm	h	11 mm

The first and second layers of the AFST structure are designed with a RO4230 substrate with a dielectric constant of 3 and thickness of 1 mm. Each unit cell of the AFST structure consists of 4 lumped resistors with a resistance value of 294 Ω (Part No.:ERA-2AEB2940X). The airspace between the first and second layers

(a) Unit cell of AFST

(b) First layer

(b) Second layer

FIGURE 6.2: Unit cell structure of AFST based on square loop hybrid resonators.

is 11 mm. The physical dimensions of the hybrid resonator-based AFST unit cell are summarized in Table 6.1. The simulated S-parameter results of AFST unit cell using ANSYS HFSS are shown in Figure 6.3. It is shown that a passband with 0.29 dB insertion loss at 6.1 GHz can be achieved. The bandwidth with $|S_{11}| < -10$ dB ranges from 2.8 GHz to 9.8 GHz, corresponding to a fractional bandwidth of 112.4%. In addition, the passband with -3 dB insertion loss is from 5.7 to 6.5 GHz, which represents a fractional bandwidth of 13%. A parametric scan of different resistor values is performed to examine the effect of resistor values on the reflection and transmission responses. It is observed that the resistance values have a minor effect on the transmission frequency. Meanwhile, the absorption performance is

related to the resistance value, which is selected with an optimized value of 294 Ω for a maximum absorption bandwidth.

FIGURE 6.3: Parametric scan results of different resistor values.

The absorption rate is defined as the energy absorbed by the lossy layer, which is calculated as

$$Absorption = 1 - |S_{11}|^2 - |S_{21}|^2 \quad (6.1)$$

Furthermore, the current distribution on the first layer at the absorption frequency and transmission frequency are plotted in Figure 6.4. At the absorption band of 4 GHz, current is induced at the metallic strips. It is clearly observed that the current density has reached a maximum at the location of the resistor, resulting in a substantial absorption.

(a) At absorption frequency of 4 GHz

(b) At transmission frequency of 6 GHz

FIGURE 6.4: Current distribution of square loop hybrid resonator.

At the passband frequency of 6 GHz, the induced current is mainly concentrated within the square loops region due to the existence of an open-circuit resonance. As a result, almost no current flows through the resistors, which leads to a low loss

in the resistors. Since the patterns of AFST structure are symmetrical, it is easy to find that the frequency response of unit cell is polarization insensitive under the normal incidence.

6.3.2 Equivalent Circuit model

The corresponding equivalent circuit model for the unit cell of the AFST structure is illustrated in Figure 6.5. The bandpass FSS located at the bottom layer is represented by a pair of inductor L_f and capacitor C_f connected in parallel. Meanwhile, the top layer is equivalent to a square-loop hybrid resonator in series with lumped resistors. To facilitate the analysis, the admittance of the air spacer and a bandpass FSS is represented by Y_{in} . If $Y_{in} = 0$, the input impedance of Port 1 is only determined by the impedance of lossy path. Meanwhile, the impedance of the lossy path is a resistive value and good matching can be achieved to the characteristic impedance of incident waves. To simplify this equation of absorption frequency, we may ignore the effects of the substrate due to its small electrical length. The lower and upper absorption bands at f_{ro} and f_{r2} can be analytically obtained by solving the equation of $Y_{in} = 0$, which is derived as

$$\tan(\theta_2) = Z_2 \left(\frac{1}{2\pi f L_f} - 2\pi f C_f \right) \quad (6.2)$$

where $\theta_2 = 2\pi f h/c$, c is the speed of light and h is the air gap between two substrates. Two analytical solutions of lower and upper absorption frequencies for this transcendental equation can yield approximate results of 4.1 GHz and 9.1 GHz, respectively. The analytical results agree with the full-wave simulation results, which verifies with the effectiveness and validating of the equivalent circuit model. The parameters of the equivalent circuit model are listed in Table 6.2.

To create a transmission band within the absorption band, the series and parallel resonances of a square loop hybrid resonator (SLHR) are utilized in this design. According to the equivalent circuit model in Figure 6.5, the parallel resonant frequency f_p is expressed as

$$f_p = \frac{1}{2\pi \sqrt{C_p(L_{p1} + L_{p2})}} \quad (6.3)$$

(a) Equivalent circuit model

(b) Variation of impedance versus frequency

(c) Full-wave simulation and equivalent circuit results

FIGURE 6.5: Equivalent circuit model and full-wave simulated results of AFST structure.

Meanwhile, a series resonance frequency is created by the SLHR resonator at f_s

$$f_s = \frac{1}{2\pi\sqrt{C_p L_{p2}}} \quad (6.4)$$

where the inductance values of L_{p1} and L_{p2} are related to the length of inner and outer loops in the top layer, respectively. The capacitance C_p is determined by the gap distance g_1 between the inner and outer loops. In addition, the bandpass FSS will provide a open circuit resonance at frequency f_b , which is expressed as

$$f_b = \frac{1}{2\pi\sqrt{C_f L_f}} \quad (6.5)$$

For the bandpass FSS, the square patch and cross-shaped frame have an equivalent inductance of L_f . The slot etched on the backplane has an equivalent capacitance of C_f . According to the transmission conditions derived in [82], both the top and bottom layers have to satisfy a parallel resonance at the same frequency of f_{r1} . Therefore, the condition of transmission is obtained by tuning the circuit parameters to achieve $f_p = f_b = f_{r1}$. In this case, the parallel resonance of double square loop resonator is utilized to create the transmission band. Meanwhile the short circuit resonance of double square loop resonator at frequency f_s is very close to the absorption resonance at f_r , which can expand the bandwidth of the upper absorption band.

TABLE 6.2: Equivalent circuit parameters of AFST unit cell.

Parameters	Values	Parameters	Length (mm)
Z_0	377 Ω	L_{p1}	2.6 nH
Z_1	217 Ω	L_{p2}	2.4 nH
Z_2	377 Ω	L_f	5 nH
L_s	4.1 nH	C_g	0.08 pF
θ_1	12° at 10 GHz	C_p	0.14 pF
θ_2	132° at 10 GHz	C_f	0.16 pF
R_a	294 Ω		

The operating principle of the AFST structure can be explained from another perspective. It is observed in Figure 6.5 that the equivalent input admittance Y_{in} is zero at the lower and upper absorption frequencies of f_{r0} and f_{r2} , respectively. In this case the characteristic impedance of the incident wave is matched with the impedance of the lossy path, resulting in a substantial absorption. Meanwhile, a parallel resonance is create by the double square loop resonator at frequency f_{r1} .

FIGURE 6.6: Simulated frequency responses of hybrid resonator based AFST structure under the normal and oblique incidence.

Under this situation, the lossy path is disconnected at the frequency of f_{r1} and the induced current is not able to flow through the resistors. Therefore, the incident waves can pass through the absorption layer with ideally no loss. In addition, the bandpass FSS is also at resonance at frequency f_{r1} , providing a transmission window for the incident wave to pass through.

6.3.3 Normal and Oblique Incidence Performance

The simulated reflection, transmission and absorption performance of the hybrid resonator-based AFST structure for normal and oblique incidences is presented in Figure 6.6. The frequency range of simulation is set from 1 GHz to 11 GHz. A passband response with 1 dB bandwidth of 6.3% centered at 6 GHz and an absorption band with 10 dB reflection reduction bandwidth of 135% are simulated. After changing the angle of incidence, the passband response remains in existence

FIGURE 6.7: Fabricated samples of AFST structure and measurement setup.

with a slight frequency shift for both TE and TM polarized incident waves, respectively. In the meantime, the performance of absorption is stable under the oblique incidence, which is mainly contributed by the lumped resistors with ohmic losses.

The simulated absorption rate of the hybrid resonator-based AFST structure under normal and oblique incidences with TE and TM polarized incident waves is shown in Figure 6.6. Two absorption bands are observed, which are located at the lower and upper bands of the transmission window. The simulated absorption rate is higher than 80% and the performance of absorption is stable under the oblique incidence for TE and TM polarized incident waves.

6.3.4 Experimental Verifications

In order to verify the presented design concept, a design example of the hybrid resonator based AFST structure is fabricated and tested. Figure 6.7 shows the photograph of the fabricated prototype, which consists of 1×6 unit cells with a size of $12 \times 72 \text{ mm}^2$. It is mentioned that the total number of lumped resistor required in each unit cell has reduced by four times, compared with the previous design in [82]. The fabricated AFST structure is then measured in a parallel plate

(a) Measured $|S_{11}|$ and $|S_{21}|$

(b) Measured absorption rate

FIGURE 6.8: Simulated and measured frequency response and absorption rate for the fabricated AFST under the normal incidence.

waveguide (PPW) test set-up. The design details of PPW measurement setup and procedures can be found in [10].

The fabricated AFST structure for measurement in a parallel plate waveguide is shown in Figure 6.7, which consists of two substrates with different etched patterns. The structure of PPW consists of two main parts. The top and bottom covers of the waveguide are conducting plates, while the two side walls of PPW are covered with microwave absorbers to reduce the reflection. The traveling TEM wave inside the PPW is launched by one conical disk and then received by another one, both are located at the two ends of the waveguide sections. For a clear view of the arrangement of AFST structure inside the waveguide, the upper plate of PPW is not assembled when taking the photograph. It is mentioned that the test frequency

range of this waveguide set-up is from 1.5 GHz to 12 GHz. The lowest operating frequency of PPW is determined by the dimensions of the cone disk, while the highest frequency is the cut-off frequency for TM_1 mode inside the parallel-plate waveguide. The test frequency range of PPW has sufficiently covered the operating frequency range of our AFST structure to be measured.

The simulated and measured transmission and reflection coefficients of the proposed hybrid resonator based AFST structure under the normal incidence are shown in Figure 6.8. It is clear that the full-wave simulation results show a good match with the measurement ones. The small difference between the simulated and measured results is attributed to the errors of fabrication and measurement. In addition, the measured results indicate a transmission window with 0.6 dB insertion loss at 6.3 GHz. The fractional bandwidth for this passband with less than 3 dB insertion loss is 21.3%. It is shown that the reflection coefficient is below -10 dB from 2.1 to 9.8 GHz, corresponding to a fractional bandwidth of 129.1%.

TABLE 6.3: Comparison of the existing 2-D AFST design in the literature.

Reference	Insertion loss (dB)	Lower absorption bandwidth ²	Upper absorption bandwidth ²	Thickness λ_L	Number of lumped elements ³
[99] ¹	0.33	N.A.	71.1%	0.080	N.A.
[103] ¹	0.30	N.A.	105.1%	0.103	8
[102] ¹	0.20	103.2%	N.A.	0.120	4
[173] ¹	0.50	92.8%	N.A.	0.112	4
[82]	0.36	25.1%	11.5%	0.113	18
[174]	0.40	61.1%	37.7%	0.124	16
[92]	0.36	41.3%	20.7%	0.110	8
This work	0.29	91.4%	31.9%	0.104	4

¹ These designs have only one-sided absorption band.

² The bandwidths are for 80% absorption rate

³ Number of lumped components required in a unit cell.

In summary, it is expected that the structure has more than 80% absorption rate for an incident wave from 1.9 to 5.1 GHz and from 7.1 to 9.8 GHz, which corresponds to fractional bandwidth of 91.4% and 31.9%, respectively. It is seen that the measured absorption bandwidth is slightly larger than the simulated one and the reason for this difference is possibly due to the parasitic effect of lumped resistors. Since the parasitic effect is not taken into the consideration of the full-wave simulations and an ideal model for resistor is applied. The comparisons between our design of

AFST and other designs in the literature are listed in Table I. As expected from the comparisons, our proposed AFST structure has an advantage of two-sided absorption bands and operates under dual-polarized incident wave. Meanwhile, it also has a reasonably low profile with the minimum requirement for the quantity of lumped resistors

6.4 FSS-Based AFST Structure

6.4.1 Absorption Performance

In the previous section, we have presented a design of an AFST structure based on hybrid resonators and lumped resistors to achieve the function of transmitting and absorbing incident waves. However, circuit analog absorber has its limitation for the total profile and maximum absorption bandwidth. To overcome this limitation, commercial absorbers with both electric and magnetic losses are utilized for the design of an improved AFST structure. Moreover, one of the main advantages of using commercial absorber is that it has a larger power handling capacity, compared with the design using lumped resistors. Therefore, it is straightforward to construct an AFST structure using commercial absorber, which may able to achieve better absorption performance by taking the advantage of commercial absorbers.

An absorber is characterized by its electric permittivity ε_r and magnetic permeability μ_r , which are both complex numbers with imaginary parts representing the attenuation of electric and magnetic field inside the absorber. In this case, a thin and broadband magnetic absorber FGM-125 fabricated by Eccosorb is chosen as the absorber for our design of AFST. This FGM-125 has an appearance similar to most of silicon rubber absorbers. It is intended for free space reflection reduction with effective frequency range from 2 GHz to 12 GHz. The material property of FGM-125 is listed in Figure 6.9. For a single layer of flat absorber backed with a perfect conducting plate, the reflection loss (RL) in dB is calculated by

$$RL = 20 \log \left| \frac{\sqrt{\mu_r/\varepsilon_r} \tanh(j\omega d \sqrt{\mu_r \varepsilon_r}/c) - 1}{\sqrt{\mu_r/\varepsilon_r} \tanh(j\omega d \sqrt{\mu_r \varepsilon_r}/c) + 1} \right| \quad (6.6)$$

where ϵ_r and μ_r are the relative permittivity and permeability of the absorber. c is the speed of light in vacuum and ω is the angular frequency. d represents the thickness of absorber.

FIGURE 6.9: Relative permittivity and permeability of FGM-125 absorber.

The reflectivity reduction of a single-layer FGM-125 absorber with a thickness of 3.3 mm is shown in Figure 6.10. The reflection coefficient is obtained by attaching a metallic plate at the back of absorber. Under the normal incidence, the bandwidth of 10 dB reflection reduction rate is 8.3 GHz. Two absorption peaks are observed at 4 GHz and 10 GHz, respectively. Following this, a strategy of greatly extend the absorption band of flat absorbers by cascading them into a multi-layered structure.

FIGURE 6.10: Simulated reflection coefficient of single-layer and three-layer cascaded absorbers.

The unit cell structure of the cascaded absorber is shown in Figure 6.10, which has three layers of square blocks with different sizes. The side length of square blocks is decreasing from the bottom to the top with dimensions of 8 mm, 6 mm and 5 mm, respectively. Meanwhile, each layer has a uniform thickness of 3.3 mm.

The absorption performance of the cascaded absorber is simulated with HFSS and the reflection coefficient is extracted from the simulation results. It is shown that the simulated reflection coefficient is less than -10 dB from 1.4 GHz to 15 GHz, corresponding to a fractional bandwidth of 165.8%. The extension of the absorption bandwidth is due to the integration of multiple resonances and edge diffraction effects, which are introduced after cascading multiple flat absorbers together.

6.4.2 Bandpass and Bandstop FSS

For absorbers backed with a conducting plate, the incident wave is either absorbed or reflected. In this case, transmission of incident wave is almost impossible due to the existence of an ideal back reflector. However, the shape and design of the back reflector can be modified to allow the incident wave to pass through within a certain frequency range.

FIGURE 6.11: Simulated frequency response of absorber with an open slot.

In particular, a traditional bandpass FSS is designed by cutting slots on a conducting plate. The resonance of slots will generate a pass band for incident waves at a certain frequency and reject waves at other frequencies. Based on the same principle, a transmission band can also be created by cutting slots on the back plate of an absorber. Compared with a lossless bandpass FSS, it is expected that the insertion loss of transmitted wave is increased, which is due to the fact that a portion of incident waves is inevitably dissipated within the absorbers. To prevent the dissipation of incident wave at the passband, the sidewalls of absorbers are coated with a conductor layer.

The simulated transmission and reflection coefficients of a three-layer cascaded absorber with an open slot are shown in Figure 6.12. It is observed that a pass band has been introduced within the absorption band. The insertion loss of the transmission band is around -4 dB and the reflection coefficient is -6 dB at the passband of 10 GHz. From the current result, the pass band is created by a single-resonance of slot, which can be expanded by introducing higher order resonances.

FIGURE 6.12: Simulated transmission and reflection coefficients of bandpass FSS.

Here, we propose an improved design of bandpass FSS with two transmission poles and extended transmission bandwidth. Compared with the previous design, a via hole is inserted in the middle of PEC-walls, which generates a second-order resonance for the incident electric field polarized in the y -direction, which generates a relative flat pass band at 10 GHz.

FIGURE 6.13: Simulated and measured frequency response and absorption rate for the fabricated AFST under the normal incidence.

The simulated transmission and reflection responses are shown in Figure 6.12. From the simulation results, it is seen that the insertion loss is less than 1 dB from 9.5 to 10.5 GHz, corresponding to 10% fractional bandwidth. Furthermore, we have combined the design of second-order bandpass FSS with absorbers. It is observed in Figure 6.13 that the insertion loss is around 5 dB with two transmission poles observed at 9.4 GHz and 10.3 GHz. However, the insertion loss is high due to the fact that a portion of incident wave is dissipated within the absorbers. The power loss at the passband can be reduced by designing a bandstop FSS to prevent the wave dissipations inside the absorbers.

6.4.3 Integrated AFST Structure

The structure of our AFST unit cell is shown in Figure 6.14 and it comprises a bandpass FSS, a bandstop FSS and a pyramidal shaped absorbers. The bottom base of the AFST structure is a stack of pyramidal absorber, consisting of three layers of FGM-125 absorbers with different sizes. These rubber-based absorbers are manufactured by Laird Performance Materials with a thickness of 3.3 mm and it is convenient to cut them into customized pieces. For the consideration of absorbing performance, these absorbers are backed with conducting plates.

FIGURE 6.14: Simulated performance of AFST structure.

The original absorption performance of this absorber has 10 dB reflection reduction from 2.5 GHz to 12 GHz, which can be referred to from its data-sheet [175]. The absorbers are stacked with three layers of different sizes to utilize the effect of multiple resonances and edge diffractions for the extension of absorption bandwidth.

FIGURE 6.15: Simulated and measured frequency response and absorption rate for the fabricated AFST under the normal incidence.

TABLE 6.4: Physical dimensions of Commercial absorber based AFST unit cell

Parameters	Length (mm)	Parameters	Length (mm)
p	20	L_s	5
w_1	3.6	L_1	8
w_2	3.5	L_2	6
w_3	0.5	L_3	4
w_4	1	h_a	3.3
h	14.5	D	1
t	0.5		

Since the structure is a single-polarized one, we only consider the incident electric field polarized in the x -direction. In addition, it is shown that the transmission efficiency will be significantly enhanced after combining the bandpass FSS with bandstop FSS. A transmission band is produced by cutting a period array of patterned rectangular slots on the conducting backplates. Similar to a bandpass FSS, it is a natural way of generating a pass band. However, the slot apertures are covered with absorbers, so the transmission efficiency is inherently limited by the losses in absorbers. In order to minimize the losses of incident waves when it passes through the absorber, four side walls of slots are covered with conducting plates. In addition, the bandwidth of the pass band is expanded by inserting a metallic pin between the two metallic sidewalls, as shown in Figure 6.12. Moreover, the efficiency of transmission is further enhanced by designing two ‘I’ shaped short-circuit resonators to prevent wave dissipations inside the absorbers. The substrate material of the bandstop FSS has a permittivity of 2.2 and a thickness of 0.5 mm, which is assembled parallel to the front face of the bandpass FSS. The physical dimensions of the AFST unit cell are summarized in Table 6.4.

The simulated reflection and transmission coefficients of proposed commercial absorber based AFST unit cell under the normal incidence are shown in Figure 6.14. At the out-of-band frequencies, the reflection coefficients of our AFST structure is small due to the substantial absorption of the incoming waves. It is observed that AFST structure exhibits more than 10 dB reflection reduction from 1.6 GHz to 15 GHz, which corresponds to a fractional bandwidth of 161.4%. In the meantime, the in-band transmission efficiency is significantly enhanced with less than 0.8 dB from 9.5 GHz to 10.8 GHz, which corresponds to a fractional bandwidth of 12.8%.

FIGURE 6.16: Simulated electric field distribution of commercial absorber based AFST structure at different frequencies.

The simulated electric field distributions inside AFST structure for incident waves of three different frequencies are plotted in Figure 6.16. The incident wave enters from the left side of the AFST and exits from its right side. It is observed that at the pass band frequency of 10 GHz, the incident waves are blocked outside of the absorbers due the resonance of bandstop FSS. Therefore, the transmission efficiency is greatly enhanced by tunneling through the bandpass FSS. Meanwhile, at the lower and upper absorption bands, the incident waves are substantially absorbed by the cascaded absorbers. Therefore, a significant reduction of reflection is observed at these frequencies.

6.4.4 Measurements and Results

In order to verify the simulation results, the proposed design of our AFST structure is fabricated and then measured. The fabricated prototype consists of 7 unit cells and they are arranged in a single linear array, which is shown in Figure 6.17. The bandstop FSS pattern is etched on a RT5880 substrate and the metallic walls is fabricated with FR4 substrates, respectively. The commercial absorber of FGM-125 is precisely cut into small pieces and then stacked into the pyramid shape with adhesive glues.

FIGURE 6.17: Fabricated sample of commercial absorber based AFST structure.

The reflection and transmission coefficients of the proposed one-dimensional array of AFST structure are measured in a wideband parallel-plate waveguide (PPW) set-up described in [10]. Since the PPW is a two-port device, a procedure of calibration is performed to obtain accurate measurement results. The measured reflection and transmission of the AFST prototype are under the normal incidence with TE-polarized incident wave. More specifically, the transmission coefficients are obtained by two measurements of an empty waveguide and an AFST in waveguide, respectively. The measured two-port results are represented by $|S_{21}|_{air}$ and $|S_{21}|_{device}$, the insertion loss of the AFST structure is calculated by

$$|S_{21}|_{AFST} = |S_{21}|_{device} - |S_{21}|_{air} \quad (6.7)$$

The reflection coefficient of AFST structure is obtained by measuring an AFST in waveguide and a conducting plate in the waveguide. The conducting plate in the waveguide represents the case of full reflection, which is utilized as a reference to access the reflection from the device under test. The measured one-port reflection coefficients are represented by S_{11}^{pec} and S_{11}^{device} . It is noted that the measured reflection coefficients contain the contributions from the feeding antenna, which may affect the accuracy of the final measurement results. Although the reflections from the feeding antenna and AFST structure are mixed in the frequency domain,

FIGURE 6.18: Simulated and measured results of commercial absorber based AFST structure.

they can be easily distinguished in the time domain. Therefore, an inverse Fourier transform is taken for S_{11}^{pec} and S_{11}^{device} to obtain the time responses of their reflection signals.

$$S(t)_{PEC} = IFFT(S_{11}^{pec}) \quad (6.8)$$

$$S(t)_{device} = IFFT(S_{11}^{device}) \quad (6.9)$$

Moreover, a rectangular window function $w(t)$ is applied to extract the reflection signals from the measured AFST structure, which is defined as

$$w(t) = \begin{cases} 1, & t_1 \leq t \leq t_2 \\ 0, & otherwise \end{cases} \quad (6.10)$$

where t_1 and t_2 are the start and end time of reflection signals from the AFST structure under the test. After adding the window function to the original signals, the pure reflected signals contributed by the AFST structure and PEC plate are represented by

$$S'(t)_{PEC} = w(IFT(S_{11}^{pec})) \quad (6.11)$$

$$S'(t)_{device} = w(IFT(S_{11}^{device})) \quad (6.12)$$

Following this, the frequency responses of the AFST structure and PEC plate are obtained by taking a Fourier transform to convert the time-domain signal to frequency domain again, which are expressed as

$$|S_{11}|'_{PEC} = |FFT[w(IFT(S_{11}^{pec}))]| \quad (6.13)$$

$$|S_{11}|'_{device} = |FFT[w(IFT(S_{11}^{device}))]| \quad (6.14)$$

The final measured reflection coefficient of the AFST structure is calculated by comparing $|S_{11}|'_{device}$ with $|S_{11}|'_{PEC}$, which is expressed by

$$|S_{11}|_{AFST} = |S_{11}|'_{device} - |S_{11}|'_{PEC} \quad (6.15)$$

Figure 6.16 presents the measured transmission and reflection results of the fabricated AFST structure based on commercial absorber. The measured insertion loss is 0.75 dB at 10 GHz and the bandwidth for less than -10 dB reflection is 152.9% from 1.9 GHz to 14.1 GHz. Moreover, the performance of absorption rate is also quite satisfactory, which shows more than 90% absorption rate for lower and upper absorption bands, ranging from 2 GHz to 8.5 GHz and 11.8 GHz to 14 GHz, respectively. Meanwhile, the absorption rate is less than 20% at the pass-band frequencies. It is observed that a slight frequency shift occurs, which might be attributed to the fabrication errors and assembly tolerance. Despite this, an acceptable agreement is obtained between the simulated and measured results, which has verified our design.

6.5 Summary

In this chapter, absorptive frequency-selective transmission structures have been proposed with a characteristic frequency response of absorption-transmission-absorption.

We have introduced two different design methods of using hybrid resonator and commercial absorber for the design of AFST structure. Our methods are able to produce a passband with low insertion loss and maintain a wide absorption band, simultaneously. For the first introduced hybrid resonator based AFST structure, open-circuit and short-circuit resonances are utilized to reduce insertion loss and expand the absorption band, respectively. Furthermore, the performance of AFST structure is improved with a simplified design using commercial absorbers. To obtain wider absorption bandwidth, commercial absorbers are cascaded into multiple layers. Meanwhile, a transmission band with low loss is created by co-designing bandpass and bandstop FSSs inside the AFST unit cell. We expect that the designs of AFST structure can be utilized as radomes to reduce the scattering cross sections of antenna systems.

Chapter 7

Conclusion and Future Work

7.1 Conclusion

AFSR and AFST structures have been applied for the RCS reduction of a variety types of antennas. In this thesis, the investigations on the design concepts and practical implementations of proposed AFSR and AFST structures have been presented. The proposed designs of AFSR and AFST structures are served as an alternative approach to reducing the RCS of antennas with maintained or enhanced radiation performance. The design objectives stated in Chapter 1 have been fulfilled by our research work presented from Chapter 3 to Chapter 6.

In Chapter 3, the design concept of AFSR structure with a frequency response of absorption-reflection-absorption has been proposed, which is further realized by using a periodic array of 3-D lossless and lossy resonators. The presented 3-D AFSR structure is capable of suppressing the RCS of antenna while improving or maintaining its radiation performance. Simulated surface current distribution has revealed a larger area of induced current on an AFSR structure than a PEC ground plane, which leads to a higher gain for AFSR-backed dipole antenna. In addition, the experiments on AFSR-backed dipole and AFSR-backed monopole antennas have been performed. Compared with their counterparts of PEC-backed dipole and monopole antennas, our proposed ones have shown a significant RCS reduction and good radiation performance.

In Chapter 4, the application range of AFSR structure is expanded by exploiting its phase response. More specifically, the reflection phase of an AFSR structure is manipulated by controlling the physical size of its top-loaded patches or varying the height of AFSR resonators. These two approaches serve as a guidance for the designs of low RCS reflectarray antennas and a low RCS parabolic reflector antenna. The radiation performance of our proposed low RCS reflectarray antenna and parabolic reflector antenna is maintained. The scattering cross sections are significantly reduced at the out-of-band frequencies, compared with their counterparts of conventional ones. The experimental results have verified the effectiveness of phase-controllable AFSR structure in RCS reduction of these reflector antennas.

In Chapter 5, a low scattering and beam steerable antenna array is implemented by co-locating active antenna array element with AFSR structure inside the same unit cell. Two prototypes of PEC-backed and AFSR-backed phased array antennas are proposed. The radiating and scattering performance of two antenna arrays are verified through both simulations and experiments. The radiation patterns of AFSR-backed dipole array are capable of achieving a maximum $\pm 45^\circ$ beam steering angles. The aperture efficiency of AFSR-backed array is maintained with 70% with an operating bandwidth of 15%. Significant out-of-band RCS reduction is achieved with 96% and 30% fractional bandwidth at lower and upper bands for co-polarized incident wave. A fractional reduction bandwidth of 142% is obtained for cross-polarized incident wave, compared with the reference PEC-backed dipole antenna array.

In Chapter 6, the designs of low-RCS radomes are proposed, which are implemented by utilizing AFST structures with a frequency response of absorption-transmission-absorption. A transmission band is created within the absorption band for signals within certain frequency band to pass through. Two absorption bands are created at the lower and upper bands of the transmission band, which are intended for the scattering reduction of out-of-band incident waves. Two different design methodologies are proposed, which are based on hybrid resonators and integrated FSSs structures. Improvements are made in the first design by introducing a dual-polarized AFST structure based on hybrid resonators, which has reduced 50% quantity of lumped elements required in a unit cell, compared with the state-of-art designs. In the second design, a single-polarized AFST structure without any

lumped elements has been proposed. Different from all previous designs, commercial absorbers are utilized for the absorption of incident waves. A remarkable reflection reduction of 152.9% fractional bandwidth has been achieved with 10 dB reflection reduction from 1.9 to 14.1 GHz. In addition, a stable transmission band with 10% fractional bandwidths is observed. The transmission and reflection responses of the proposed two AFST structures are stable under an oblique incident angle of 30° . The dispersion of the resonator varies with incidence angle, which leads to the shift of frequency responses. Future work might be focused on improving the performance of AFST structure under large oblique incident angle.

In conclusion, the design concept of AFST and AFSR structure is presented in this thesis with their applications in RCS reduction of antennas. The proposed AFSR structure find its application by incorporating with antennas to achieve RCS reduction while maintaining or enhancing the radiation performance of antennas. The amplitude and phase response of the AFSR structure are both fully exploited for demonstrating the low RCS antennas. Examples of antennas integrated with AFSR structure are introduced, including single antenna, reflectarrays and phase array antenna. Furthermore, the low RCS antenna radomes are demonstrated with improved AFST structures, including a topological optimized AFST structure with reduced number of lumped resistors and FSS-based AFST structure free of lumped resistors. The proposed approaches of low RCS antennas might be proven useful in future communication systems with reduced scattering cross sections.

7.2 Future Works

In this thesis, two EM structures of AFSR and AFST were presented to reduce the RCS of antennas while maintaining or enhancing its radiation performance. In addition to the above mentioned topics, some further research may be carried out in the following directions.

Tunable or switchable low-RCS antennas

In recent years, with the rapid development of wireless communications, functionalities such as frequency hopping and dynamic spectrum allocations are becoming more and more important. Therefore, it is demanded that the radiation pattern

or operating frequency of antennas can be reconfigured. In Chapter 3, we have presented the design concept of using AFSR structure for RCS reduction and gain enhancement of dipole and monopole antennas. However, the introduced AFSR-backed dipole and monopole antennas are not able to change their functionalities. Therefore, reconfigurable or tunable functionalities can be introduced in the design of low-RCS AFSR structures. The resonate frequency of AFSR structure and antenna can be tuned with pin-diodes or varactors. As a result, the operating frequency of antennas can be reconfigured simultaneously with the AFSR structure. Meanwhile, the RCS of such reconfigurable or tunable antennas will be significantly lower than the conventional ones.

RCS reduction with parabolic surface profile

This thesis introduces the low scattering reflectarray antenna and parabolic reflector antenna based on AFSR structure in Chapter 4. It is observed that the scattering from the reflector is low at the in-band frequency of antenna. The main reason is attributed to the unique reflection phase distribution of a planar reflector, which is characterized by a parabolic phase profile. This specific phase profile is capable of converging incident plane wave at its focal point and then diverges after passing through the focal point. Therefore, compared with a flat PEC plate, significant RCS reduction can be obtained for this planar reflector with size variable patches. The parabolic phase distribution pattern is different from the existing chessboard surface or other randomly distributed coding metasurface, which has a potential to be applied in the RCS reduction of planar surfaces in the future.

Low-RCS transmitarray antennas

The AFST structure based on integrated FSSs is introduced in Chapter 6. The bandpass FSS provides the pass band and the bandstop FSS bypasses the losses at the transmission band. This design is featured free of lumped elements; only thin commercial absorbers are utilized for the absorption. This commercial absorber based design is more suitable for high power applications. The limitation of the current proposed design is that it is sensitive to the polarization of incident waves. In future works, the design can be expanded into a dual-polarized one. Apart from that, the design concept of controlling reflection phase of an AFSR structure is applied in the low scattering reflectarray antenna. For an AFST structure, it is

featured by an amplitude response of absorption-transmission-absorption. However, at the transmission band of an AFST structure, the phase delay is a constant value. More functionalities can be introduced by controlling the transmitted phase, such as wavefront transformation and beam steering. One potential demonstration of this phase-controllable AFST is to design a transmitarray antenna with high gain and low RCS. At the transmission band, the radiated wave front can be transformed from a cylindrical wave into a planar wave with a high directivity. At the absorption band, the scattering of transmitarray is reduced due to the absorption of the absorbers. Apart from that, the transmission phase of each unit cell can be further tuned or reconfigured. The phase delay profiles can be implemented with different preset values to achieve reconfigurable radiation patterns.

List of Author's Awards and Publications

Awards

- **2018 Best Student Paper Competition Award (First Prize)**
 - IEEE Singapore AP/MTT Joint Chapter
 - For the journal paper titled “3-D absorptive frequency selective reflector for antenna radar cross section reduction”.
- **Honorable Mention**
 - Student Paper Competition in 2018 IEEE AP-S/URSI Symposium.
 - For the conference paper titled “Realization of low-RCS parabolic reflector antenna using curved 3-D frequency-selective structure”.

Journal Articles

- **H. Huang** and Z. Shen, “Absorptive frequency-selective transmission structure with square-loop hybrid resonator,” *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 3212-3215, 2017.
- **H. Huang**, Z. Shen and A. A. Omar, “3-D absorptive frequency selective reflector for antenna radar cross section reduction,” *IEEE Trans. Antennas Propag.*, vol. 65, no. 11, pp. 5908-5917, Nov. 2017.
- **H. Huang** and Z. Shen, “Low-RCS reflectarray with phase controllable absorptive frequency-selective reflector,” *IEEE Trans. Antennas Propag.*, vol. 67, no. 1, pp. 190-198, Jan. 2019.

- **H. Huang**, A. A. Omar and Z. Shen, "Low-RCS and beam-steerable dipole array using absorptive frequency-selective reflection structures," *IEEE Trans. Antennas Propag.*, vol. 68, no. 3, pp. 2457-2462, Mar. 2020.
- **H. Huang** and Z. Shen, "Brewster lens with perfect wave refraction," *IEEE Trans. Antennas Propag.*, vol. 68, no. 8, pp. 6204-6213, Aug. 2020.
- **H. Huang** and Z. Shen, "Angle-selective surface based on uniaxial dielectric-magnetic slab," *IEEE Antennas Wireless Propag. Lett.*, Early Access, doi: 10.1109/LAWP.2020.3035439.

Conference Proceedings

- **H. Huang** and Z. Shen, "An absorptive frequency-selective reflector based on miniaturized square-loop resonator," *12th European Conference on Antennas and Propagation (EuCAP)*, London, UK, Apr. 2018.
- **H. Huang** and Z. Shen, "Realization of low-RCS parabolic reflector antenna using curved 3-D frequency-selective structure," *IEEE International Symposium on Antennas and Propagation (AP-S/URSI)*, Boston, MA, USA, Jul. 2018, pp. 1723-1724.
- **H. Huang**, A. A. Omar and Z. Shen, "Absorptive frequency-selective reflectors," *IEEE International Conference on Computational Electromagnetics (ICCEM)*, Shanghai, China, Mar. 2019.
- **H. Huang** and Z. Shen, "High-efficiency Brewster lens for arbitrary wave-front control," *IEEE MTT-S International Wireless Symposium (IWS)*, Guangzhou, China, May 2019.
- **H. Huang** and Z. Shen, "Ultra-Broadband 3-D absorptive frequency-selective transmission structure using commercial absorber," *IEEE International Symposium on Antennas and Propagation (AP-S/URSI)*, Montréal, QC, CA, Jul. 2020.

Bibliography

- [1] E. Knott, J. Schaeffer, and M. Tulley, *Radar Cross Section*. New York, USA: SciTech Publishing, 2004.
- [2] D. Lynch, *Introduction to RF Stealth*. Raleigh, NC, USA: SciTech, 2004.
- [3] R. C. Hansen, “Relationships between antennas as scatterers and as radiators,” *Proc. of the IEEE*, vol. 77, no. 5, pp. 659–662, May 1989.
- [4] J. P. Turpin, P. E. Sieber, and D. H. Werner, “Absorbing ground planes for reducing planar antenna radar cross-section based on frequency selective surfaces,” *IEEE Antennas Wireless Propag. Lett.*, vol. 12, pp. 1456–1459, 2013.
- [5] S. Genovesi, F. Costa, and A. Monorchio, “Wideband radar cross section reduction of slot antennas arrays,” *IEEE Trans. Antennas Propag.*, vol. 62, no. 1, pp. 163–173, Jan. 2014.
- [6] Y. Liu, N. Li, Y. Jia, W. Zhang, and Z. Zhou, “Low RCS and high-gain patch antenna based on a holographic metasurface,” *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 3, pp. 492–496, 2019.
- [7] Q. Zheng, C. Guo, G. A. E. Vandenbosch, and J. Ding, “Low-profile circularly polarized array with gain enhancement and RCS reduction using polarization conversion ebg structures,” *IEEE Trans. Antennas Propag.*, vol. 68, no. 3, pp. 2440–2445, Mar. 2020.
- [8] Y. Zheng, Y. Zhou, J. Gao, X. Cao, H. Yang, S. Li, L. Xu, J. Lan, and L. Jidi, “Ultra-wideband polarization conversion metasurface and its application cases for antenna radiation enhancement and scattering suppression,” *Sci. Rep.*, vol. 7, p. 16137, 2017.
- [9] Y. Han, W. Che, C. Christopoulos, Y. Xiong, and Y. Chang, “A fast and efficient design method for circuit analog absorbers consisting of resistive square-loop arrays,” *IEEE Trans. Electromagn. Compat.*, vol. 58, no. 3, pp. 747–757, Mar. 2016.
- [10] Y. Shang, Z. Shen, and S. Xiao, “On the design of single-layer circuit analog absorber using double-square-loop array,” *IEEE Trans. Antennas Propag.*, vol. 61, no. 12, pp. 6022–6029, Dec. 2013.

- [11] J. W. Yu, Y. Cai, X. Q. Lin, and X. Wang, "Perforated multilayer ultrawideband absorber based on circuit analog absorber with optimal air spaces," *IEEE Antennas Wireless Propag. Lett.*, vol. 19, no. 1, pp. 34–38, 2020.
- [12] A. A. Omar and Z. Shen, "Double-sided parallel-strip line resonator for dual-polarized 3-D frequency-selective structure and absorber," *IEEE Trans. Microw. Theory Tech.*, vol. 65, no. 10, pp. 3744–3752, Oct. 2017.
- [13] F. Costa, A. Monorchio, and G. Manara, "Analysis and design of ultra thin electromagnetic absorbers comprising resistively loaded high impedance surfaces," *IEEE Trans. Antennas Propag.*, vol. 58, no. 5, pp. 1551–1558, May 2010.
- [14] L. Li and Z. Lv, "Ultra-wideband polarization-insensitive and wide-angle thin absorber based on resistive metasurfaces with three resonant modes," *J. Appl. Phys.*, vol. 122, no. 5, p. 055104, 2017.
- [15] Y. Zheng, K. Chen, T. Jiang, J. Zhao, and Y. Feng, "Multi-octave microwave absorption via conformal metamaterial absorber with optical transparency," *J. Phys. D: Appl. Phys.*, vol. 52, no. 33, p. 335101, Jun. 2019.
- [16] Z. Zhou, K. Chen, J. Zhao, P. Chen, T. Jiang, B. Zhu, Y. Feng, and Y. Li, "Metasurface Salisbury screen: achieving ultra-wideband microwave absorption," *Opt. Exp.*, vol. 25, p. 30241, Nov. 2017.
- [17] J. Chen, Z. Hu, G. Wang, X. Huang, S. Wang, X. Hu, and M. Liu, "High-impedance surface-based broadband absorbers with interference theory," *IEEE Trans. Antennas Propag.*, vol. 63, no. 10, pp. 4367–4374, Oct. 2015.
- [18] W. W. Salisbury, "Absorbent body for electromagnetic waves," Patent US2 599 944, Jun., 1952.
- [19] B. Chambers and A. Tennant, "Characteristics of a Salisbury screen radar absorber covered by a dielectric skin," *Electron. Lett.*, vol. 30, no. 21, pp. 1797–1799, 1994.
- [20] R. L. Fante and M. T. McCormack, "Reflection properties of the Salisbury screen," *IEEE Trans. Antennas Propag.*, vol. 36, no. 10, pp. 1443–1454, Oct. 1988.
- [21] B. Munk, *Finite Antenna Arrays and FSS*. New York, USA: Wiley, 2003.
- [22] W. Li, T. Wu, W. Wang, P. Zhai, and J. Guan, "Broadband patterned magnetic microwave absorber," *J. Appl. Phys.*, vol. 116, no. 4, p. 044110, 2014.
- [23] W. Li, T. Wu, W. Wang, J. Guan, and P. Zhai, "Integrating non-planar metamaterials with magnetic absorbing materials to yield ultra-broadband microwave hybrid absorbers," *Appl. Phys. Lett.*, vol. 104, no. 2, p. 022903, 2014.

- [24] F. Ding, Y. Cui, X. Ge, Y. Jin, and S. He, "Ultra-broadband microwave metamaterial absorber," *Appl. Phys. Lett.*, vol. 100, no. 10, p. 103506, 2012.
- [25] K. N. Rozanov, "Ultimate thickness to bandwidth ratio of radar absorbers," *IEEE Trans. Antennas Propag.*, vol. 48, no. 8, pp. 1230–1234, Aug. 2000.
- [26] J. Chen, Y. Shang, and C. Liao, "Double-layer circuit analog absorbers based on resistor-loaded square-loop arrays," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 4, pp. 591–595, 2018.
- [27] J. W. Yu, Y. Cai, X. Q. Lin, and X. Wang, "Perforated multilayer ultrawideband absorber based on circuit analog absorber with optimal air spaces," *IEEE Antennas Wireless Propag. Lett.*, vol. 19, no. 1, pp. 34–38, 2020.
- [28] A. Sharma, R. Panwar, and R. Khanna, "Experimental validation of a frequency-selective surface-loaded hybrid metamaterial absorber with wide bandwidth," *IEEE Magn. Lett.*, vol. 10, pp. 1–5, 2019.
- [29] Y. Shen, J. Zhang, Y. Pang, J. Wang, and S. Qu, "Hybrid metamaterial absorber based on the combination of plasmonic structure and magnetic absorber," in *2019 IEEE Int. Conf. Comput. Electromagn. (ICCEM)*, 2019, pp. 1–2.
- [30] X. Qi, J. Xu, Q. Hu, Y. Deng, R. Xie, Y. Jiang, W. Zhong, and Y. Du, "Metal-free carbon nanotubes: synthesis, and enhanced intrinsic microwave absorption properties," *Sci. Rep.*, vol. 6, p. 28310, 2016.
- [31] M. Green and X. Chen, "Recent progress of nanomaterials for microwave absorption," *J. Materiomics*, vol. 5, no. 4, pp. 503 – 541, 2019.
- [32] W. Chen, C. A. Balanis, and C. R. Birtcher, "Checkerboard EBG surfaces for wideband radar cross section reduction," *IEEE Trans. Antennas Propag.*, vol. 63, no. 6, pp. 2636–2645, Jun. 2015.
- [33] M. Amano and Y. Kotsuka, "A method of effective use of ferrite for microwave absorber," *IEEE Trans. Microw. Theory Tech.*, vol. 51, no. 1, pp. 238–245, Jan. 2003.
- [34] L. Zhang, P. Zhou, H. Zhang, L. Lu, G. Zhang, H. Chen, H. Lu, J. Xie, and L. Deng, "A broadband radar absorber based on perforated magnetic polymer composites embedded with FSS," *IEEE Trans. Magn.*, vol. 50, no. 5, pp. 1–5, May 2014.
- [35] L. Sun, H. Cheng, Y. Zhou, and J. Wang, "Design of a lightweight magnetic radar absorber embedded with resistive FSS," *IEEE Antennas Wireless Propag. Lett.*, vol. 11, pp. 675–677, 2012.
- [36] D. Sievenpiper, Lijun Zhang, R. F. J. Broas, N. G. Alexopolous, and E. Yablonovitch, "High-impedance electromagnetic surfaces with a forbidden frequency band," *IEEE Trans. Microw. Theory Tech.*, vol. 47, no. 11, pp. 2059–2074, Nov. 1999.

- [37] M. Paquay, J. Iriarte, I. Ederra, R. Gonzalo, and P. de Maagt, “Thin AMC structure for radar cross-section reduction,” *IEEE Trans. Antennas Propag.*, vol. 55, no. 12, pp. 3630–3638, Dec. 2007.
- [38] Y. Jia, Y. Liu, Y. J. Guo, K. Li, and S. Gong, “Broadband polarization rotation reflective surfaces and their applications to rcs reduction,” *IEEE Trans. Antennas Propag.*, vol. 64, no. 1, pp. 179–188, Jan. 2016.
- [39] J. C. Iriarte Galarregui, A. Tellechea Pereda, J. L. M. de Falcón, I. Ederra, R. Gonzalo, and P. de Maagt, “Broadband radar cross-section reduction using AMC technology,” *IEEE Trans. Antennas Propag.*, vol. 61, no. 12, pp. 6136–6143, Dec. 2013.
- [40] K. Wang, J. Zhao, Q. Cheng, T. J. Cui *et al.*, “Broadband and broad-angle low-scattering metasurface based on hybrid optimization algorithm,” *Sci. Rep.*, vol. 4, p. 5935, 2014.
- [41] S. H. Esmaeli and S. H. Sedighy, “Wideband radar cross-section reduction by AMC,” *Elect. Lett.*, vol. 52, no. 1, pp. 70–71, 2016.
- [42] A. Y. Modi, C. A. Balanis, C. R. Birtcher, and H. N. Shaman, “Novel design of ultrabroadband radar cross section reduction surfaces using artificial magnetic conductors,” *IEEE Trans. Antennas Propag.*, vol. 65, no. 10, pp. 5406–5417, Oct. 2017.
- [43] M.-J. Haji-Ahmadi, V. Nayyeri, M. Soleimani, and O. Ramahi, “Pixelated checkerboard metasurface for ultra-wideband radar cross section reduction,” *Sci. Rep.*, vol. 7, p. 11437, 2017.
- [44] Y. Lu, J. Su, J. Liu, Q. Guo, H. Yin, Z. Li, and J. Song, “Ultrawideband monostatic and bistatic RCS reductions for both copolarization and cross polarization based on polarization conversion and destructive interference,” *IEEE Trans. Antennas Propag.*, vol. 67, no. 7, pp. 4936–4941, Jul. 2019.
- [45] J. Su, Y. Lu, J. Liu, Y. Yang, Z. Li, and J. Song, “A novel checkerboard metasurface based on optimized multielement phase cancellation for super-wideband rcs reduction,” *IEEE Trans. Antennas Propag.*, vol. 66, no. 12, pp. 7091–7099, Dec. 2018.
- [46] M. Akbari, F. Samadi, A. Sebak, and T. A. Denidni, “Superbroadband diffuse wave scattering based on coding metasurfaces: Polarization conversion metasurfaces,” *IEEE Antennas Propag. Mag.*, vol. 61, no. 2, pp. 40–52, Feb. 2019.
- [47] E. Ameri, S. Esmaeli, and S. H. Sedighy, “Ultra wideband radar cross section reduction by using polarization conversion metasurfaces,” *Sci. Rep.*, vol. 9, p. 478, 2019.

- [48] H. Chen, J. Wang, H. Ma, S. Qu, Z. Xu, A. Zhang, M. Yan, and Y. Li, “Ultra-wideband polarization conversion metasurfaces based on multiple plasmon resonances,” *J. Appl. Phys.*, vol. 115, no. 15, p. 154504, 2014.
- [49] Z. Bomzon, V. Kleiner, and E. Hasman, “Pancharatnam-Berry phase in space-variant polarization-state manipulations with subwavelength gratings,” *Opt. Lett.*, vol. 26, no. 18, pp. 1424–1426, Sep. 2001.
- [50] L. Marrucci, C. Manzo, and D. Paparo, “Pancharatnam-Berry phase optical elements for wave front shaping in the visible domain: switchable helical mode generation,” *Appl. Phys. Lett.*, vol. 88, no. 22, p. 221102, 2006.
- [51] G. Milione, S. Evans, D. A. Nolan, and R. R. Alfano, “Higher order Pancharatnam-Berry phase and the angular momentum of light,” *Phys. Rev. Lett.*, vol. 108, p. 190401, May 2012.
- [52] H. Sun, C. Gu, X. Chen, L. Liu, B. Xu, and Z. Zhou, “Broadband and broad-angle polarization-independent metasurface for radar cross section reduction,” *Sci. Rep.*, vol. 7, p. 40782, 2016.
- [53] P. Su, Y. Zhao, S. Jia, W. Shi, and H. Wang, “An ultra-wideband and polarization-independent metasurface for RCS reduction,” *Sci. Rep.*, vol. 6, p. 20387, 2016.
- [54] Y. Jia, Y. Liu, Y. J. Guo, K. Li, and S. Gong, “A dual-patch polarization rotation reflective surface and its application to ultra-wideband RCS reduction,” *IEEE Trans. Antennas Propag.*, vol. 65, no. 6, pp. 3291–3295, Jun. 2017.
- [55] H. Rajabalipanah and A. Abdolali, “Ultrabroadband monostatic/bistatic RCS reduction via high-entropy phase-encoded polarization conversion metasurfaces,” *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 6, pp. 1233–1237, 2019.
- [56] W. Pan, C. Huang, M. Pu, X. Ma, J. Cui, B. Zhao, and X. Luo, “Combining the absorptive and radiative loss in metasurfaces for multi-spectral shaping of the electromagnetic scattering,” *Sci. Rep.*, vol. 6, p. 21462, 2016.
- [57] C. Ji, C. Huang, X. Zhang, J. Yang, J. Song, and X. Luo, “Broadband low-scattering metasurface using a combination of phase cancellation and absorption mechanisms,” *Opt. Express*, vol. 27, no. 16, pp. 23 368–23 377, Aug. 2019.
- [58] S. Sui, H. Ma, J. Wang, Y. Pang, M. Feng, Z. Xu, and S. Qu, “Absorptive coding metasurface for further radar cross section reduction,” *J. Phys. D: Appl. Phys.*, vol. 51, no. 6, p. 065603, Jan. 2018.
- [59] F.-F. Li, Q. Lou, P. Chen, Y. Poo, and R.-X. Wu, “Broadband backscattering reduction realized by array of lossy scatterers,” *Opt. Express*, vol. 26, no. 26, pp. 34 711–34 718, Dec. 2018.

- [60] M. Gustafsson, "RCS reduction of integrated antenna arrays and radomes with resistive sheets," in *2006 IEEE Antennas Propag. Soc. Int. Symp.*, 2006, pp. 3479–3482.
- [61] Y. Li, H. Zhang, Y. Fu, and N. Yuan, "RCS reduction of ridged waveguide slot antenna array using EBG radar absorbing material," *IEEE Antennas Wireless Propag. Lett.*, vol. 7, pp. 473–476, 2008.
- [62] Q. Chen and Y. Fu, "A planar stealthy antenna radome using absorptive frequency selective surface," *Microw. Opt. Tech. Lett.*, vol. 56, no. 8, pp. 1788–1792, 2014.
- [63] Y. Liu and X. Zhao, "Perfect absorber metamaterial for designing low-RCS patch antenna," *IEEE Antennas Wireless Propag. Lett.*, vol. 13, pp. 1473–1476, 2014.
- [64] C. Huang, W. Pan, X. Ma, and X. Luo, "Wideband radar cross-section reduction of a stacked patch array antenna using metasurface," *IEEE Antennas Wireless Propag. Lett.*, vol. 14, pp. 1369–1372, 2015.
- [65] T. Liu, X. Cao, J. Gao, Q. Zheng, W. Li, and H. Yang, "RCS reduction of waveguide slot antenna with metamaterial absorber," *IEEE Trans. Antennas Propag.*, vol. 61, no. 3, pp. 1479–1484, Mar. 2013.
- [66] Q. Chen, M. Guo, D. Sang, Z. Sun, and Y. Fu, "RCS reduction of patch array antenna using anisotropic resistive metasurface," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 6, pp. 1223–1227, 2019.
- [67] Y. Han, L. Zhu, Y. Bo, W. Che, and B. Li, "Novel low-RCS circularly polarized antenna arrays via frequency-selective absorber," *IEEE Trans. Antennas Propag.*, vol. 68, no. 1, pp. 287–296, Jan. 2020.
- [68] Y. Zheng, J. Gao, X. Cao, Z. Yuan, and H. Yang, "Wideband RCS reduction of a microstrip antenna using artificial magnetic conductor structures," *IEEE Antennas Wireless Propag. Lett.*, vol. 14, pp. 1582–1585, 2015.
- [69] L. Jidi, X. Cao, W. Li, J. Gao, Y. Zheng, T. Li, and Y. Tang, "X-band and circularly polarized antenna with inborn RCS reduction," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 8, pp. 1501–1504, Aug. 2018.
- [70] Y. Zhao, J. Gao, X. Cao, T. Liu, L. Xu, X. Liu, and L. Cong, "In-band RCS reduction of waveguide slot array using metasurface bars," *IEEE Trans. Antennas Propag.*, vol. 65, no. 2, pp. 943–947, Feb. 2017.
- [71] Y. Zhao, X. Cao, J. Gao, X. Yao, T. Liu, W. Li, and S. Li, "Broadband low-RCS metasurface and its application on antenna," *IEEE Trans. Antennas Propag.*, vol. 64, no. 7, pp. 2954–2962, Jul. 2016.
- [72] Z. Liu, Y. Liu, and S. Gong, "Gain enhanced circularly polarized antenna with RCS reduction based on metasurface," *IEEE Access*, vol. 6, pp. 46 856–46 862, 2018.

- [73] Y. Liu, Y. Hao, K. Li, and S. Gong, "Radar cross section reduction of a microstrip antenna based on polarization conversion metamaterial," *IEEE Antennas Wireless Propag. Lett.*, vol. 15, pp. 80–83, 2016.
- [74] W. Pan, C. Huang, P. Chen, X. Ma, C. Hu, and X. Luo, "A low-RCS and high-gain partially reflecting surface antenna," *IEEE Trans. Antennas Propag.*, vol. 62, no. 2, pp. 945–949, Feb. 2014.
- [75] Y. Zheng, J. Gao, Y. Zhou, X. Cao, H. Yang, S. Li, and T. Li, "Wideband gain enhancement and RCS reduction of Fabry-Perot resonator antenna with chessboard arranged metamaterial superstrate," *IEEE Trans. Antennas Propag.*, vol. 66, no. 2, pp. 590–599, Feb. 2018.
- [76] L. Zhang, X. Wan, S. Liu, J. Y. Yin, Q. Zhang, H. T. Wu, and T. J. Cui, "Realization of low scattering for a high-gain Fabry-Perot antenna using coding metasurface," *IEEE Trans. Antennas Propag.*, vol. 65, no. 7, pp. 3374–3383, Jul. 2017.
- [77] M. Long, W. Jiang, and S. Gong, "Wideband RCS reduction using polarization conversion metasurface and partially reflecting surface," *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 2534–2537, 2017.
- [78] C. Huang, W. Pan, X. Ma, and X. Luo, "A frequency reconfigurable directive antenna with wideband low-RCS property," *IEEE Trans. Antennas Propag.*, vol. 64, no. 3, pp. 1173–1178, Mar. 2016.
- [79] W. B. M. Walter S. Arceneaux, Rickey D. Akins, "Absorptive/transmissive radome," Patent US5 400 043A, March, 1994.
- [80] S. Yu, N. Kou, Z. Ding, and Z. Zhang, "Harmonic-suppressed frequency selective rasorber using resistive-film sheet and square-loops resonator," *IEEE Antennas Wireless Propag. Lett.*, vol. 19, no. 2, pp. 292–296, 2020.
- [81] B. Munk, *Metamaterials: Critique and Alternatives*. Wiley, 2009.
- [82] Y. Shang, Z. Shen, and S. Xiao, "Frequency-selective rasorber based on square-loop and cross-dipole arrays," *IEEE Trans. Antennas Propag.*, vol. 62, no. 11, pp. 5581–5589, Nov. 2014.
- [83] X. Xiu, W. Che, W. Yang, Y. Han, and Q. Xue, "A highly selective rasorber based on second-order resonance," *IEEE Antennas Wireless Propag. Lett.*, vol. 19, no. 2, pp. 223–227, 2020.
- [84] X. Zhang, W. Wu, L. Huang, Y. Ma, and N. Yuan, "Design of dual-absorptive-bands frequency selective rasorber with Minkowski loop arrays," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 9, pp. 1843–1847, 2019.
- [85] Q. Yu, S. Liu, A. Monorchio, X. Kong, Y. Wen, and Z. Huang, "A miniaturized high-selectivity frequency selective rasorber based on subwavelength resonance and interdigital resonator," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 9, pp. 1833–1837, 2019.

- [86] M. Guo, Q. Chen, D. Sang, Y. Zheng, and Y. Fu, "Dual-polarized dual-band frequency selective rasorber with low insertion loss," *IEEE Antennas Wireless Propag. Lett.*, vol. 19, no. 1, pp. 148–152, 2020.
- [87] M. Guo, Q. Chen, Z. Sun, D. Sang, and Y. Fu, "Design of dual-band frequency-selective rasorber," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 5, pp. 841–845, 2019.
- [88] H. Chen, W. Lu, Z. Liu, and Z. H. Jiang, "Flexible rasorber based on graphene with energy manipulation function," *IEEE Trans. Antennas Propag.*, vol. 68, no. 1, pp. 351–359, Jan. 2020.
- [89] L. Wu, S. Zhong, J. Huang, and T. Liu, "Broadband frequency-selective rasorber with varactor-tunable interabsorption band transmission window," *IEEE Trans. Antennas Propag.*, vol. 67, no. 9, pp. 6039–6050, Sep. 2019.
- [90] G. Qian, J. Zhao, X. Ren, K. Chen, T. Jiang, Y. Feng, and Y. Liu, "Switchable broadband dual-polarized frequency-selective rasorber/absorber," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 12, pp. 2508–2512, 2019.
- [91] Y. Han, W. Che, X. Xiu, W. Yang, and C. Christopoulos, "Switchable low-profile broadband frequency-selective rasorber/absorber based on slot arrays," *IEEE Trans. Antennas Propag.*, vol. 65, no. 12, pp. 6998–7008, Dec. 2017.
- [92] X. Xiu, Y. Han, W. Che, and W. Yang, "Double-polarization frequency selective rasorber based on cross-frame and circle ring slot arrays," in *2017 10th Global Symp. Millimeter-Waves*, May 2017, pp. 109–111.
- [93] X. Yan, X. Kong, Q. Wang, L. Xing, F. Xue, Y. Xu, S. Jiang, and X. Liu, "Water-based reconfigurable frequency selective rasorber with thermally tunable absorption band," *IEEE Trans. Antennas Propag.*, pp. 1–1, 2020.
- [94] W. Yu, G. Q. Luo, Y. Yu, Z. Liao, H. Jin, and Z. Shen, "Broadband band-absorptive frequency-selective rasorber with a hybrid 2-D and 3-D structure," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 8, pp. 1701–1705, 2019.
- [95] Y. Yu, Z. Shen, T. Deng, and G. Luo, "3-D frequency-selective rasorber with wide upper absorption band," *IEEE Trans. Antennas Propag.*, vol. 65, no. 8, pp. 4363–4367, Aug. 2017.
- [96] Y. Zhang, B. Li, L. Zhu, Y. Tang, Y. Chang, and Y. Bo, "Frequency selective rasorber with low insertion loss and dual-band absorptions using planar slotline structures," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 4, pp. 633–636, 2018.
- [97] Z. Shen, J. Wang, and B. Li, "3-D frequency selective rasorber: Concept, analysis, and design," *IEEE Trans. Microw. Theory Tech.*, vol. 64, no. 10, pp. 3087–3096, Oct. 2016.

- [98] B. Li and Z. Shen, "Wideband 3D frequency selective rasorber," *IEEE Trans. Antennas Propag.*, vol. 62, no. 12, pp. 6536–6541, Dec. 2014.
- [99] F. Costa and A. Monorchio, "A frequency selective radome with wideband absorbing properties," *IEEE Trans. Antennas Propag.*, vol. 60, no. 6, pp. 2740–2747, Jun. 2012.
- [100] S. Hamid, H. Shakhtour, and D. Heberling, "Frequency selective radome with enhanced transmissive and absorptive response," in *2014 Loughborough Antennas Propag. Conf. (LAPC)*, 2014, pp. 36–39.
- [101] G. I. Kiani, A. R. Weily, and K. P. Esselle, "A novel absorb/transmit FSS for secure indoor wireless networks with reduced multipath fading," *IEEE Microw. Wireless Compon. Lett.*, vol. 16, no. 6, pp. 378–380, 2006.
- [102] Q. Chen, S. Yang, J. Bai, and Y. Fu, "Design of absorptive/transmissive frequency-selective surface based on parallel resonance," *IEEE Trans. Antennas Propag.*, vol. 65, no. 9, pp. 4897–4902, Sep. 2017.
- [103] Q. Chen, J. Bai, L. Chen, and Y. Fu, "A miniaturized absorptive frequency selective surface," *IEEE Antennas Wireless Propag. Lett.*, vol. 14, pp. 80–83, 2015.
- [104] H. Zhou, L. Yang, S. Qu, K. Wang, J. Wang, H. Ma, and Z. Xu, "Experimental demonstration of an absorptive/transmissive FSS with magnetic material," *IEEE Antennas Wireless Propag. Lett.*, vol. 13, pp. 114–117, 2014.
- [105] F. Costa, A. Monorchio, and G. Manara, "Ultra-thin absorbers by using high impedance surfaces with resistive frequency selective surfaces," in *2007 IEEE Antennas Propag. Society Int. Symp.*, 2007, pp. 861–864.
- [106] G. I. Kiani, A. R. Weily, and K. P. Esselle, "Frequency selective surface absorber using resistive cross-dipoles," in *2006 IEEE Antennas Propag. Society Int. Symp.*, 2006, pp. 4199–4202.
- [107] Z. Sun, Q. Chen, M. Guo, H. Yu, and Y. Fu, "Frequency selective rasorber and reflector with two-sided absorption bands," *IEEE Access*, vol. 7, pp. 6025–6031, 2019.
- [108] Z. Wang, Q. Zeng, J. Fu, W. Chen, B. Lv, M. Song, and T. A. Denidni, "A high-transmittance frequency-selective rasorber based on dipole arrays," *IEEE Access*, vol. 6, pp. 31 367–31 374, 2018.
- [109] X. Zhang, W. Wu, Y. Ma, C. Wang, C. Li, and N. Yuan, "Design dual-polarization frequency selective rasorber using split ring resonators," *IEEE Access*, vol. 7, pp. 101 139–101 146, 2019.
- [110] Y. Sun, S. Xiao, Z. Yao, and Q. Shi, "A wideband frequency-selective rasorber based on interdigital resonator and fractal shaped slot," in *2019 IEEE MTT-S Int. Wireless Symp. (IWS)*, 2019, pp. 1–3.

- [111] M. Qu, S. Sun, L. Deng, and S. Li, "Design of a frequency-selective rasorber based on notch structure," *IEEE Access*, vol. 7, pp. 3704–3711, 2019.
- [112] W. Yu, G. Q. Luo, Y. Yu, Y. Pan, W. Cao, Y. Pan, and Z. Shen, "Dual-polarized band-absorptive frequency selective rasorber using meander-line and lumped resistors," *IEEE Trans. Antennas Propag.*, vol. 67, no. 2, pp. 1318–1322, Feb. 2019.
- [113] L. Wang, S. Liu, X. Kong, H. Zhang, Q. Yu, and Y. Wen, "Frequency-selective rasorber with a wide high-transmission passband based on multiple coplanar parallel resonances," *IEEE Antennas Wireless Propag. Lett.*, vol. 19, no. 2, pp. 337–340, 2020.
- [114] Q. Chen, D. Sang, M. Guo, and Y. Fu, "Frequency-selective rasorber with interabsorption band transparent window and interdigital resonator," *IEEE Trans. Antennas Propag.*, vol. 66, no. 8, pp. 4105–4114, Aug. 2018.
- [115] Z. Wang, J. Fu, Q. Zeng, M. Song, and T. A. Denidni, "High-transmittance absorptive structure based on effect of electromagnetically induced transparency," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 12, pp. 2463–2467, 2019.
- [116] X. Xiu, W. Che, Y. Han, and W. Yang, "Low-profile dual-polarization frequency-selective rasorbers based on simple-structure lossy cross-frame elements," *IEEE Antennas Wireless Propag. Lett.*, vol. 17, no. 6, pp. 1002–1005, 2018.
- [117] Y. Yu, G. Q. Luo, Q. Liu, W. Yu, H. Jin, Z. Liao, and Z. Shen, "3D band-absorptive frequency selective rasorber: Concept and analysis," *IEEE Access*, vol. 7, pp. 2520–2528, 2019.
- [118] A. A. Omar, Z. Shen, and H. Huang, "Absorptive frequency-selective reflection and transmission structures," *IEEE Trans. Antennas Propag.*, vol. 65, no. 11, pp. 6173–6178, Nov. 2017.
- [119] Y. Wang, S. Qi, Z. Shen, and W. Wu, "Ultra-thin 3-D frequency selective rasorber with wide absorption bands," *IEEE Trans. Antennas Propag.*, vol. 68, no. 6, pp. 4697–4705, Jun. 2020.
- [120] T. Deng, Y. Yu, Z. Shen, and Z. N. Chen, "Design of 3-D multilayer ferrite-loaded frequency-selective rasorbers with wide absorption bands," *IEEE Trans. Microw. Theory Tech.*, vol. 67, pp. 108–117, Jan. 2019.
- [121] Y. Yu, T. Deng, and Z. Shen, "Wideband 3D frequency selective rasorber based on ferrite absorber," in *2017 IEEE Int. Symp. Antennas Propag. US-NC/URSI Nat. Rad. Sci. Meeting*, 2017, pp. 681–682.
- [122] G. I. Kiani, K. L. Ford, K. P. Esselle, A. R. Weily, and C. J. Panagamuwa, "Oblique incidence performance of a novel frequency selective surface absorber," *IEEE Trans. Antennas Propag.*, vol. 55, no. 10, pp. 2931–2934, Oct. 2007.

- [123] Q. Chen, L. Liu, L. Chen, J. Bai, and Y. Fu, "Absorptive frequency selective surface using parallel lc resonance," *Electron. Lett.*, vol. 52, no. 6, pp. 418–419, 2016.
- [124] Q. Chen, J. Bai, L. Chen, and Y. Fu, "A miniaturized absorptive frequency selective surface," *IEEE Antennas Wireless Propag. Lett.*, vol. 14, pp. 80–83, 2015.
- [125] A. K. Rashid and Z. Shen, "A novel band-reject frequency selective surface with pseudo-elliptic response," *IEEE Trans. Antennas Propag.*, vol. 58, no. 4, pp. 1220–1226, Apr. 2010.
- [126] B. Li and Z. Shen, "Three-dimensional bandpass frequency-selective structures with multiple transmission zeros," *IEEE Trans. Microw. Theory Tech.*, vol. 61, no. 10, pp. 3578–3589, Oct. 2013.
- [127] Y. Wang, S. Qi, W. Wu, and Z. Shen, "An ultra-thin wideband 3-D frequency selective rasorber based on ferrite absorber and slow wave structure," in *2019 IEEE Int. Symp. Antennas Propag. and USNC-URSI Rad. Sci. Meeting*, 2019, pp. 957–958.
- [128] A. A. Omar and Z. Shen, "Thin 3-D bandpass frequency-selective structure based on folded substrate for conformal radome applications," *IEEE Trans. Antennas Propag.*, vol. 67, no. 1, pp. 282–290, Jan. 2019.
- [129] X. Q. Lin, P. Mei, P. C. Zhang, Z. Z. D. Chen, and Y. Fan, "Development of a resistor-loaded ultrawideband absorber with antenna reciprocity," *IEEE Trans. Antennas Propag.*, vol. 64, no. 11, pp. 4910–4913, 2016.
- [130] P. Mei, X. Q. Lin, J. W. Yu, and P. C. Zhang, "A band-notched absorber designed with high notch-band-edge selectivity," *IEEE Trans. Antennas Propag.*, vol. 65, no. 7, pp. 3560–3567, Jul. 2017.
- [131] A. A. Omar, H. Huang, and Z. Shen, "Absorptive frequency-selective reflection/transmission structures: A review and future perspectives," *IEEE Antennas Propag. Mag.*, vol. 62, no. 4, pp. 62–74, Aug. 2020.
- [132] Y. Yu, G. Q. Luo, A. A. Omar, X. Liu, W. Yu, Z. C. Hao, and Z. Shen, "3D absorptive frequency-selective reflection and transmission structures with dual absorption bands," *IEEE Access*, vol. 6, pp. 72 880–72 888, 2018.
- [133] B. Zhang, C. Jin, and Z. Shen, "Absorptive frequency selective reflector based on bent metallic strip embedded with chip-resistors," *IEEE Trans. Antennas Propag.*, vol. 68, no. 7, pp. 5736–5741, Jul. 2020.
- [134] H. Huang and Z. Shen, "An absorptive frequency-selective reflector based on miniaturized square-loop resonator," in *12th European Conference on Antennas and Propagation (EuCAP 2018)*, 2018, pp. 1–4.

- [135] Y. Han, L. Zhu, Y. Chang, and B. Li, "Dual-polarized bandpass and band-notched frequency-selective absorbers under multimode resonance," *IEEE Trans. Antennas Propag.*, vol. 66, no. 12, pp. 7449–7454, Dec. 2018.
- [136] A. A. Omar and Z. Shen, "Multiband high-order bandstop 3-D frequency-selective structures," *IEEE Trans. Antennas Propag.*, vol. 64, no. 6, pp. 2217–2226, Jun. 2016.
- [137] J. L. Wallace, "Broadband magnetic microwave absorbers: fundamental limitations," *IEEE Trans. Magnetics*, vol. 29, no. 6, pp. 4209–4214, Jun. 1993.
- [138] B. Zhang, G. Lu, Y. Feng, J. Xiong, and H. Lu, "Electromagnetic and microwave absorption properties of alnico powder composites," *J. Magn. Magnetic Mater.*, vol. 299, no. 1, pp. 205–210, 2006.
- [139] W. Li, T. Wu, W. Wang, P. Zhai, and J. Guan, "Broadband patterned magnetic microwave absorber," *J. Appl. Phys.*, vol. 116, no. 4, p. 044110, 2014.
- [140] B. Wadell, *Transmission Line Design Handbook*. Artech House, 1991.
- [141] I. J. Bahl, M. Bozzi, and R. Garg, *Microstrip Lines and Slotlines*, 3rd ed. Boston, USA: Artech House, 2013.
- [142] D. Pozar, *Microwave Engineering*, 4th ed. New Jersey, USA: Wiley, 2011.
- [143] N. Marcuvitz, *Waveguide Handbook*. New York, USA: McGraw-Hill, 1951.
- [144] S. S. Holland and M. N. Vouvakis, "The planar ultrawideband modular antenna (PUMA) array," *IEEE Trans. Antennas Propag.*, vol. 60, no. 1, pp. 130–140, Jan. 2012.
- [145] W. Jiang, Y. Liu, S. Gong, and T. Hong, "Application of bionics in antenna radar cross section reduction," *IEEE Antennas Wireless Propag. Lett.*, vol. 8, pp. 1275–1278, 2009.
- [146] M. Selvanayagam and G. V. Eleftheriades, "Experimental demonstration of active electromagnetic cloaking," *Phys. Rev. X*, vol. 3, p. 041011, Nov. 2013.
- [147] S. Varault, M. Soiron, A. Barka, A. C. Lepage, and X. Begaud, "RCS reduction with a dual polarized self-complementary connected array antenna," *IEEE Trans. Antennas Propag.*, vol. 65, no. 2, pp. 567–575, Feb. 2017.
- [148] H. Jiang, Z. Xue, Q. Zeng, W. Li, and W. Ren, "High-gain low-RCS slot antenna array based on checkerboard surface," *IET Microw. Antennas Propag.*, vol. 12, no. 2, pp. 237–240, 2018.
- [149] G. V. Trentini, "Partially reflecting sheet arrays," *IRE Trans. Antennas Propag.*, vol. 4, no. 4, pp. 666–671, Oct. 1956.

- [150] W. Pan, C. Huang, P. Chen, X. Ma, C. Hu, and X. Luo, "A low-RCS and high-gain partially reflecting surface antenna," *IEEE Trans. Antennas Propag.*, vol. 62, no. 2, pp. 945–949, Feb. 2014.
- [151] H. Jiang, Z. Xue, W. Li, W. Ren, and M. Cao, "Low-RCS high-gain partially reflecting surface antenna with metamaterial ground plane," *IEEE Trans. Antennas Propag.*, vol. 64, no. 9, pp. 4127–4132, Sep. 2016.
- [152] N. Misran, R. Cahill, and V. F. Fusco, "RCS reduction technique for reflectarray antennas," *Electron. Lett.*, vol. 39, no. 23, p. 1630, Nov. 2003.
- [153] L.-S. Ren, Y.-C. Jiao, J.-J. Zhao, and F. Li, "RCS reduction for a FSS-backed reflectarray using a ring element," *Prog. In Electromagn. Res. Lett.*, vol. 26, pp. 115–123, Jan. 2011.
- [154] S. Genovesi, F. Costa, and A. Monorchio, "Low-profile array with reduced radar cross section by using hybrid frequency selective surfaces," *IEEE Trans. Antennas Propag.*, vol. 60, no. 5, pp. 2327–2335, May 2012.
- [155] A. W. Love, *Reflector Antennas*. New York, USA: IEEE Press, 1978.
- [156] Y. Rahmat-Samii and R. Haupt, "Reflector antenna developments: A perspective on the past, present and future," *IEEE Antennas Propag. Mag.*, vol. 57, no. 2, pp. 85–95, Apr. 2015.
- [157] A. A. Omar and Z. Shen, "Multiband high-order bandstop 3-D frequency-selective structures," *IEEE Trans. Antennas Propag.*, vol. 64, no. 6, pp. 2217–2226, Jun. 2016.
- [158] J. Huang and J. Encinar, *Reflectarray Antennas*. NJ, USA: John Wiley & Sons, 2007.
- [159] D. M. Pozar, S. D. Targonski, and H. D. Syrigos, "Design of millimeter wave microstrip reflectarrays," *IEEE Trans. Antennas Propag.*, vol. 45, no. 2, pp. 287–296, Feb. 1997.
- [160] H. Rajagopalan, S. Xu, and Y. Rahmat-Samii, "On understanding the radiation mechanism of reflectarray antennas: An insightful and illustrative approach," *IEEE Antennas Propag. Mag.*, vol. 54, no. 5, pp. 14–38, Oct. 2012.
- [161] R. C. Hansen, *Phased Array Antennas*. NJ, USA: John Wiley & Sons, 2009.
- [162] J. Mitola and G. Q. Maguire, "Cognitive radio: making software radios more personal," *IEEE Pers. Commun.*, vol. 6, no. 4, pp. 13–18, Aug. 1999.
- [163] G. C. Tavik, C. L. Hilterbrick, J. B. Evins, J. J. Alter, J. G. Crnkovich, J. W. de Graaf, W. Habicht, G. P. Hrin, S. A. Lessin, D. C. Wu, and S. M. Hagewood, "The advanced multifunction RF concept," *IEEE Trans. Microw. Theory Tech.*, vol. 53, no. 3, pp. 1009–1020, Mar. 2005.

- [164] B. Munk, *Frequency Selective Surfaces: Theory and Design*. New York, USA: Wiley, 2005.
- [165] D. Sang, Q. Chen, L. Ding, M. Guo, and Y. Fu, "Design of checkerboard AMC structure for wideband RCS reduction," *IEEE Trans. Antennas Propag.*, vol. 67, no. 4, pp. 2604–2612, Apr. 2019.
- [166] Y. Liu, K. Li, Y. Jia, Y. Hao, S. Gong, and Y. J. Guo, "Wideband RCS reduction of a slot array antenna using polarization conversion metasurfaces," *IEEE Trans. Antennas Propag.*, vol. 64, no. 1, pp. 326–331, Jan. 2016.
- [167] T. Hong, S. Wang, Z. Liu, and S. Gong, "RCS reduction and gain enhancement for the circularly polarized array by polarization conversion metasurface coating," *IEEE Antennas Wireless Propag. Lett.*, vol. 18, no. 1, pp. 167–171, Jan. 2019.
- [168] M. Moccia, S. Liu, R. Y. Wu, G. Castaldi, A. Andreone, T. J. Cui, and V. Galdi, "Coding metasurfaces for diffuse scattering: Scaling laws, bounds, and suboptimal design," *Adv. Opt. Mater.*, vol. 5, no. 19, p. 1700455, 2017.
- [169] L. Zhang, X. Wan, S. Liu, J. Y. Yin, Q. Zhang, H. T. Wu, and T. J. Cui, "Realization of low scattering for a high-gain fabry-perot antenna using coding metasurface," *IEEE Trans. Antennas Propag.*, vol. 65, no. 7, pp. 3374–3383, Jul. 2017.
- [170] S. Xiao, S. Yang, Y. Chen, S. Qu, and J. Hu, "An ultra-wideband tightly coupled dipole array co-designed with low scattering characteristics," *IEEE Trans. Antennas Propag.*, vol. 67, no. 1, pp. 676–680, Jan. 2019.
- [171] P. Yang, F. Yan, F. Yang, and T. Dong, "Microstrip phased-array in-band RCS reduction with a random element rotation technique," *IEEE Trans. Antennas Propag.*, vol. 64, no. 6, pp. 2513–2518, Jun. 2016.
- [172] S. Xiao, S. Yang, H. Zhang, H. Bao, Y. Chen, and S. Qu, "A low-profile wide-band tightly coupled dipole array with reduced scattering using polarization conversion metamaterial," *IEEE Trans. Antennas Propag.*, vol. 67, no. 8, pp. 5353–5361, Aug. 2019.
- [173] Q. Chen, L. Chen, J. Bai, and Y. Fu, "Design of absorptive frequency selective surface with good transmission at high frequency," *Electron. Lett.*, vol. 51, no. 12, pp. 885–886, 2015.
- [174] K. Zhang, W. Jiang, and S. Gong, "Design bandpass frequency selective surface absorber using LC resonators," *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 2586–2589, 2017.
- [175] Datasheet of Eccosorb FGM series thin broadband microwave absorbers, Laird Tech., 2018. [Online]. Available: <https://www.laird.com/sites/default/files/2018-11/Datasheet%20Eccosorb%20FGM.pdf>