Realization of miniaturized multi-/wideband microwave front-ends

Shamaileh, Khair Al

University of Toledo
A Dissertation
titled
Realization of Miniaturized Multi-/Wideband Microwave Front-Ends
by
Khair A. Al Shamaileh
Submitted to the Graduate Faculty as partial fulfillment of the requirements for the
Doctor of Philosophy Degree in Engineering

Vijay Devabhaktuni, Ph.D., Committee Chair

Mansoor Alam, Ph.D., Committee Member

Junghwan Kim, Ph.D., Committee Member

Daniel Georgiev, Ph.D., Committee Member

Mohammad Almalkawi, Ph.D., Committee Member

Douglas Nims, Ph.D., Committee Member

Abdelrazik Sebak, Ph.D., Committee Member

Patricia Komuniecki, Ph.D., Dean
College of Graduate Studies

The University of Toledo

August 2015
An Abstract of

Realization of Miniaturized Multi-/Wideband Microwave Front-Ends

by

Khair Al Shamaileh

Submitted to the Graduate Faculty as partial fulfillment of the requirements for the
Doctor of Philosophy Degree in Engineering

The University of Toledo

August 2015

The ever-growing demand toward designing microwave front-end components
with enhanced access to the radio spectrum (e.g., multi-/wideband functionality) and
improved physical features (e.g., miniaturized circuitry, ease and cost of fabrication) is
becoming more paramount than ever before. This dissertation proposes new design
methodologies, simulations, and experimental validations of passive front-ends (i.e.,
antennas, couplers, dividers) at microwave frequencies. The presented design concepts
optimize both electrical and physical characteristics without degrading the intended
performance. The developed designs are essential to the upcoming wireless technologies.

The first proposed component is a compact ultra-wideband (UWB) Wilkinson
power divider (WPD). The design procedure is accomplished by replacing the uniform
transmission lines in each arm of the conventional single-frequency divider with
impedance-varying profiles governed by a truncated Fourier series. While such non-
uniform transmission lines (NTLs) are obtained through the even-mode analysis, three
isolation resistors are optimized in the odd-mode circuit to achieve proper isolation and
output ports matching over the frequency range of interest. The proposed design
methodology is systematic, and results in single-layered and compact structures.
For verification purposes, an equal split WPD is designed, simulated, and measured. The obtained results show that the input and output ports matching as well as the isolation between the output ports are below –10 dB; whereas the transmission parameters vary between –3.2 dB and –5 dB across the 3.1–10.6 GHz band. The designed divider is expected to find applications in UWB antenna diversity, multiple-input-multiple-output (MIMO) schemes, and antenna arrays feeding networks.

The second proposed component is a wideband multi-way Bagley power divider (BPD). Wideband functionality is achieved by replacing the single-frequency matching uniform microstrip lines in the conventional design with NTLs of wideband matching nature. To bring this concept into practice, the equivalent transmission line model is used for profiling impedance variations. The proposed technique leads to flexible spectrum allocation and matching level. Moreover, the resulting structures are compact and planar. First, the analytical results of three 3-way BPDs of different fractional bandwidths are presented and discussed to validate the proposed approach. Then, two examples of 3- and 5-way BPDs with bandwidths of 4–10 GHz and 5–9 GHz, respectively, are simulated, fabricated, and measured. Simulated and measured results show an acceptable input port matching of below –15 dB and –12.5 dB for the 3- and 5-way dividers, respectively, over the bands of interest. The resulting transmission parameters of the 3- and 5-way dividers are –4.77±1 dB and –7±1 dB, respectively, over the design bands; which are in close proximity to their theoretical values. The proposed wideband BPD dividers find many applications in microwave front-end circuitry, especially in only-transmitting antenna subsystems, such as multi-/broad-cast communications, where neither output ports matching nor isolation is a necessity.
The third proposed component is a 90° hybrid branch-line coupler (BLC) with multi-/wideband frequency matching. To obtain a multi-frequency operation, NTLs of lengths equal to those in the conventional design are incorporated through the even- and odd-mode analysis. The proposed structure is relatively simple and is fabricated on a single-layered substrate. Two design examples of dual-/triple-frequency BLCs suitable for GSM, WLAN, and Wi-Fi applications are designed, fabricated and evaluated experimentally to validate the proposed methodology. The same concept is extended to realize a broadband BLC with arbitrary coupling levels. Based on how impedances are profiled, the fractional bandwidth of a single-section 90° 3-dB BLC is extended to 57%, and the realization of broadband BLCs with predefined coupling levels is also achieved. Furthermore, higher-order harmonics are suppressed by enforcing BLC S-parameters to match design requirements only at a given frequency range. Three examples of 3-dB, 6-dB, and 9-dB BLCs are demonstrated at 3 GHz center frequency. The obtained analytical response, EM simulations, and measurements justify the design concept.

The fourth proposed component is an UWB antipodal Vivaldi antenna (AVA) with high-Q stopband characteristics based on compact electromagnetic bandgap (EBG) structures. First, an AVA is designed and optimized to operate over an UWB spectrum. Then, two pairs of EBG cells are introduced along the antenna feed-line to suppress the frequency components at 3.6–3.9 and 5.6–5.8 GHz (i.e., WiMAX and ISM bands, respectively). Simulated and measured voltage standing wave ratio (VSWR) are below 2 for the entire 3.1–10.6 GHz band with high attenuation at the two selected sub-bands. This simple yet effective approach eliminates the need to deform the antenna radiators with slots/parasitic elements or comprise multilayer substrates.
For my parents and my wife Rand
Acknowledgements

I would like to express my sincere appreciation to Prof. Vijay Devabhaktuni for his supportive recommendations that led this research to a successful conclusion.

I am very grateful to Dr. Mohammad Almalkawi and Dr. Nihad Dib for their expert assistance throughout the course of this research.

I would also like to thank Dr. Amin Abbosh, Dr. Saeed Abushamleh, and Dr. Lee Cross for their help in fabrications and measurements.

And most of all, I am grateful to my wife Rand for her support through all of this.

I thank you all for I could not have done it without each one of you.
# Table of Contents

Abstract .............................................................................................................................. iii
Acknowledgements........................................................................................................... vii
Table of Contents ............................................................................................................. viii
List of Tables ..................................................................................................................... xi
List of Figures ................................................................................................................... xii
List of Abbreviations ...................................................................................................... xvii

1 Introduction .................................................................................................................. 1
  1.1 Motivation .................................................................................................................. 1
  1.2 Research Objective .................................................................................................... 2
  1.3 Literature Survey ...................................................................................................... 3
    1.3.1 The Wilkinson Power Divider ............................................................................ 3
    1.3.2 The Bagley Power Divider ................................................................................ 4
    1.3.3 The Quadrature Branch-line Coupler ................................................................. 5
    1.3.4 The Antipodal Vivaldi Antenna ........................................................................... 8
  1.4 Organization .............................................................................................................. 9

2 Non-Uniform Microstrip Transmission Lines .......................................................... 11
  2.1 Non-Uniform Transmission Line Optimization .................................................... 12
  2.2 Non-Uniform Transmission Line Modeling .......................................................... 15
  2.3 Results and Discussions ........................................................................................ 16
2.4 Conclusions........................................................................................................21

3 Ultra-Wideband Wilkinson Power Divider ..................................................22

3.1 Design.............................................................................................................23

3.1.1 Even-Mode Analysis .............................................................................24

3.1.2 Odd-Mode Analysis ..............................................................................25

3.2 Simulations and Measurements ..................................................................27

3.3 Non-Uniform Ultra-Wideband Divider Modeling ......................................29

3.4 Conclusions...................................................................................................33

4 Wideband Multi-Way Bagley Power Divider ..............................................34

4.1 Design.............................................................................................................35

4.2 Analytical Examples ....................................................................................38

4.3 Simulations and Measurements ....................................................................40

4.4 Conclusions...................................................................................................45

5 Multi-/broadband Quadrature Branch-Line Coupler .......................................46

5.1 Multi-Frequency Branch-Line Coupler ......................................................47

5.1.1 Dual-Frequency Example ......................................................................51

5.1.2 Triple-Frequency Example ....................................................................53

5.2 Broadband Branch-Line Coupler ................................................................55

5.2.1 Design .......................................................................................................55

5.2.2 Analytical Results ...................................................................................60

5.2.3 Simulations and Measurements .............................................................66

5.3 Conclusions...................................................................................................73

6 Dual-Band Notch Antipodal Vivaldi Antenna ................................................75
6.1 Antenna Configuration.................................................................76
6.2 Performance Analysis .................................................................77
6.3 Simulations and Measurements ....................................................81
6.4 Conclusions..................................................................................84

7 Conclusions and Future Work ..........................................................85
7.1 Summary .....................................................................................85
7.2 Impact on Different Disciplines ....................................................89
    7.2.1 Global EARS Community .....................................................89
    7.2.2 Academia, Society, and Industry ...........................................90
7.3 Future Work ................................................................................91
7.4 Research Publications and Outcomes .........................................95

References .......................................................................................97
List of Tables

2.1 Comparison between optimized and ANN-based non-uniform transformers. .....21
3.1 Comparison between optimized and modeled WPDs.................................33
4.1 Optimized Fourier series coefficients for the three 3-way BPD examples.........39
4.2 Measured metrics of the proposed dividers magnitude/phase imbalances. ....42
5.1 NTL coefficients of the dual-band BLC......................................................51
5.2 NTL coefficients of the triple-band BLC.....................................................53
5.3 Theoretical values of the through and coupled parameters. ................. 62
5.4 Fourier coefficients of the impedances of the three couplers. ............ 65
5.5 Comparison between electrical and physical characteristics of recent broadband branch-line couplers.................................................................72
List of Figures

2-1 (a) conventional microstrip line; (b) proposed miniaturized NTL.................13

2-2 ANN model of a NTL transformer trained with backpropagation, quasi-Newton, and conjugate gradient techniques. .................................................................16

2-3 $ABCD$ parameters comparison between the conventional uniform transformer; compact optimized NTL transformer; and the ANN-modeled NTL transformer:
(a) parameter $A$; (b) parameter $B$; (c) parameter $C$; (d) parameter $D$...............17

2-4 Optimized and ANN-based NTL transformers variations as a function of length:
(a) width $W(x)$; (b) impedance $Z(x)$. .................................................................19

2-5 Optimized and ANN-based NTL transformers $S$-parameters: (a) $|S_{11}|$; (b) $|S_{21}|$ ....19

2-6 Simulations of the optimized and ANN QN-based NTL transformers..............20

3-1 Schematic diagrams of (a) conventional single-frequency WPD; (b) proposed UWB WPD utilizing NTLs.................................................................23

3-2 Proposed non-uniform WPD: (a) even-mode; (b) odd-mode circuits. ..........23

3-3 Flowchart demonstrating the design of the proposed UWB divider; green and red enclosures present the even- and odd-mode analyses, respectively. .................26

3-4 Simulated and measured $S$-parameters of the proposed UWB divider...........28

3-5 (a) measured amplitude and phase imbalance of the proposed UWB NTL divider;
(b) simulated and measured group delay. .............................................................28

3-6 Proposed ANN model of the UWB non-uniform WPD. ...............................29
3-7 Optimized and ANN-based non-uniform WPD arm variations as a function of length: (a) width; (b) impedance .................................................................30

3-8 Calculated $S$-parameters of the UWB WPD for optimized and modeled resistors of $\{R_1,R_2,R_3\} = \{151,237.6,147.4\}$ and $\{156.6,252.8,148.8\}$, respectively: (a) $|S_{11}|$; (b) $|S_{21}|$; (c) $|S_{22}|$; (d) $|S_{23}|$. .................................................................31

3-9 Full-wave EM simulations of the optimized and ANN-based UWB WPD: (a) $|S_{11}|$, (b) $|S_{21}|$, (c) $|S_{22}|$, and (d) $|S_{23}|$. ..............................................................................32

4-1 (a) proposed wideband multi-way impedance-varying BPD; (b) equivalent transmission line model. .................................................................36

4-2 Flowchart showing the design of the proposed wideband BPD; red enclosure presents formulations based on the equivalent transmission line model. .........37

4-3 NTL transformer designs for the three different proposed fractional bandwidths: (a) impedance variations; (b) width variations. ........................................39

4-4 $S$-parameters for three fractional bandwidths: (a) $|S_{11}|$; (b) $|S_{21}|$. ......................40

4-5 Photographs of the fabricated BPD structures: (a) 3-way; (b) 5-way. .................40

4-6 Simulated and measured $S$-parameters of the proposed 3-way NTL BPD: (a) $|S_{11}|$; (b) $|S_{21}|$; (c) $|S_{31}|$. ..............................................................................41

4-7 Simulated and measured $S$-parameters of the proposed 5-way NTL BPD: (a) $|S_{11}|$; (b) $|S_{21}|$; (c) $|S_{31}|$; (d) $|S_{41}|$. ..............................................................................41

4-8 BPDs simulated and measured group delays: (a) 3-way; (b) 5-way. ...................42

4-9 Measured imbalance of the 3-way BPD: (a) magnitude; (b) phase. .....................43

4-10 Measured imbalance of the 5-way BPD: (a) magnitude; (b) phase. .....................43

4-11 Output ports isolation of the 3-way BPD: (a) $|S_{23}| = |S_{34}|$; (b) $|S_{24}|$. ...............44
Output ports matching of the 3-way BPD: (a) $|S_{22}| = |S_{44}|$; (b) $|S_{33}|$. ........................................44

Schematics of: (a) conventional single-frequency BLC; (b) proposed multi-frequency BLC utilizing NTLs. ........................................................................................................................................47

Proposed non-uniform BLC circuits: (a) even-mode; (b) odd-mode. ......................47

Flowchart showing the design procedure of the multi-frequency non-uniform BLC; green and red enclosures present the theoretical formulation based on even- and odd-mode equivalent transmission line circuits, respectively. ..............................50

Simulated and measured results of the dual-frequency BLC: (a) $S$-parameters magnitude; (b) phase difference between $S_{21}$ and $S_{31}$. .................................................................52

Simulated and measured results of the triple-frequency BLC: (a) $S$-parameters magnitude; (b) phase difference between $S_{21}$ and $S_{31}$. .................................................................54

Schematic diagram of the proposed broadband BLC. The dashed blue box represents the portion where the even-odd mode analysis is carried out. ......................55

Even-odd mode circuit outlines of the proposed impedance-varying broadband BLC: (a) even-even; (b) even-odd; (c) odd-even; (d) odd-odd. ............................55

Pseudocode of the proposed broadband impedance-varying BLC. .......................59

Variations as a function of length: (a) 3-dB; (b) 6-dB; (c) 9-dB broadband BLCs. Solid, dotted, and dashed lines represent $Z_1(x)$, $Z_2(x)$, and $Z_3(x)$, respectively. .....61

Analytical response of the proposed broadband BLCs with different values of $C$. Magnitudes of $S$-parameters for: (a) $C = 3$-dB; (b) $C = 6$-dB; (c) $C = 9$-dB. Phase difference between the through and coupled ports for: (d) $C = 3$-dB; (e) $C = 6$-dB; (f) $C = 9$-dB. .........................................................................................................................62
5-11 Response of a broadband 6-dB BLC over an extended frequency range. S-parameter magnitudes: (a) design equations in [49]; (b) the proposed method. Phase differences between through and coupled ports: (c) design equations reported in [49]; (d) the proposed method.

5-12 Photographs of the fabricated BLCs: (a) 3-dB; (b) 6-dB; (c) 9-dB.

5-13 Magnitude response of: (a) 3-dB; (b) 6-dB; (c) 9-dB BLCs. Dashed, dotted, solid, and dashed-dotted lines represent the simulated $S_{21}$, $S_{31}$, $S_{11}$, and $S_{41}$, respectively; whereas the plus, star, circle, and cross markers represent the measured $S_{21}$, $S_{31}$, $S_{11}$, and $S_{41}$, respectively.

5-14 Simulated and measured phase difference between the through and coupled ports: (a) $C = 3$-dB; (b) $C = 6$-dB; (c) $C = 9$-dB.

5-15 S-parameter magnitude of impedance-varying broadband 6-dB BLCs optimized for three different fractional bandwidths.

5-16 Phase differences between through and coupled ports of the impedance-varying broadband 6-dB BLCs optimized for three different fractional bandwidths.

6-1 Proposed dual-band notched AVA; black and gray strips refer to upper and lower flares, respectively.

6-2 Notch characteristics for pair and single EBG cells.

6-3 Effect of changing EBG$_1$ (a) radius $r_1$; (b) width $w_{m1}$; (c) separation $s_l$.

6-4 VSWR simulations for four different $d_s$ values.

6-5 VSWR simulation results for four different scenarios.

6-6 Current distribution of the proposed dual-notch AVA at frequencies: (a) 3.8 GHz; and (b) 5.7 GHz.
6-7 Simulated and measured VSWRs of the proposed AVA.................................81
6-8 Maximum gain for conventional and proposed AVAs.................................82
6-9 Proposed dual-notched AVA radiation patterns: (a) 5 GHz, (b) 7 GHz, (c) 9 GHz,
(d) 3.8 GHz; center frequency of the 1st notch, and (e) 5.7 GHz; center frequency
of the 2nd notch. ........................................................................................................83
6-10 Group delay of the proposed dual-band notched AVA antenna..................83
List of Abbreviations

ANN........................... artificial neural network
AVA........................... antipodal Vivaldi antenna

BLC.........................branch-line coupler
BP.........................backpropagation
BPD.........................Bagley power divider

CPW..........................coplanar waveguide
CG..........................conjugate gradient

EARS........................enhancing access to the radio spectrum
EBG........................electromagnetic bandgap
EM........................electromagnetic

FB..........................fractional bandwidth

GSM........................global system for mobile communications

HFSS........................high frequency structural simulator

IEEE........................institute of electrical and electronics engineers
IPD........................integrated passive device
ISM........................industrial-scientific-medical

m/s..........................meter per second
MIMO........................multiple-input-multiple-output
MLP........................multi-layer perceptron
MMIC........................monolithic microwave integrated circuit
mm-wave......................millimeter wave

NTL..........................non-uniform transmission line

PCB..........................printed circuit board

QN..........................quasi Newton
RF...............................radio frequency
SMA...........................subminiature version A
S-parameter ...............scattering parameter
STEM..........................science, technology, engineering, and math
UWB .........................ultra-wideband, uses 3.1–10.6 GHz frequency range
VNA..........................vector network analyzer
VSWR.........................voltage standing wave ratio
Wi-Fi.........................wireless fidelity, a technology based on IEEE 802.11 standard
WLAN........................wireless local area network
WPĐ..........................Wilkinson power divider
Chapter 1

Introduction

Front-end components are of an essence to any microwave subsystem, such as transceiver modules, medical instruments, and imaging devices. Hence, a tremendous effort is relentlessly placed to enhance their electrical performance while maintaining a compact size, reasonable fabrication complexity, and – above all else – cost. This dissertation presents novel designs of microwave front-ends that address multi-/wideband performance consistency concerns and optimize realization ease and cost.

In this chapter, the motivations of this investigation are discussed in Section 1.1; dissertation objectives are listed in Section 1.2; a literature survey on the covered front-end components in this research is provided in Section 1.3; followed by a brief overview of each chapter in Section 1.4.

1.1 Motivation

The microwave frequency range, loosely defined as 0.3–30 GHz, is a portion of the electromagnetic (EM) spectrum commonly used for wireless communications, audio and video broadcast, radars, imaging, and sensors. With the continuous development in such arenas, the need for front-end components of advanced electrical properties and
improved physical characteristics is of utmost significance. Microwave components are now expected to support concurrent applications by switching from a single-frequency to multi-/wideband functionalities. Such an interchange must come at no expense to neither the resulting circuitry occupation nor design complexity and cost. At the same time, improved bandwidth utilizations must not impact other coexisting technologies.

Motivated by these challenges, this dissertation seeks to provide novel concepts to fulfill the requirements mentioned above, with an emphasis on the following widely exploited front-end RF/microwave components: Wilkinson and Bagley power dividers, quadrature branch-line couplers, and antipodal Vivaldi antenna. Though, the developed approaches and design methodologies are valid for a variety of other front-ends.

1.2 Research Objective

The main goal of this dissertation is to design front-end microwave components with an improved frequency response and bandwidth accessibility. The development of such components must rely on systematic platforms that are tunable to the given design requirements. Furthermore, the realization of the resulting schematics is considered as a point of concern, by avoiding any increase in the structural complexity, size as well as manufacturing ease and cost.

In order to bring such objectives into reality, mathematical representations of all addressed components are derived based on microwave engineering and transmission line theory. The developed foundations are analytically tested and justified by means of professional full-wave EM simulations. The realization of the proposed designs are performed by means of fabrications and measurements. Finally, simulated and measured outcomes are compared to judge the validity of the proposed structures.
1.3 Literature Survey

Power dividers, couplers, and antennas are integral components in many front-end RF/microwave systems. Hence, the advanced designs and miniaturization of these components are ongoing research topics. Scholars strive to achieve set of targets (e.g., broadened bandwidth, suppressed harmonics) while minimizing size and fabrication cost. This section presents a literature survey on the components proposed in this dissertation. Section 1.3.1 presents the progress in UWB WPD design, and describes the recent reported methods that obtained this performance. The BPD is introduced in Section 1.3.2, adjoined with the latest research associated with this component. The quadrature BLC is investigated in Section 1.3.3, in which contributions to multi-/wideband designs are reported. Finally the AVA is introduced in Section 1.3.4, where Different band-notch techniques are presented and thoroughly discussed.

1.3.1 The Wilkinson Power Divider

The WPD, invented by E. Wilkinson [1], is a passive component that gained much interest in literature, due to its capacity in achieving high isolation between the output ports while maintaining a matched condition at all ports. These significant properties qualify its adoption in arrays feeding networks, and MIMO applications. Due to the fact that conventional WPDs support only a single frequency, their exploitation to wideband systems are limited. For example, the conventional WPD is incompatible with the widely utilized UWB spectrum that spans the 3.1–10.6 GHz frequency range; and thus, is unemployable to technologies that use this spectrum. In [2], a reduced-size UWB divider was proposed by implementing the transmission lines of a two-stage WPD using
bridged T-coils. However, the accompanied complexity in the design and fabrication is a major drawback. Bialkowski et al. proposed a compact UWB out-of-phase uniplanar power divider formed by a slotline and a microstrip line T-junction along with wideband microstrip to slotline transitions [3]. A miniaturized three-way power divider with UWB feature was presented in [4] by utilizing broadside coupling via multilayer microstrip/slot transitions of elliptical shape. A very similar approach was utilized in [5] to design a planar in-phase power divider via circular microstrip/slot transitions for 2–5 GHz wideband applications. Tapered line transformers, which exhibit almost a constant input impedance over a wide range of frequencies, were incorporated in the design of an UWB divider [6]. Nevertheless, the resulting circuit area was relatively large. Different kinds of stubs, such as open stubs [7], delta stubs [8], radial stubs [9], and coupled lines [10] were introduced as an approach in designing modified WPDs with extended bandwidth. As such, extra transmission lines were utilized. Other efforts enhanced spectrum accessibility of the WPD by proposing multi-frequency topologies based on lumped elements [11]-[13] and stubs [14]-[15]. However, the increased integration complexity and circuitry occupation were major disadvantages.

1.3.2 The Bagley Power Divider

Unlike the WPD, the output ports of the BPD can be easily extended to any number according the application requirements. BPDs also offer structural compactness, excellent input port matching and transmission, and a planar geometry without any added design complexity or lumped elements (e.g., resistors, inductors, capacitors). However, the operational bandwidth of the conventional BPD does not support wideband-based communication schemes. In this regard, this divider requires modifications to flexibly
cover a wideband design spectrum with the required matching level. In [16], reduced size 3- and 5-way BPDs using open stubs were presented. An optimum design of a modified 3-way Bagley rectangular power divider was investigated in [17]. However, the half-wavelength impedances in the conventional design result in a considerable increase in BPD circuitry. To this end, the conventional BPD design was redefined in [18]-[19] by eliminating the half-wavelength arbitrarily chosen impedances connecting the output ports with specific impedance values of unconditional lengths.

In order to improve the BPD bandwidth, composite right-/left-handed (CRLH) transmission lines [20], dual-passband sections [21], two-section quarter-wavelength transformers [22], dual-band matching networks [23], and coupled lines [24] were investigated to achieve dual-frequency functionality. Compact multi-band multi-way BPDs utilizing NTLs were proposed in [25]. In [26], a generalized design procedure for an unequal split multi-way BPD was elaborated. It is noteworthy to point out that BPDs investigated in [16]-[26] have an odd number of output ports; whereas novel BPDs with an even number of output ports was introduced in [27].

1.3.3 The Quadrature Branch-line Coupler

The 90° BLC is found in many modern systems, such as measurement setups, radars, and RF mixers, where reduced-size circuitry and multi-/wideband operation are two main objectives. However, the conventional BLC suffers from the inherent single-frequency matching nature due to the narrowband properties of the quarter-wavelength transformers that form its branches. Hence, introducing systematic and realistic multi-/wideband methodologies that support the current and simultaneous wireless technologies are steering the research momentum in the most recent BLC studies.
Normally, dual-frequency characteristics are achieved through the use of dual-band quarter-wavelength impedance transformers [28] attained by the proper selection of circuit parameters. Nevertheless, the increase in the circuitry size was a major concern. Other ways to realize the dual-band characteristic of BLCs were either by: a) using unequal arms lengths adjoined with a center-tapped stub [29]; b) incorporating stepped-impedance-stub placed at the middle of each quarter-wave branch of the conventional coupler [30]; c) using four open-ended quarter-wave transmission lines at each port of the BLC [31], where the lengths of the additional stubs as well as the main branches, are evaluated at the middle frequency of the two operating bands; or d) employing coupled-line sections as demonstrated in [32].

Feng Lin et al. proposed a tri-band BLC with three controllable operating frequencies employing four matching stubs at each port [33]. A similar technique stands in [34], in which triple-broadband matching techniques employing matching stubs were considered in designing a 3-dB BLC. A tri-band BLC using double-Lorentz transmission lines was introduced in [35], where lumped capacitors and inductors were incorporated in the middle of each branch. The design of a tri-band coupler for WiMAX applications was investigated in [36]. However, the lack of detailed design procedure and analysis makes the design of such multi-band couplers a difficult task. Recently, triple- and quad-band 3-dB couplers were proposed by adopting optimized compensation techniques to satisfy the matching conditions [37]-[38]. It is paramount to point out that multi-band couplers reported in [28]-[34] and [36]-[38] were realized by adding extra transmission lines and/or matching stubs, which remarkably increases the overall circuit size; while in [35], lumped elements were incorporated in the BLC design topology.
A considerable effort was also devoted to broaden the operational bandwidth of BLCs with equal [39]-[47] as well as arbitrary [48]-[50] coupling levels, all of which showed excellent performance. In general, increasing the fractional bandwidth of a BLC while maintaining an equal coupling level (i.e., 3-dB) is achieved by incorporating matching networks at each of its ports, such as short- and open-circuited stubs [39]-[41], double quarter-wave transformers [42], and open-circuited coupled lines [43]. Coplanar waveguide (CPW) structures [44]-[45], and stub-loaded air-filled rectangular coaxial lines [46] were adopted to improve the bandwidth of the 3-dB BLC. The concept of integrated passive device (IPD) technology was recently introduced in [47] to increase the bandwidth of a BLC in the mm-wave range. The methods mentioned in [39]-[47] reported a fractional bandwidth (FB) of 50%. Nevertheless, some investigations resulted in increased circuit area, fabrication complexity, and cost. Similar approaches were used to design BLCs with arbitrary coupling levels, such as double [48] and single [49] quarter-wave sections, and CPW open-circuited series stubs [50]. However, the associated fabrication challenges at low coupling levels were serious disadvantages. Other approaches showed that the BLC bandwidth can be further enhanced by integrating multiple couplers in a cascaded manner [51]-[53]. Nevertheless, the subsequent increase in size and internal impedance levels were limitations that necessitate miniaturization and impedance compensation techniques, such as fractal shapes [54]-[55], stubs [56]-[60], coupled lines [61], and defected ground structures [62]-[64].

In short, new BLC designs with an emphasis on physical compactness, ease/cost of fabrication, multi-/broadband frequency response, and arbitrary coupling levels are seriously required in these days’ sophisticated RF subsystems.
1.3.4 The Antipodal Vivaldi Antenna

The AVA is investigated because of its merit in the field of UWB systems, owing to its UWB frequency characteristics, high gain, and directive radiation. Because of these desirable features, AVA proved to be a competitive candidate for several applications. However, the UWB characteristics of AVAs may cause interference to other coexisting wireless technologies; and thus, negatively affect their intended performance.

Although planar antennas with stopband characteristics have been extensively investigated [65]-[69], little efforts have been done to mitigate interference between AVAs and other wireless channel users, basically due to the non-uniformity of the AVA radiators which follows either an exponential or elliptical taper. In addition, the relatively low current distribution of such an antenna makes it difficult for antenna designers to drive the structure toward exhibiting frequency notch(s). Recent research articles have reported several band-reject resonators to alleviate the interference impact mainly by one of the following techniques: 1) etched slots on the antenna radiators [70]-[71]; and 2) parasitic elements on the radiation surface [72]. In [70], Ω-shaped slot was made in each radiating flare to create a frequency notch at 5.5 GHz. Following a similar approach, [71] proposed an AVA with a U-shaped slot to realize a band-notch at the same frequency. This technique was also applied to different antennas in order to obtain notches at predefined stopbands [65]-[66]. It is noteworthy mentioning that apertures excavating techniques reported in [70]-[71] require extensive parametric simulations and may degrade antenna gain due to copper etching [73]. The concept of applying split ring resonators as parasitic elements on the radiation surface was applied in [72] to create a notch within the 5–6 GHz band. A similar concept was applied in [67] taking into
account a different antenna structure, in which a triple-notched UWB antenna was designed using three parasitic strips accompanied with a deformed ground plane. Besides the increased computational demand, approaches in [67]-[72] negatively impact the overall design complexity. Other efforts incorporated multilayers stacked together to achieve multiple frequency notches. In [68]-[69], stopband resonators in multilayer configurations were introduced. However, the increase in fabrication cost and structural complexity/assembly were major drawbacks. Thus, this research focuses on convenient-to-realize band-notched AVAs with controllable notches’ number/locations.

1.4 Organization

Chapter 2 presents the theoretical platform and mathematical formulation of microstrip NTLs, which are the main applied technique throughout this dissertation. Then, the analytical results of a proof-of-principle example of an NTL transformer is presented. Such results are obtained using two different approaches: 1) optimizations; and 2) modeling to show the merit of impedance-varying lines as competitive candidates not only in achieving a certain electrical performance, but also in miniaturizing the overall circuitry of the entire design. Trust-region-reflective algorithm as well as artificial neural networks (ANNs) are utilized as optimization and modeling tools, respectively.

Chapter 3 presents the applications of NTLs in the design and realization of a compact and planar UWB equal-split WPD. First, the conventional and proposed power dividers are presented. Then, a theoretical approach of how NTLs are incorporated to obtain the required frequency response with no added fabrication limitations, complexity, or cost are emphasized. Further validations are given through simulated and measured results, in which both are elaborated and compared to verify the design concept.
Chapter 4 presents the utilization of NTLs in the design of wideband multi-way BPDs. Mathematical derivations of a wideband multi-way divider is firstly provided. Then, different examples of 3- and 5-way BPDs with different fractional bandwidths and matching levels are presented. The theoretical results of the proposed design approach are further validated through means of full-wave EM simulations, fabrications, and testing of two 3- and 5-way divider prototypes.

Chapter 5 presents NTLs multi-frequency and broadband quadrature BLCs. The design and realization of these couplers are established to be suitable for either multi-frequency or broadband applications. The first part of this chapter is to develop a systematic design accompanied with supporting simulations and measurements for dual- and triple-frequency BLCs. More advanced BLC characteristics (i.e., broadband response, arbitrary coupling levels, and higher harmonics suppression) are elaborated and thoroughly discussed in the second part.

Chapter 6 presents a simple but effective way to design an UWB AVA with dual-band notch. The design concept based on mushroom-like EBG structures, is given. The effects of the EBG dimensions on the overall response is studied by performing extensive parametric analysis. The number, positions, and level of notches are set as benchmarks. Then, the effect of utilizing the underlined principle on the antenna gain and radiation is presented. It will be shown that such a technique does not result in an increased antenna dimensions or extra incorporated substrates/layers.

Chapter 7 concludes this dissertation and suggests future research that is aligned with the scope of this investigation. It also summarizes the scientific contribution of this study and lists the resulting publications.
Chapter 2

Non-Uniform Microstrip Transmission Lines

Microstrip transmission line technology, developed by ITT laboratories as a competitor to its counterpart stripline [74], is a transmission medium for electromagnetic waves, and is fabricated using the conventional printed circuit board (PCB) photo-etching process. It consists of a conducting strip separated from the ground plane by a dielectric layer with a predefined permittivity known as the substrate.

Microstrip lines are extensively utilized in microwave components, such as power dividers, couplers, filters, and antennas. If adopted, the entire component is designed and fabricated based on a specific metallization pattern built on the substrate. Microstrip transmission lines are also widely utilized in monolithic microwave integrated circuits (MMICs) and high-speed digital PCB interconnects, where signals need to be routed from one part of the assembly to another with minimal distortion.

Microstrip technology is considerably less expensive than other transmission media, such as waveguides. Besides, designs built with this technology are significantly lighter, compact, and easier to realize. Hence, microstrip lines are exploited in almost all modern applications. However, such advantages come at the expense of lower power handling capacity, higher losses, and increased susceptibility to cross-talk.
This chapter presents a new representation of conventional microstrip lines. Instead of being constant, the proposed methodology suggests continuously varying the impedance along the propagation direction of the electromagnetic wave. These variations lead, by basic definition, to a non-uniform width profile. Based on how the impedance is varied, advanced physical and electrical features are obtained as compared to uniform microstrip lines. For example, NTLs can be designed to be more compact, and can have customized electrical performances set as goals during the profiling phase. At the same time, planarity, structural complexity, and fabrication cost are left unaltered.

This chapter is summarized as follows: Section 2.1 discusses the proposed theory and the design concept of compact microstrip NTLs. Then, Section 2.2 presents the incorporated optimization and modeling approaches in realizing the mathematical foundation. Section 2.3 elaborates the analytical results of a simple matching transformer example to validate the underlined concept. Finally, conclusions are given in Section 2.4.

2.1 Non-Uniform Transmission Line Optimization

Figure 2-1 shows a schematic diagram of a conventional microstrip transmission line with a fixed impedance $Z$, propagation constant $\beta$, and length $d_0$, along with its counterpart compact NTL with varying characteristic impedance $Z(x)$, propagation constant $\beta(x)$, and length $d < d_0$. The analysis starts by obtaining the $ABCD$ parameters of the NTL by subdividing it into $K$ uniform short segments each of length $\Delta x \ll \lambda$, where $\Delta x = d/K$, and $\lambda$ is the guided wavelength. The $ABCD$ matrix of the whole NTL is obtained by multiplying the $ABCD$ matrices of all sections as follows [75]:

$$
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_{Z(x)} = \begin{bmatrix}
A_i & B_i \\
C_i & D_i
\end{bmatrix} \cdots \begin{bmatrix}
A_i & B_i \\
C_i & D_i
\end{bmatrix} \cdots \begin{bmatrix}
A_K & B_K \\
C_K & D_K
\end{bmatrix}
$$

(2.1)
where the \(ABCD\) parameters of the \(i^{th}\) \((i = 1, 2 \ldots K)\) segment are [75]:

\[
A_i = D_i = \cos(\Delta \theta) \tag{2.2}
\]

\[
B_i = Z^2 \left((i-0.5)\Delta x\right) C_i = jZ \left((i-0.5)\Delta x\right) \sin(\Delta \theta) \tag{2.3}
\]

\[
\Delta \theta = \frac{2\pi}{\lambda} \Delta x = \frac{2\pi}{c} f \sqrt{\varepsilon_{\text{eff}}} \Delta x \tag{2.4}
\]

where \(c \approx 3\times10^8\) m/s is the speed of light and \(f\) is the design frequency. The effective dielectric constant, \(\varepsilon_{\text{eff}}\), of each section is calculated using the well-known microstrip line formulas given in [75]. Then, the following normalized non-uniform profile of \(Z(x)\), written in terms of a truncated Fourier series, is considered:

\[
Z(x) = Z \times \exp \left( c_0 + \sum_{n=0}^{N} a_n \cos \left( \frac{2\pi nx}{d} \right) + b_n \sin \left( \frac{2\pi nx}{d} \right) \right) \tag{2.5}
\]

In simple design examples (e.g., two port matching network), the series described in (2.5) can be reduced to the following representation:

\[
Z(x) = Z \times \exp \left( \sum_{n=0}^{N} a_n \cos \left( \frac{2\pi nx}{d} \right) \right) \tag{2.6}
\]
where \( N \) represents the number of series terms. Downgrading the series terms from (2.5) to (2.6) results in valued design benefits, such as reduced optimization time and physical NTL symmetry around the propagation direction. However, better performance is observed in the case of adopting (2.5), since more impedance variations along the NTL are allowed to meet the design objective.

An optimum compact NTL transformer of length \( d \) should have its \( ABCD \) parameters at a certain design frequency \( f \) as close as possible to those of the uniform one of length \( d_0 \) \((d < d_0)\). Hence, the optimum Fourier coefficients values are obtained by minimizing the following error function [76]:

\[
\text{Error} = \sqrt{\frac{1}{4} \left( |A-A_0|^2 + Z^{-2} |B-B_0|^2 + Z^{-2} |C-C_0|^2 + |D-D_0|^2 \right)}
\]

(2.7)

where \( A_0, B_0, C_0, \) and \( D_0 \) are the \( ABCD \) parameters of the uniform transmission line. The resulting \( Z(x) \) must be within reasonable fabrication tolerances and meet matching conditions. That is, the following physical constraints are set [76]:

\[
Z_{\min} \leq Z(x) \leq Z_{\max}
\]

(2.8)

\[
Z(0) = Z(d) = Z
\]

(2.9)

The constraint presented in (2.8) confines the impedance profile within minimum and maximum widths so that fabrication limitations are not exceeded; whereas (2.9) ensures that both NTL terminations are equal and match the uniform line impedance \( Z \). To minimize the non-linear bound-constrained error function in (2.7), an optimization procedure is carried out, in which the series coefficients are set as the variables to be optimized. It is noteworthy to point out that such coefficients \( \in [-1,1] \). Trust-region-reflective algorithm is used in this context for its strong convergence properties [77].
2.2 Non-Uniform Transmission Line Modeling

An optimization procedure is carried out to solve the bound-constrained non-linear minimization problem at the expense of simulation time and computational effort. ANNs, in this context, are one of the best candidates in addressing the above challenges, owing to their ability to process the interrelation between the electrical and physical characteristics of an NTL in a superfast manner. The basis of ANN modeling is to capture the inherent input-output functional relationship and model any complexity with ease. Because of the various training algorithms, ANNs can be trained to achieve a better convergence. Furthermore, the dynamic allocation of the hidden neurons significantly assists the learning phase as compared with other modeling approaches, such as splines or polynomials. Hence, ANNs were broadly applied in modeling modern microwave components that possess a high degree of non-linearity [78]-[82].

Based on the universal approximation theorem, a three-layer multi-layer perceptron (MLP) neural network, also known as MLP-3, can model any non-linearity with tolerable error [83]. The proposed model, thus, utilizes supervised MLP-3 neural networks. During the training process, weights and biases of the ANN are adjusted to determine the appropriate number of hidden neurons required to minimize the prediction error [84]. It will be seen that the achieved accuracy and the quick prediction of the impedance variations are two key advantages of the proposed model.

Although NTLs modeling is based on training data from an optimization-driven procedure, the design approach is valid for EM simulations, and has a particular usefulness to electronic manufacturing industry where PCB layouts are often reused with repeated modifications to the existing time-tested designs.
2.3 Results and Discussions

A design example of compact NTL transformer that matches a source impedance \(Z_s\) to a load impedances \(Z_l\) such that \(Z = (Z_sZ_l)^{0.5}\). Here, \(Z_s\) and \(Z_l\) equal 100 \(\Omega\) and 25 \(\Omega\), respectively, and \(Z = 50 \Omega\). The design frequency is set to 0.5 GHz. The used substrate is 1.6-mm-thick FR4 with a relative permittivity of 4.6. The NTL transformer is designed to have a length \(d = 56 \text{ mm}\) (shortest possible, obtained after multiple optimization trials with different values of the parameters involved) and width variation between \(w_{\text{min}} = 0.2 \text{ mm}\) and \(w_{\text{max}} = 10 \text{ mm}\). This length is 32% shorter than that of the uniform transformer of a length \(d_0 = 82 \text{ mm}\). Series given in (2.6) is adopted, with terms \(N\) and sections \(K\) of 10 and 50, respectively. During the optimization process, Matlab function ‘fmincon’ is utilized considering 1000 iterations. The modeling procedure is performed after obtaining a reasonable size of training data by running adequate optimizations, taking into account that the design frequency \(f\), the minimum and maximum widths \(w_{\text{min}}\) and \(w_{\text{max}}\), respectively, and the length \(d\) are considered as input parameters; whereas the series coefficients \(a_0, a_1 \ldots a_{10}\) in (2.6) are set as outputs which once determined, \(Z(x)\) is obtained. Figure 2-2 shows the proposed model (trained with three different techniques) associated with the training statistics.

```
Model Statistics
Data Size 3617×15
Training Samples (%) 66
Validation Samples (%) 34
No. of Inputs 4
Hidden Neurons 12
No. of Outputs 11
Training Method BP, QN, CG
Training Error (%)
BP: 4.607, QN: 2.872, CG: 4.014
Validation Error (%)
BP: 4.596, QN: 2.856, CG: 3.982
```

Figure 2-2: ANN model of a NTL transformer trained with backpropagation, quasi-Newton, and conjugate gradient techniques.
Figure 2-3 shows the resulting $ABCD$ parameters of various NTL transformers generated from: (1) optimizations, (2) backpropagation (BP)-based, (3) quasi Newton (QN)-based, and (4) conjugate gradient (CG)-based models. The developed ANN model is valid for the specific FR4 substrate. Thus, updated optimizations should be performed to re-gather new sets of training data in the case of changing the substrate.

Figure 2-3: $ABCD$ parameters comparison between the conventional uniform transformer; compact optimized NTL transformer; and the ANN-modeled NTL transformer: (a) parameter $A$; (b) parameter $B$; (c) parameter $C$; (d) parameter $D$. 
The \textit{ABCD} parameters of the optimized design and those generated by the variously-trained ANN are close to those of the conventional $\lambda/4$ transformer at the frequency of interest (0.5 GHz). It is also seen that QN- and CG-based algorithms have better accuracy than BP, as the QN approach utilizes the 1\textsuperscript{st} and 2\textsuperscript{nd} derivatives to interrelate input/output data; whereas the conjugate direction (instead of the gradient direction) in CG results in a faster convergence. The differences between the \textit{ABCD} parameters of the optimized NTL transformer (and thus, from ANN modeling) and those representing the conventional uniform matching transformer, especially at the higher frequencies, are fundamentally due to enforcing the equivalency characterized by (2.7) only at a single design frequency. Hence, \{\textit{A},\textit{A}_0\}, \{\textit{B},\textit{B}_0\}, \{\textit{C},\textit{C}_0\}, \text{ and } \{\textit{D},\textit{D}_0\} \text{ are almost equivalent up to } 0.5 \text{ GHz. As a result, higher order harmonics that present in the conventional matching transformer are efficiently suppressed.}

Width and impedance variations as a function of the NTL transformer length are shown in Figure 2-4. An excellent confinement of $W(x)$ in the allowed variation boundary [0.2,10] mm is achieved, with QN- and CG-based training methods being in more proximity to the NTL transformer obtained by optimizations than BP-trained model. $Z(x)$ varies within the interval [22,150] $\Omega$, complying with the minimum and maximum widths. The final designs obey the condition given in (2.9), since $Z(0) = Z(d) = Z = 50 \text{ } \Omega$. Discrepancies between the developed ANN model and optimized NTLs are due to the generated error in the training phase (various error values are described in Figure 2-2).

Upon examining the physical properties of the optimized and modeled NTL transformers, electrical performance assessment is carried out by calculating and plotting the $S$-parameters of the NTLs as depicted in Figure 2-5, keeping in mind that [85]-[86]:

```
Figure 2-4: Optimized and ANN-based NTL transformers variations as a function of length: (a) width $W(x)$; (b) impedance $Z(x)$.

\[
S_{11} = \frac{AZ_l + B - CZ_l Z_i - DZ_s}{AZ_l + B + CZ_l Z_i + DZ_s}
\]

(2.10)

\[
S_{21} = 2 \left( \frac{AD - BC}{AZ_l} \right) (Z_l Z_i)^{0.5}
\]

(2.11)

where $Z_i$ and $Z_l$ equal 100 $\Omega$ and 25 $\Omega$, respectively.

Figure 2-5: Optimized and ANN-based NTL transformers $S$-parameters: (a) $|S_{11}|$; (b) $|S_{21}|$. 
As seen in Figure 2-5, input port matching ($S_{11}$) and transmission ($S_{21}$) parameters of the proposed variously-trained ANN model are close to those obtained by the time- and effort-consuming optimizations. Alike the optimization-driven transformer, an excellent input matching of below –30 dB at 0.5 GHz is obtained by the BP-, QN-, and CG-based training techniques. Furthermore, the transmission in the optimized and ANN-trained transformers at the design frequency is less than –0.1 dB. A 0.1 GHz positive frequency shift is seen in the results of BP-based modeling due to the associated larger error in the training and testing phases.

The optimized and modeled NTLs are also validated by full-wave EM simulations with the finite element method-based tool ANSYS HFSS [87]. Figure 2-6 shows the $S$-parameters of the optimized and QN-based ANN NTL profiles. A clear resonance at the design frequency (0.5 GHz) is achieved, with $S_{11}$ values below –15 dB and $S_{21}$ values of around –0.3 dB. Differences between such results and those presented in Figure 2-5 are mainly due to different types of losses (e.g., conductor and dielectric losses).

Figure 2-6: Simulations of the optimized and ANN QN-based NTL transformers.
Table 2.1 shows a comparison between the NTL transformers obtained from optimization and modeling. Non-uniform widths are in a close match. Besides, ANN modeling has a major improvement in the simulation time and allocated memory. Although $w_{\text{min}}$ is set to 0.2 mm during optimizations, the resulting $w_{\text{min}}$ is 0.15693 mm due to the optimization error. Thus, the inputs used in the training and validation phase, including $w_{\text{min}}$ and $w_{\text{max}}$, are also affected by optimization errors.

Table 2.1: Comparison between optimized and ANN-based non-uniform transformers.

<table>
<thead>
<tr>
<th></th>
<th>$w_{\text{min}}$ (mm)</th>
<th>$w_{\text{max}}$ (mm)</th>
<th>Simulation time (sec)</th>
<th>memory (Kb)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optimized</td>
<td>0.15693</td>
<td>10.02</td>
<td>16.03319</td>
<td>23920</td>
</tr>
<tr>
<td>ANN Model-BP</td>
<td>0.17838</td>
<td>8.1</td>
<td>Order of 0.01</td>
<td>&lt;10</td>
</tr>
<tr>
<td>ANN Model-QN</td>
<td>0.17681</td>
<td>10</td>
<td>Order of 0.01</td>
<td>&lt;10</td>
</tr>
<tr>
<td>ANN Model-CG</td>
<td>0.20258</td>
<td>10</td>
<td>Order of 0.01</td>
<td>&lt;10</td>
</tr>
</tbody>
</table>

Simulation time of only one (the best) trial is included in the above table. However, it normally requires the designer several optimization trials to obtain acceptable values of the series coefficients to proceed with.

2.4 Conclusions

The optimization and modeling of miniaturized microstrip NTLs are discussed. Both physical and electrical characteristics are optimized to match given load and source impedances at a predefined frequency utilizing the trust-region-reflective algorithm. The computational complexity of the optimization process is tackled by utilizing a MLP-3 ANN. A case study involving a compact NTL transformer is provided, and the achieved optimization accuracy and the superfast modeling of the impedance variations are justified. S-parameters of the conventional uniform transformer and those obtained by optimization and modeling are in excellent agreement at the frequency of interest. The analysis presented in this chapter illustrates a conceptual example of a simple two-port network operating at a single frequency. Extending the problem to address multi-port networks with advanced electrical characteristics significantly increases optimization time and computational complexity as will be seen in the next chapter.
Chapter 3

Ultra-Wideband Wilkinson Power Divider

This chapter presents a technique with clear guidelines to design a compact UWB WPD. The design procedure is accomplished by replacing the uniform transmission lines in each arm of the conventional power divider with impedance-varying profiles. Impedance variations are governed by a truncated Fourier series with coefficients optimized to achieve UWB frequency matching. The design procedure is divided into two main steps: 1) even-mode analysis, carried out to optimize the series coefficients according to the intended performance; 2) odd-mode analysis, utilized to obtain the optimum isolation resistors that guarantee an acceptable isolation and output ports matching over the frequency range of interest. The proposed design procedure results in an easy-to-fabricate single-layered structure. The optimization-driven framework is also modeled utilizing a QN-based trained ANN to address the burden in optimization time and complexity, leading to valued benefits to design engineers.

The chapter is organized as follows Section 3.1 presents the detailed design procedure of the two analysis modes. Simulated and measured results of a designed in-phase equal-split UWB divider are given in Section 3.2. ANN modeling of the same example is presented in Section 3.3. Finally, conclusions are given in Section 3.4.
3.1 Design

A schematic diagram of the conventional and proposed UWB component is shown in Figure 3-1. Each uniform impedance in the conventional divider is replaced with a NTL of length \(d\), and varying characteristic impedance and propagation constant, \(Z(x)\) and \(\beta(x)\), respectively, to achieve UWB operation. Such a response is obtained by properly profiling the impedance of the NTL.

![Schematic diagram of conventional and proposed UWB component](image)

(a) (b)

Figure 3-1: Schematic diagrams of (a) conventional single-frequency WPD; (b) proposed UWB WPD utilizing NTLs.

Figure 3-2 demonstrates the corresponding even and odd mode circuits of the proposed design. In Section 3.1.1 (even-mode analysis), the design of the NTL is presented; whereas in Section 3.1.2 (odd-mode analysis), the values of the isolation resistors are optimized to meet acceptable output ports’ isolation and matching.

![Proposed non-uniform WPD](image)

(a) (b)

Figure 3-2: Proposed non-uniform WPD: (a) even-mode; (b) odd-mode circuits.
3.1.1 Even-Mode Analysis

The even-mode equivalent circuit is shown in Figure 3-2(a). The goal is to match a source impedance \( Z_s \) to a load impedance \( Z_l \) across 3.1–10.6 GHz. In our case, \( Z_s = 2Z_0 \) and \( Z_l = Z_0 \). Due to the symmetric excitation at the two output ports, the isolation resistors \( R_r/2 (r = 1, 2, 3) \) are open-circuited. The NTL is designed by enforcing the magnitude of input reflection coefficient, \( |\Gamma_{in}| \), to be zero (or very small) over the intended frequency range. \( |\Gamma_{in}| \) at the input port can be expressed in terms of \( Z_{in}' \), where \( Z_{in}' \) is calculated after obtaining the \( ABCD \) parameters of the NTL as indicated in equations (2.1–2.4) presented in Section 2.1. During the calculations of the \( ABCD \) parameters, the non-uniform profile given in (2.5) is considered for the characteristic impedance \( Z(x) \). The impedance \( Z \), which equals to \((Z_sZ_l)^{0.5}\), is the characteristic impedance of the conventional WPD arm. \( Z(x) \) should be restricted by the constraints given in equations (2.8) and (2.9) stated in Section 2.1. An optimum designed NTL has \( |\Gamma_{in}| \) at each \( f \in [f_l,f_h] \), where \( f_l = 3.1 \) GHz and \( f_h = 10.6 \) GHz are the lowest and highest frequencies, respectively, with an increment of \( \Delta f \), as close as possible to zero. Hence, the optimum values of the coefficients are obtained by minimizing the following error function [88]:

\[
\text{Error}_{in} = \max(E_{f_l}^{in},...,E_{f_h}^{in})
\]

(3.1)

where,

\[
E_{f_i}^{in} = |\Gamma_{in}|^2
\]

(3.2)

\[
\Gamma_{in} = \frac{Z_{in}' - Z_s}{Z_{in}'+Z_s}
\]

(3.3)

\[
Z_{in}' = \frac{A_{Z(x)}Z_i + B_{Z(x)}}{C_{Z(x)}Z_i + D_{Z(x)}}
\]

(3.4)
3.1.2 Odd-Mode Analysis

The odd-mode analysis is carried out to obtain the resistors’ values needed to achieve the optimum output ports isolation and matching conditions. Figure 3-2(b) shows the equivalent odd-mode circuit of the proposed divider [89], where the isolation resistors are distributed uniformly along the NTL (a resistor every $d/3$ distance). Three resistors are adequate to achieve the desired isolation and matching. Interested scholars may refer to [90] for a detailed study on the effect of the number of resistors on the performance.

The asymmetric excitation of the output ports results in terminating each $R_{r}/2$ resistor with a short circuit. Once the optimum values of the Fourier coefficients are determined by following the procedure described in Section 3.1.1, the NTL is subdivided into 3 sections, and the $ABCD$ matrix for each section is calculated. Then, the total $ABCD$ matrix of the whole network shown in Figure 3-2(b) can be calculated as follows [90]:

$$
[ABCD]_{\text{Total}} = [ABCD]_{\frac{R_{l}}{2}} \cdot [ABCD]_{\text{1st Section}} \cdot [ABCD]_{\frac{R_{l}}{2}} \cdot [ABCD]_{\text{2nd Section}} \cdot [ABCD]_{\frac{R_{l}}{2}} \cdot [ABCD]_{\text{3rd Section}}
$$

Finally, and as illustrated in Figure 3-2(b), the following equation can be written:

$$
\begin{bmatrix}
    V_1 \\
    I_1 \\
\end{bmatrix}
= 
\begin{bmatrix}
    A & B \\
    C & D \\
\end{bmatrix}
\begin{bmatrix}
    V_2 \\
    I_2 \\
\end{bmatrix}
$$

Setting $V_2$ in (3.6) to zero, and solving for $V_1/I_1$, one obtains:

$$
\frac{V_1}{I_1} = \frac{B}{D} = Z_{in}^{o}
$$

For a perfect output matching at each at $f_0$, the following error are minimized:

$$
\text{Error}_{out} = \max(E_{f_1}^{out}, \ldots, E_{f_j}^{out}, \ldots, E_{f_h}^{out})
$$

where,
This optimization problem is solved keeping in mind that $R_1$, $R_2$, and $R_3$ are the optimization variables to be determined, which in order to obtain, the series coefficients must first be optimized. Figure 3-3 illustrates the design steps of proposed divider.

\[
E_f^{\text{out}} = |\Gamma_{\text{out}}|^2 \tag{3.9}
\]

\[
\Gamma_{\text{out}} = \frac{Z_{\text{in}}^* - Z_0}{Z_{\text{in}}^* + Z_0} \tag{3.10}
\]

Figure 3-3: Flowchart demonstrating the design of the proposed UWB divider; green and red enclosures present the even- and odd-mode analyses, respectively.
As seen in Figure 3-3, two consequent optimization routines are carried out within each analysis mode (i.e., even and odd) to realize the design approach of the proposed UWB impedance-varying WPD: First, series coefficients that result in UWB matching are first optimized during the even-mode analysis; Second, the obtained coefficients are fed to a new optimization process that is governed by the odd-mode equations to acquire the values of the isolation resistors that best achieve output ports matching and isolation.

### 3.2 Simulations and Measurements

Based on the design procedure presented in Section 3.1, an example of an in-phase equal-split UWB power divider is designed, simulated, fabricated, and measured. A characteristic impedance of $Z_0 = 50 \, \Omega$ is considered taking into account, a 0.813-mm-thick Rogers RO4003C substrate with a relative permittivity of 3.55, and a loss tangent of 0.0027. The length of each NTL arm of the proposed WPD is set to $d = 10 \, \text{mm}$ (almost equal the length at center frequency of 6.85 GHz), and the widths are bounded by 0.15–2.5 mm. $K$ and $N$ are set to 50 and 5, respectively. The frequency increment $\Delta f$ is set to 0.5 GHz. Minimization of the objective functions in (3.1) and (3.8) are performed using two separate but subsequent subroutines using Matlab, each of 3000 iterations.

Figure 3-4 illustrates the resulting simulated and measured $S$-parameters. The input and output ports matching ($S_{11}$) and ($S_{22}$), respectively, as well as the isolation ($S_{23}$) are below $-10 \, \text{dB}$ over the 3.1–10.6 GHz frequency band. The transmission parameter ($S_{21}$) varies between $-3.2 \, \text{dB}$ and $-5 \, \text{dB}$, and is in close proximity to its theoretical value of $-3 \, \text{dB}$. Here, $S_{ij} = S_{ji}$ based on reciprocity concept; while $S_{22} = S_{33}$ and $S_{21} = S_{31}$ as the divider of an equal split type. Discrepancies between simulations and measurements are due to the fabrication process and measurement errors.
Figure 3-4: Simulated and measured $S$-parameters of the proposed UWB divider.

Figure 3-5(a) shows the measured amplitude and phase imbalances between the two output ports of the proposed equal-split in-phase UWB divider. The measured phase imbalance is less than $\pm 10^\circ$ over the entire design frequency range. The obtained amplitude imbalance is around $\pm 0.1$ dB over the 3.1–10.6 GHz band. Such imbalance values prove an excellent symmetry degree of the implemented structure. Figure 3-5(b) depicts the simulated and measured group delays of the designed WPD. Both results are almost flat over the UWB range, and are less than 0.2 ns with a mismatch thought to be due to the inhomogeneous substrate material used in this project.

Figure 3-5: (a) measured amplitude and phase imbalance of the proposed UWB NTL divider; (b) simulated and measured group delay.
3.3 Non-Uniform Ultra-Wideband Divider Modeling

According to the analysis presented in Section 3.1, the design of an NTL-based UWB divider requires two optimization phases: 1) in the even-mode circuit, the series coefficients are set as optimization variables to obtain an NTL with an UWB frequency response; and 2) in the odd-mode circuit, and upon determining the optimum series coefficients, the values of the three isolation resistors are optimized to achieve acceptable output ports matching and isolation.

Two distinctive optimization routines are required in the overall UWB divider design, which adds to the computational demand and design time. As such, an ANN model is built and trained to meet these challenges. Figure 3-6 shows the proposed model of the 3.1–10.6 GHz power divider.

Based on the proposed model, all design parameters are achieved with only a single stage (in contrast to the optimization approach that requires two extensive sets of calculations to find the Fourier series coefficients and the isolation resistors values). The proposed ANN is trained with different techniques. However, only QN-based method (which results in the highest accuracy) is included for the sake of brevity.

**ANN Model Statistics**

| Data Size | 54×17 |
| Training Samples (%) | 70 |
| Validation Samples (%) | 30 |
| No. of Inputs | 3 |
| No. of Hidden Neurons | 11 |
| No. of Outputs | 14 |
| Training Method | QN |
| Training Error (%) | 5.168 |
| Validation Error (%) | 7.976 |

**Figure 3-6:** Proposed ANN model of the UWB non-uniform WPD.
Width and impedance variations of the optimized and modeled NTL divider arms are presented in Figure 3-7. An acceptable match between the optimized and modeled results is noticed, and the non-uniform width is bounded within the predefined \(w_{\text{min}}\) and \(w_{\text{max}}\); whereas \(Z(x) \in [40,128] \ \Omega\). Widths and impedance variations are not symmetric around the y-axis (on the contrary to the ones shown in Figure 2-4), mainly due to the added ‘sine’ terms in the truncated series.

![Figure 3-7: Optimized and ANN-based non-uniform WPD arm variations as a function of length: (a) width; (b) impedance.](image)

The calculated \(S\)-parameters for both optimized and ANN-based UWB NTL dividers are shown in Figure 3-8. Such parameters are generated with the following equations being considered [91]:

\[
S_{21} = S_{31} = 20 \log \left( \frac{1-S_{11}^2}{2} \right) \tag{3.11}
\]

\[
S_{22} = S_{33} = 20 \log \left( \frac{1}{2} \left( \frac{Z_{\text{out}}^e - Z_{i}}{Z_{\text{out}}^e + Z_{i}} + \frac{Z_{\text{out}}^o - Z_{i}}{Z_{\text{out}}^o + Z_{i}} \right) \right) \tag{3.12}
\]

\[
S_{32} = S_{23} = 20 \log \left( \frac{1}{2} \left( \frac{Z_{\text{out}}^e - Z_{i}}{Z_{\text{out}}^e + Z_{i}} - \frac{Z_{\text{out}}^o - Z_{i}}{Z_{\text{out}}^o + Z_{i}} \right) \right) \tag{3.13}
\]
where $S_{11}$ is calculated using (2.10). The generated input port matching, output port matching, isolation, and transmission parameters from the proposed model over the frequency range of interest are in excellent agreement with those generated by optimizations. $S_{11}$, $S_{22}$, and $S_{23}$ are all below $-10$ dB across 3.1–10.6 GHz; whereas $S_{21}$ is around $-3.2$ dB over the same UWB range. The small differences between simulated and modeled results are due to the error induced by training.

Figure 3-8: Calculated $S$-parameters of the UWB WPD for optimized and modeled resistors of $\{R_1, R_2, R_3\} = \{151, 237.6, 147.4\}$ and $\{156.6, 252.8, 148.8\}$, respectively: (a) $|S_{11}|$; (b) $|S_{21}|$; (c) $|S_{22}|$; (d) $|S_{23}|$.
Figure 3-9 shows the simulated $S$-parameters of the optimized and modeled UWB WPDs. Both results are in a good agreement, with small discrepancies between the optimized and modeled designs basically due to errors induced during the training phase. Both approaches result in input/output ports matching and isolation of less than $-10$ dB, and acceptable transmission over the intended frequency range. Degradations of $S_{21}$ as frequency increases are due to different types of losses (e.g., dielectric and conductor losses). Table 3.1 shows a comparison between the optimized and modeled WPDs.

Figure 3-9: Full-wave EM simulations of the optimized and ANN-based UWB WPD: (a) $|S_{11}|$, (b) $|S_{21}|$, (c) $|S_{22}|$, and (d) $|S_{23}|$. 

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>$S_{11}$ (dB)</th>
<th>Optimized</th>
<th>ANN Model-QN</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>-30</td>
<td>-25</td>
<td>-20</td>
</tr>
<tr>
<td>4</td>
<td>-15</td>
<td>-10</td>
<td>-5</td>
</tr>
<tr>
<td>6</td>
<td>-5</td>
<td>-2.5</td>
<td>-1</td>
</tr>
<tr>
<td>8</td>
<td>-0.5</td>
<td>0</td>
<td>1.5</td>
</tr>
<tr>
<td>10</td>
<td>2</td>
<td>5</td>
<td>10</td>
</tr>
<tr>
<td>12</td>
<td>10</td>
<td>15</td>
<td>20</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>$S_{21}$ (dB)</th>
<th>Optimized</th>
<th>ANN Model-QN</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>-30</td>
<td>-25</td>
<td>-20</td>
</tr>
<tr>
<td>4</td>
<td>-15</td>
<td>-10</td>
<td>-5</td>
</tr>
<tr>
<td>6</td>
<td>-5</td>
<td>-2.5</td>
<td>-1</td>
</tr>
<tr>
<td>8</td>
<td>-0.5</td>
<td>0</td>
<td>1.5</td>
</tr>
<tr>
<td>10</td>
<td>2</td>
<td>5</td>
<td>10</td>
</tr>
<tr>
<td>12</td>
<td>10</td>
<td>15</td>
<td>20</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>$S_{22}$ (dB)</th>
<th>Optimized</th>
<th>ANN Model-QN</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>-30</td>
<td>-25</td>
<td>-20</td>
</tr>
<tr>
<td>4</td>
<td>-15</td>
<td>-10</td>
<td>-5</td>
</tr>
<tr>
<td>6</td>
<td>-5</td>
<td>-2.5</td>
<td>-1</td>
</tr>
<tr>
<td>8</td>
<td>-0.5</td>
<td>0</td>
<td>1.5</td>
</tr>
<tr>
<td>10</td>
<td>2</td>
<td>5</td>
<td>10</td>
</tr>
<tr>
<td>12</td>
<td>10</td>
<td>15</td>
<td>20</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>$S_{23}$ (dB)</th>
<th>Optimized</th>
<th>ANN Model-QN</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>-30</td>
<td>-25</td>
<td>-20</td>
</tr>
<tr>
<td>4</td>
<td>-15</td>
<td>-10</td>
<td>-5</td>
</tr>
<tr>
<td>6</td>
<td>-5</td>
<td>-2.5</td>
<td>-1</td>
</tr>
<tr>
<td>8</td>
<td>-0.5</td>
<td>0</td>
<td>1.5</td>
</tr>
<tr>
<td>10</td>
<td>2</td>
<td>5</td>
<td>10</td>
</tr>
<tr>
<td>12</td>
<td>10</td>
<td>15</td>
<td>20</td>
</tr>
</tbody>
</table>
Table 3.1: Comparison between optimized and modeled WPDs.

<table>
<thead>
<tr>
<th></th>
<th>$w_{\min}$ (mm)</th>
<th>$w_{\max}$ (mm)</th>
<th>Simulation time (sec)</th>
<th>memory (Kb)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Optimized$^*$</td>
<td>0.22039</td>
<td>2.6</td>
<td>337.89418</td>
<td>101732</td>
</tr>
<tr>
<td>ANN Model-QN</td>
<td>0.18526</td>
<td>2.6</td>
<td>Order of 0.01</td>
<td>&lt;10</td>
</tr>
</tbody>
</table>

*Simulation time of only one (the best) trial is included in the above table. However, it normally requires the designer several optimizations to obtain an acceptable response.

3.4 Conclusions

A general design of an UWB WPD incorporating Fourier-based impedance-varying profiles is presented. The design of the UWB NTLs is obtained from the even-mode analysis of the WPD. Three isolation resistors are optimized through the odd mode circuit. For verification purposes, an equal-split UWB power divider is designed, simulated, and measured. The good agreement between both simulated and measured results over the 3.1–10.6 GHz frequency range proves the validity of the design procedure. The differences between simulation and experimental results could be due to the fabrication process, the effect of the connectors, and measurement errors.

Furthermore, the modeling of the computationally-expensive impedance-varying physical and electrical characteristics of the proposed UWB NTL divider utilizing MLP-3 ANNs is presented and discussed. The results of the two optimization routines (series coefficients and isolation resistors) are considered in a single-staged model. The achieved accuracy and the superfast modeling of the NTL impedance variations are two major advantages of the proposed model. $S$-parameters derived from the trained ANN outputs are in excellent agreement with those obtained by the time-consuming optimization procedure, and show excellent electrical performance across the UWB frequency range. Although modeling examples are based on training data derived from analytical optimizations, the overall design is accurate as justified by EM simulation results.
Chapter 4

Wideband Multi-Way Bagley Power Divider

The BPD offers structural compactness and excellent input port matching and transmission. Furthermore, its output ports are easily extended to any number according to the given design requirements while maintaining a planar geometry without added design complexity or lumped elements. However, the output ports of BPDs are unmatched, and the isolation is not as good as that of other dividers (e.g., WPD).

In this chapter, the concept of impedance-varying microstrip transmission lines optimized to wideband multi-way BPD is presented. The proposed procedure is based on substituting the single-frequency matching quarter-wave sections in the conventional design by impedance-varying transmission lines of flexible bandwidth allocation and matching levels. Impedance variations are profiled according to a truncated Fourier series with coefficients determined by an optimization procedure.

This chapter is organized as follows: Section 4.1 mathematically discusses the proposed design approach; Section 4.2 presents the obtained analytical results of a 3-way BPD for different design bands; simulated and measured results for 3- and 5-way BPDs of fractional bandwidths 86% and 57%, respectively, are provided in Section 4.3. Finally, conclusions are given in Section 4.4.
4.1 Design

In this section, the design procedure of the proposed impedance-varying divider is presented. Figure 4-1(a) shows a schematic diagram of the wideband multi-way BPD. Figure 4-1(b) depicts the equivalent transmission line model with which design equations are derived based on. As shown in Figure 4-1(b), if $Z_1$ is set such that $Z_1 = 2Z_0$, where $Z_0$ is the characteristic impedance of the ports, the length $d_1$ can be arbitrarily chosen. Hence, the input impedance $Z_{in}^{(1)}$ equals the parallel combination $Z_0 // 2Z_0 = 2Z_0/3$. In general, for a multi-way BPD with $N_o$ odd output ports:

$$Z_{in}^{(n_o)} = \frac{2}{2n_o + 1} Z_0$$

(4.1)

where $n_o = 1, 2, ..., (N_o-1)/2$. If the impedances interconnecting the output ports are chosen such that $Z_2, Z_3, ..., Z_{(N_o-1)/2}$ equal $Z_{in}^{(1)}, Z_{in}^{(2)}, ..., Z_{in}^{((N_o-1)/2 - 1)}$, respectively, lengths $d_2, d_3, ..., d_{(N_o-1)/2}$ can be assigned any values. Consequently, a single-frequency matching uniform quarter-wave length transformer in the conventional BPD design must satisfy the following equation:

$$Z = \sqrt{2Z_0Z_{in}^{((N_o-1)/2)}} = \frac{2Z_0}{\sqrt{N_o}}$$

(4.2)

To obtain a wideband frequency characteristic, the uniform matching transformer is replaced with a NTL of varying impedance and propagation constant $Z(x)$ and $\beta(x)$, respectively, and length $d$. Mathematical formulations start by obtaining the $ABCD$ matrix of the whole NTL transformer by adopting (2.1–2.4) presented in Section 2.1. During the calculations of the $ABCD$ matrix, the impedance profile given equation (2.5) is considered, where $Z$ is the impedance of the conventional multi-way BPD transformer.
The resulting $Z(x)$ must be within reasonable fabrication tolerances and meet matching conditions. That is, the physical constraints expressed by (2.8) and (2.9) in Section 2.1 are taken into account. A wideband NTL transformer has its input reflection coefficient magnitude $|\Gamma_{in}|$ at each $f$ within the frequency range $[f_l, f_h]$ with an increment $\Delta f$ as close as possible to zero. Therefore, we set and minimize the following objective function w.r.t the truncated Fourier series:

$$\text{Objective} = \sum_{j=0}^{(f_h-f_l)/\Delta f} E(f_j + j\Delta f)$$  \quad (4.4)
where,

\[ E(f) = |\Gamma_{in}|^2 \]  

(4.5)

\[ \Gamma_{in} = \frac{Z_{in}' - Z_s}{Z_{in}'+Z_s} \]  

(4.6)

\[ Z_{in}' = \frac{A_{Z(x)}Z_x^{(N_x-1)/2} + B_{Z(x)}Z_x^{(N_x-1)/2} + D_{Z(x)}}{C_{Z(x)}Z_x^{(N_x-1)/2} + D_{Z(x)}} \]  

(4.7)

where \( Z_0 = 2Z_0 \) and \( Z_{in}' \) is the total input impedance shown in Figure 4-1(b). The design steps of the proposed wideband BPD are presented in Figure 4-2.

---

**Figure 4-2:** Flowchart showing the design of the proposed wideband BPD; red enclosure presents formulations based on the equivalent transmission line model.
As seen in Figure 4-2, a wideband BPD with flexible bandwidth allocation can be designed based on the predefined values $f_l$ and $f_h$. Furthermore, the proposed procedure is reasonably simple, and depends on properly modulating impedance variations of the matching transformer during the minimization of the objective function given in (4.4).

### 4.2 Analytical Examples

Three 3-way BPD examples are discussed. Three different frequency bands are considered in this study to demonstrate the efficiency of the proposed methodology: 6–8 GHz, 5–9 GHz, and 4–10 GHz, which correspond to fractional bandwidths of 28%, 57%, and 86%, respectively. A 0.787-mm-thick RT/duroid 5880 substrate with a relative permittivity of 2.2 and dielectric loss tangent of 0.0009 is used in all examples. For a compact BPD design, $d$ is chosen to be $\lambda/4$ at center frequency of 7 GHz. $K$ and $N$ for $Z(x)$ are set to 25 and 5, respectively, which are sufficient to achieve the required optimization goals, and $\Delta f = 0.5$ GHz. The reference impedance $Z$ is calculated using (4.2) and equals 57.735 $\Omega$. $Z(x)$ is bounded by minimum and maximum values to ensure realization and matching within fabrication limits. In other words, $Z(x) \in [38,165]$ $\Omega$, which correspond to width variations between 0.15 and 3.5 mm. The minimization of the objective function in (4.4) was performed in 1000 iterations using Matlab. Figure 4-3(a) shows the resulting impedance profiles for the proposed designs; whereas width variations are illustrated in Figure 4-3(b).

All non-uniform profiles are constrained by the impedance interval mentioned above (and thus, by the predefined width variations). Table 4.1 shows the resulting coefficients with the associated optimization error in each example.
Figure 4-3: NTL transformer designs for the three different proposed fractional bandwidths: (a) impedance variations; (b) width variations.

Analytical response of the transmission and input port matching parameters of the three designs is shown in Figure 4-4. $S_{21} = S_{31} = S_{41}$ since the proposed dividers are of an equal split type. $S$-parameters are calculated using the following equations:

\[
S_{11} = 20 \log \left( \frac{A_{Z(x)}Z_{in}^{(1)} + B_{Z(x)} - C_{Z(x)}Z_{x}Z_{in}^{(1)} - D_{Z(x)}Z_{x}}{A_{Z(x)}Z_{in}^{(1)} + B_{Z(x)} + C_{Z(x)}Z_{x}Z_{in}^{(1)} + D_{Z(x)}Z_{x}} \right) \quad (4.8)
\]

\[
S_{21} = S_{31} = S_{41} = 20 \log \left( \sqrt{1 - S_{11}^2} \right) / 3 \quad (4.9)
\]

As seen in Figure 4-4, $S_{11}$ of below –20 dB is obtained in each example, and $S_{21}$ is in close proximity to its theoretical value of –4.77 dB over the design bands. The theoretical results are validated in Section 4.3 using detailed simulated and measured studies that include: $S$-parameters, group delay, and physical symmetry.

Table 4.1: Optimized Fourier series coefficients for the three 3-way BPD examples.

<table>
<thead>
<tr>
<th>FB (%)</th>
<th>$c_0$</th>
<th>$a_1$</th>
<th>$a_2$</th>
<th>$a_3$</th>
<th>$a_4$</th>
<th>$a_5$</th>
<th>$b_1$</th>
<th>$b_2$</th>
<th>$b_3$</th>
<th>$b_4$</th>
<th>$b_5$</th>
<th>Error</th>
</tr>
</thead>
<tbody>
<tr>
<td>28</td>
<td>0.1221</td>
<td>0.0378</td>
<td>0.3798</td>
<td>-0.1638</td>
<td>0.0415</td>
<td>-0.0865</td>
<td>-0.3809</td>
<td>0.1675</td>
<td>-0.0620</td>
<td>-0.0020</td>
<td>-0.0535</td>
<td>0.001</td>
</tr>
<tr>
<td>57</td>
<td>0.1392</td>
<td>0.0180</td>
<td>0.3694</td>
<td>-0.1273</td>
<td>-0.0154</td>
<td>0.0549</td>
<td>-0.2442</td>
<td>-0.0631</td>
<td>-0.0922</td>
<td>0.1189</td>
<td>-0.1580</td>
<td>0.021</td>
</tr>
<tr>
<td>86</td>
<td>0.1448</td>
<td>0.0186</td>
<td>0.3218</td>
<td>0.0204</td>
<td>0.0265</td>
<td>-0.0540</td>
<td>0.0186</td>
<td>-0.2481</td>
<td>-0.2488</td>
<td>0.2440</td>
<td>-0.2437</td>
<td>0.112</td>
</tr>
</tbody>
</table>
4.3 Simulations and Measurements

This section discusses full-wave EM simulated and measured results for 3- and 5-way wideband BPDs with fractional bandwidths of 86% (i.e., 4–10 GHz) and 57% (i.e., 5–9 GHz), respectively. Simulations were performed with Ansys HFSS. Figure 4-5 shows photographs of the proposed dividers built with the substrate mentioned earlier. Figure 4-6 and Figure 4-7 show the simulated and measured S-parameters of the 3- and 5-way BPDs, respectively. It should be pointed out that, ideally, $S_{21} = S_{41}$ in the 3-way divider; whereas $S_{21} = S_{61}$ and $S_{31} = S_{51}$ in the 5-way divider due to structures symmetry.
Figure 4-6: Simulated and measured $S$-parameters of the proposed 3-way NTL BPD: (a) $|S_{11}|$; (b) $|S_{21}|$; (c) $|S_{31}|$.

Figure 4-7: Simulated and measured $S$-parameters of the proposed 5-way NTL BPD: (a) $|S_{11}|$; (b) $|S_{21}|$; (c) $|S_{31}|$; (d) $|S_{41}|$. 
Simulated and measured $S_{11}$ of the 3- and 5-way BPDs are in a good agreement, and are below $-15$ dB and $-12.5$ dB, respectively, over the bands of interest. The transmission parameters of the 3-way BPD equal $-4.77 \pm 1$ dB across 4–10 GHz, and are $-7 \pm 1$ dB over 5–9 GHz in the 5-way BPD. Discrepancies between the simulated and measured results are thought to be due to measurement errors (connector/cable losses).

Simulated and measured group delays of the proposed 3- and 5-way BPDs are shown in Figure 4-8. Measured results are in close proximity to those obtained from simulations, and show almost constant response of 0.15 ns for both $S_{21}$ and $S_{31}$ in the 3-way divider over the 4–10 GHz band (Figure 4-8(a)). Similarly, Figure 4-8(b) shows constant group delays for $S_{21}$, $S_{31}$, and $S_{41}$ of 0.18 ns, 0.21 ns, and 0.22 ns, respectively, in the 5-way divider across the design band. Structural symmetry of the fabricated dividers is assessed by measuring the magnitude and phase imbalance as indicated in Table 4.2.

![Figure 4-8: BPDs simulated and measured group delays: (a) 3-way; (b) 5-way](image)

Table 4.2: Measured metrics of the proposed dividers magnitude/phase imbalances.

<table>
<thead>
<tr>
<th></th>
<th>Magnitude (dB)</th>
<th>Phase (Deg.)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3-Way BPD</td>
<td>$</td>
<td>S_{21}</td>
</tr>
<tr>
<td>5-Way BPD</td>
<td>$</td>
<td>S_{21}</td>
</tr>
<tr>
<td></td>
<td>$</td>
<td>S_{31}</td>
</tr>
</tbody>
</table>
Figure 4-9 and Figure 4-10 show the magnitude and phase imbalance of the proposed 3- and 5-way BPDs, respectively. As seen in Figure 4-9, the measured magnitude imbalance of the 3-way divider equals ±1 dB; whereas the phase imbalance is measured to be ±6° over the design bandwidth. Figure 4-10 shows the measured magnitude and phase imbalance of the 5-way BPD, which equal ±0.75 dB and ±5°, respectively, across the 5–9 GHz band. Such results indicate an excellent symmetry of the two fabricated divider layouts.

![Figure 4-9: Measured imbalance of the 3-way BPD: (a) magnitude; (b) phase.](image)

![Figure 4-10: Measured imbalance of the 5-way BPD: (a) magnitude; (b) phase.](image)
Simulated and measured output ports matching and isolation of the 4–10 GHz 3-way BPD are given in Figure 4-11 and Figure 4-12, respectively; whereas those of the 5-way divider are not included for the sake of brevity. Figure 4-11 show that the isolation between output ports, characterized by $S_{23}$, $S_{34}$, and $S_{24}$, varies between $-4$ dB and $-10$ dB across the design band. Output matching parameters $S_{22}$, $S_{33}$, and $S_{44}$ are around $-5$ dB. Thus, the BPD output ports are neither isolated nor matched at the design frequency(s). However, BPDs possess a compact area, and can easily be extended to any number of output ports. Hence, BPDs are excellent candidates in only-transmitting antenna feeding networks (e.g., broadcasting applications).

Figure 4-11: Output ports isolation of the 3-way BPD: (a) $|S_{23}| = |S_{34}|$; (b) $|S_{24}|$.

Figure 4-12: Output ports matching of the 3-way BPD: (a) $|S_{22}| = |S_{44}|$; (b) $|S_{33}|$. 
4.4 Conclusions

The concept of Fourier-based impedance-varying profiles of wideband frequency matching characteristic is adopted in the design of compact wideband multi-way BPDs. The equivalent transmission line model is used to profile impedance variations by finding the optimum series coefficients that result in a wideband matching nature. The proposed methodology leads to flexible spectrum allocation and matching level. Moreover, the resulting structures are compact and planar.

Three 3-way BPDs of different fractional bandwidths are designed to validate the proposed technique. Then, two examples of 3- and 5-way BPDs with bandwidths of 4–10 GHz and 5–9 GHz, respectively, are simulated, fabricated, and measured. Simulated and measured results show an excellent agreement, with input port matching of below –15 dB and –12.5 dB for the 3- and 5-way dividers, respectively, over the bands of interest. The obtained transmission parameters of the 3- and 5-way dividers are \(-4.77 \pm 1\) dB and \(-7 \pm 1\) dB, respectively, over the design bands. Wideband Bagley dividers may find many applications, especially in only-transmitting antenna subsystems.
Chapter 5

Multi-/Broadband Quadrature Branch-Line Coupler

Microwave couplers are essential components for a host of system applications (e.g., modern radars, test equipment, RF mixers) where reduced-size circuitry, multi-/broadband operation, and arbitrary coupling levels are important requirements. The hybrid BLC is among such couplers that is extensively addressed in literature, with an emphasis on improving its spectrum accessibility by proposing multi-/broadband designs.

In this chapter, the concept of NTLs optimized to multi-/broadband BLCs is presented. The proposed procedure is based on substituting uniform matching quarter-wave branches in the conventional design by impedance-varying lines of multi-frequency or broadband nature. The adopted concept results in suppressing higher order harmonics, and have the merit of achieving arbitrary coupling levels.

This chapter is organized as follows: Section 5.1 presents a mathematical platform of a multi-frequency BLC. Two design examples of dual- and triple-frequency BLCs are conveyed for verification purposes. Section 5.2 discusses the concept of NTLs in the design of broadband BLCs with arbitrary coupling levels and higher-order harmonics suppression, where design examples of 3-dB, 6-dB, and 9-dB BLCs are given. Finally, conclusions and remarks are drawn in Section 5.3.
### 5.1 Multi-Frequency Branch-Line Coupler

Figure 5-1 shows the conventional BLC and the proposed multi-band design. The proposed BLC has six variable-impedance profiles formed from $Z_{1,2}(x)$, with lengths $d_{1,2}$. Figure 5-2 depicts the corresponding even- and odd-mode circuits of the proposed multi-frequency BLC, with which the mathematical derivation and representation is based on.

![Figure 5-1: Schematics of: (a) conventional single-frequency BLC; (b) proposed multi-frequency BLC utilizing NTLs.](image1)

![Figure 5-2: Proposed non-uniform BLC circuits: (a) even-mode; (b) odd-mode.](image2)
Overall $ABCD$ matrix of each analysis mode at each frequency $f_m$ ($m = 1, 2, \ldots M$) is found by multiplying the $ABCD$ parameters of each individual branch, that is:

$$
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_e = \begin{bmatrix}
1 & 0 \\
\left(Z_{in}^{even}\right)^{-1} & 1
\end{bmatrix}_{Z(x)} \begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_e \begin{bmatrix}
1 & 0 \\
\left(Z_{in}^{even}\right)^{-1} & 1
\end{bmatrix}_{Z(x)} (5.1)
$$

$$
\begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_o = \begin{bmatrix}
1 & 0 \\
\left(Z_{in}^{odd}\right)^{-1} & 1
\end{bmatrix}_{Z(x)} \begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_o \begin{bmatrix}
1 & 0 \\
\left(Z_{in}^{odd}\right)^{-1} & 1
\end{bmatrix}_{Z(x)} (5.2)
$$

The $ABCD$ parameters of the non-uniform impedance profiles $Z_{1,2}(x)$ can be determined by following the procedure given in Section 2.1 at each design frequency ‘$f_m$’, taking into account the series in (2.6), where $Z_1$ and $Z_2$ are set to 50 $\Omega$ and 35 $\Omega$, respectively. Microstrip lengths $d_1$ and $d_2$ are chosen to be $\lambda/8$ and $\lambda/4$, respectively, at the 1st (i.e., lowest) design frequency $f_1$. Upon determining the $ABCD$ matrix of $Z_1(x)$, the following equation can be written as:

$$
\begin{bmatrix}
V_1 \\
I_1
\end{bmatrix} = \begin{bmatrix}
A & B \\
C & D
\end{bmatrix}_{Z_1(x)} \begin{bmatrix}
V_2 \\
I_2
\end{bmatrix} (5.3)
$$

In order to obtain $Z_{in}^{even}$, $I_2$ is set to zero. Solving for $\frac{V_1}{I_1}$, one obtains:

$$
\frac{V_1}{I_1} = \frac{A Z_{1(x)}}{C Z_{1(x)}} = Z_{in}^{even} (5.4)
$$

Similarly, the odd-mode input impedance $Z_{in}^{odd}$ shown in Figure 5-2(b) is determined by setting $V_2$ in (5.3) to zero, leading to:

$$
\frac{V_1}{I_1} = \frac{B Z_{1(x)}}{D Z_{1(x)}} = Z_{in}^{odd} (5.5)
$$

Consequently, the $ABCD$ matrices for the circuit modes are calculated using (5.1) and (5.2). The total input impedance for each mode is expressed as follows:
\[ Z_{\text{total}}^{\text{in}} = \frac{A_{e,o}Z_0 + B_{e,o}}{C_{e,o}Z_0 + D_{e,o}} \]  

(5.6)

where \( Z_0 \) is the characteristic impedance of each feed port. Thus, the reflection and transmission coefficients for the NTLs BLC can be written as: \( \Gamma_{e,o} \)

\[ \Gamma_{e,o} = \frac{Z_{\text{total}}^{\text{in}} - Z_0}{Z_{\text{total}}^{\text{in}} + Z_0} \]  

(5.7)

\[ T_{e,o} = \frac{2}{A_{e,o} + \frac{B_{e,o}}{Z_0} + C_{e,o}Z_0 + D_{e,o}} \]  

(5.8)

S-parameters of the BLC are calculated using the following equations:

\[ S_{11} = \frac{\Gamma_{e} + \Gamma_{o}}{2} \]  

(5.9)

\[ S_{41} = \frac{\Gamma_{e} - \Gamma_{o}}{2} \]  

(5.10)

\[ S_{21} = \frac{T_{e} + T_{o}}{2} \]  

(5.11)

\[ S_{31} = \frac{T_{e} - T_{o}}{2} \]  

(5.12)

Finally, in order to obtain the desired response at the design frequencies, the optimum values of the Fourier coefficients (\( a_n \)'s in (2.6)), can be obtained through minimizing the following error function:

\[ E = \sqrt{\frac{\sum_{m=1}^{M} \left( \left| S_{11} \right|^2 + \left| S_{41} \right|^2 + \left( \left| S_{21} \right|^2 - \left| S_{31} \right|_{\text{des}}^2 \right)^2 + \left( \left| S_{31} \right|^2 - \left| S_{31} \right|_{\text{des}}^2 \right)^2 \right)}{16M}} \]  

(5.13)

where \( \left| S_{21} \right|_{\text{des}} = \left| S_{31} \right|_{\text{des}} = 0.707 \). The term “16M” in the denominator acts as a normalization factor. Figure 5-3 shows a flowchart summarizing the design procedure of the proposed multi-frequency BLC.
Figure 5-3: Flowchart showing the design procedure of the multi-frequency non-uniform BLC; green and red enclosures present the theoretical formulation based on even- and odd-mode equivalent transmission line circuits, respectively.
5.1.1 Dual-Frequency Example

A dual-frequency NTLs BLC with design frequencies chosen to be 0.9 GHz and 2.4 GHz is presented. The 0.9 GHz frequency band is widely used in GSM technology; whereas the 2.4 GHz band fits in many wireless applications, such as IEEE 802.11b,g,n standards (WLAN and/or WiFi).

Taking into account a 1.524-mm-thick Rogers RO4835 substrate with a relative permittivity of 3.48 and a loss tangent of 0.0037, two NTLs, $Z_{1,2}(x)$, with widths bounded by $1 \text{ mm} < W_{1,2}(x) < 10 \text{ mm}$ and lengths $d_1$ and $d_2$ of 25.18 mm and 49.27 mm, respectively, are deployed. The characteristic impedances $Z_1$, $Z_2$, and $Z_0$ are chosen to be 50, 35, and 50 Ω, respectively. $K_1$, $K_2$, and $N$ are set to 50, 50, and 10, respectively. The optimization is performed in 2000 iterations, with a resulting error value of 0.022. Table 5.1 shows the obtained Fourier series coefficients for $Z_{1,2}(x)$.

### Table 5.1: NTL coefficients of the dual-band BLC.

<table>
<thead>
<tr>
<th>Fourier coefficients for $Z_1(x)$</th>
<th>$a_0$</th>
<th>$a_1$</th>
<th>$a_2$</th>
<th>$a_3$</th>
<th>$a_4$</th>
<th>$a_5$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$a_6$</td>
<td>0.0534</td>
<td>-0.0923</td>
<td>0.0102</td>
<td>0.0057</td>
<td>0.0043</td>
<td>0.0036</td>
</tr>
<tr>
<td>$a_7$</td>
<td>0.0033</td>
<td>0.0031</td>
<td>0.0030</td>
<td>0.0029</td>
<td>0.0028</td>
<td>-</td>
</tr>
<tr>
<td>$a_8$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$a_9$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$a_{10}$</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

### Fourier coefficients for $Z_2(x)$

<table>
<thead>
<tr>
<th>$c_6$</th>
<th>$c_7$</th>
<th>$c_8$</th>
<th>$c_9$</th>
<th>$c_{10}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$a_0$</td>
<td>-0.0370</td>
<td>-0.1661</td>
<td>-0.3631</td>
<td>0.3822</td>
</tr>
<tr>
<td>$a_1$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$a_2$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$a_3$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$a_4$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$a_5$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$c_6$</td>
<td>0.0193</td>
<td>0.0112</td>
<td>0.0044</td>
<td>0.0016</td>
</tr>
<tr>
<td>$c_7$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$c_8$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$c_9$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>$c_{10}$</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Figure 5-4 shows the full-wave simulated and measured results of the dual-band BLC. $S_{11}$ is below –20 dB and –18 dB at 0.83 GHz and 2.4 GHz, respectively, and the obtained experimental results are in a good agreement with simulations. The isolation parameter ($S_{41}$) is also below –20 dB at the design frequencies. The simulated through parameter ($S_{21}$) equals to –2.9 dB and –2.7 dB at the first and second bands, respectively.
which are very close to their theoretical value of –3 dB. The results obtained from measurement are around –3.4 dB. The simulated coupled parameter \((S_{31})\) equals to –3.4 dB at 0.9 GHz and 2.4 GHz. Such values are also close to –3 dB; whereas the measured results are –3.5 dB in proximity to the two design frequencies. The slight discrepancies between the simulated and measured results are thought to be due to connector losses as well as measurement errors. Figure 5-4(b) shows the simulated and measured phase difference between the through and coupled parameters. A quadrature phase difference occurs at 0.9 GHz and 2.4 GHz; specifically, 90° and 270°, respectively.

![Figure 5-4: Simulated and measured results of the dual-frequency BLC: (a) S-parameters magnitude; (b) phase difference between \(S_{21}\) and \(S_{31}\).](image)
5.1.2 Triple-Frequency Example

After the successful implementation of a dual-frequency NTLs BLC, a triple-band coupler is implemented in a similar fashion to prove the validity, repeatability, and robustness of the underlying design procedure. The proposed triple-band NTLs BLC is designed to operate at three concurrent frequencies, specifically, 0.9 GHz, 2.4 GHz, and 5.4 GHz. Such bands find many applications in GSM, WLAN, Wi-Fi, and WiMAX technologies. Two NTLs, $Z_{1,2}(x)$ with widths bounded by $1 \text{ mm} < W_{1,2}(x) < 10 \text{ mm}$ are designed with lengths $d_1$ and $d_2$ of 25.18 mm and 49.27 mm, respectively, which equal to $\lambda/8$ and $\lambda/4$ at the lowest design frequency (i.e., 0.9 GHz). The characteristic impedances $Z_1$, $Z_2$, and $Z_0$ are chosen to be 50, 35, and 50 $\Omega$, respectively. $K_1$, $K_2$, and $N$ are set to 50, 50, and 10, respectively. The optimization is performed in 3000 iterations. The resulting error value was 0.026. Table 5.2 shows the obtained Fourier series coefficients for $Z_{1,2}(x)$.

<table>
<thead>
<tr>
<th>Fourier Coefficients for $Z_1(x)$</th>
<th>$a_0$</th>
<th>$a_1$</th>
<th>$a_2$</th>
<th>$a_3$</th>
<th>$a_4$</th>
<th>$a_5$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$d_0$</td>
<td>0.1782</td>
<td>0.1954</td>
<td>0.2223</td>
<td>-0.2163</td>
<td>-0.0881</td>
<td>-0.0589</td>
</tr>
<tr>
<td>$d_6$</td>
<td>-0.0540</td>
<td>-0.0487</td>
<td>-0.0450</td>
<td>-0.0432</td>
<td>-0.0418</td>
<td>-</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Fourier Coefficients for $Z_2(x)$</th>
<th>$a_0$</th>
<th>$a_1$</th>
<th>$a_2$</th>
<th>$a_3$</th>
<th>$a_4$</th>
<th>$a_5$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$a_0$</td>
<td>0.1680</td>
<td>0.1112</td>
<td>0.3284</td>
<td>-0.0909</td>
<td>0.2555</td>
<td>-0.2732</td>
</tr>
<tr>
<td>$d_6$</td>
<td>-0.1565</td>
<td>-0.0920</td>
<td>-0.0933</td>
<td>-0.0783</td>
<td>-0.0790</td>
<td>-</td>
</tr>
</tbody>
</table>

Figure 5-5(a) shows the simulated and measured $S$-parameters of the triple-band BLC. $S_{11}$ and $S_{41}$ are both below $-20$ dB at the design frequencies, and in a well agreement with the measured results. The simulated $S_{21}$ and $S_{31}$ are in the ranges of $-2.5$ dB to $-4$ dB; whereas the measured results are in the ranges of $-3.1$ dB to $-4.6$ dB at the
three design bands. The slight frequency shifts, as well as the increased losses are mainly due to the resulting optimization error, different types of losses, and measurement errors. Figure 5-5(b) illustrates the simulated and measured phase difference between the two output ports. A quadrature phase difference occurs around the design frequencies (i.e. 0.9 GHz, 2.4 GHz, and 5.4 GHz).

Figure 5-5: Simulated and measured results of the triple-frequency BLC: (a) S-parameters magnitude; (b) phase difference between $S_{21}$ and $S_{31}$. 
5.2 Broadband Branch-Line Coupler

5.2.1 Design

Figure 5-6 shows a schematic layout of the proposed coupler. Figure 5-7 depicts the even-odd mode circuits, characterized by non-uniform impedances $Z_i(x)$ ($i = 1, 2, 3$), propagation constants $\beta_i(x)$, and lengths $d_i$.

Figure 5-6: Schematic diagram of the proposed broadband BLC. The dashed blue box represents the portion where the even-odd mode analysis is carried out.

Figure 5-7: Even-odd mode circuit outlines of the proposed impedance-varying broadband BLC: (a) even-even; (b) even-odd; (c) odd-even; (d) odd-odd.
The mathematical formulations of the AB CD matrix of \( Z(x) \) are obtained as described in (2.1–2.5), Section 2.1. Upon determining the AB CD parameters of \( Z_{1,2,3}(x) \), \( Z_{w(1,2)}^{e} \) and \( Z_{w(2,1)}^{o} \) are calculated. We first consider the even-mode input impedances \( Z_{w(1,2)}^{e} \), where \( Z_{1,2}(x) \) are terminated by an open-circuit. Referring to Figure 5-7, and upon determining the overall AB CD matrix of \( Z_{1,2}(x) \), the following equation can be written:

\[
\begin{bmatrix}
V_{1,2}' \\
I_{1,2}'
\end{bmatrix} = \begin{bmatrix} A & B \\
C & D \end{bmatrix} \begin{bmatrix}
V_{1,2}'' \\
I_{1,2}''
\end{bmatrix}
\]  

(5.14)

which leads to:

\[
V_{1,2}' = AV_{1,2}'' + BI_{1,2}''
\]  

(5.15)

\[
I_{1,2}' = CV_{1,2}'' + DI_{1,2}''
\]  

(5.16)

Setting \( I_{1,2}' \) in (5.14) to zero, and solving for \( V_{1,2}'/I_{1,2}' \), one gets:

\[
\frac{V_{1,2}'}{I_{1,2}'} = \left(\frac{A}{C}\right)_{Z_{1,2}(x)} = Z_{w(1,2)}^{e}
\]  

(5.17)

Similarly, the odd-mode input impedance \( Z_{w(2,1)}^{o} \) is determined for \( Z_{1,2}(x) \) with short-circuit terminations by setting \( V_{1,2}'' \) in (5.14) to zero, leading to:

\[
\frac{V_{1,2}'}{I_{1,2}'} = \left(\frac{B}{D}\right)_{Z_{1,2}(x)} = Z_{w(1,2)}^{o}
\]  

(5.18)

\( Z_{w(1,2)}^{ee,oe,oo} \) seen before \( Z_{3}(x) \) are then the parallel combinations \( Z_{w(1,2)}^{e,o}/Z_{w(1,2)}^{e,o} \). The total even-odd mode input impedances of the entire network are expressed as:

\[
Z_{w(1,2)}^{ee,oe,oo} \bigg|_{in}^{(T)} = \frac{A Z_{3}(x) Z_{w(1,2)}^{ee,oe,oo} + B Z_{2}(x)}{C Z_{3}(x) Z_{w(1,2)}^{ee,oe,oo} + D Z_{2}(x)}
\]  

(5.19)

which, once determined, the reflection coefficient \( \Gamma \) can be calculated as follows:

\[
\Gamma_{ee,oe,oo} = \frac{Z_{w(1,2)}^{ee,oe,oo}}{Z_{w(1,2)}^{ee,oe,oo}} - Z_{0}
\]  

(5.20)
where \( Z_0 \) is the port characteristic impedance. \( S \)-parameters at each frequency \( f \in [f_l, f_h] \) are determined using the reflection coefficients found in (5.20):

\[
\begin{align*}
S_{11} &= \frac{\Gamma_{ee} + \Gamma_{eo} + \Gamma_{oe} + \Gamma_{oo}}{4} \quad (5.21) \\
S_{21} &= \frac{\Gamma_{ee} - \Gamma_{eo} + \Gamma_{oe} - \Gamma_{oo}}{4} \quad (5.22) \\
S_{31} &= \frac{\Gamma_{ee} - \Gamma_{eo} - \Gamma_{oe} + \Gamma_{oo}}{4} \quad (5.23) \\
S_{41} &= \frac{\Gamma_{ee} + \Gamma_{eo} - \Gamma_{oe} - \Gamma_{oo}}{4} \quad (5.24)
\end{align*}
\]

Then, the error at each frequency \( f \) is defined as:

\[
E = \left( |S_{11}|^2 + |S_{41}|^2 + |S_{21}-\sqrt{1-C^2}|^2 + |S_{31}-C|^2 \right) \frac{f_h-f_l}{\Delta f} \right)^{0.5}
\]

subject to:

\[
\angle S_{21} - \angle S_{31} = \frac{\pi}{2}
\]

where \( C = 10^{-C_{(\text{dB})}/20} \) is the desired coupling level, and \( \Delta f \) is a frequency increment. The error vector resulting from applying (5.25) to all frequency points within \( f_l \) and \( f_h \) is used to formulate and minimize the following objective function:

\[
\text{Objective} = \sum_{p=0}^{(f_h-f_l)/\Delta f} E(f_i + p\Delta f)
\]

Subject to the constraints mentioned in (2.8–2.9) for matching purposes and impedance confinement within minimum and maximum widths so that fabrication limitations are not exceeded. The trust-region-reflective algorithm is used to solve this constrained non-linear minimization problem. The general design steps to realize the proposed broadband impedance-varying BLC with arbitrary coupling levels are summarized as follows:
**Step 1:** $Z_{1,2,3}(x)$ are subdivided into uniform electrically short segments of fixed lengths, and the $ABCD$ parameters of each segment are calculated taking into account Fourier-based impedance profiles expressed by (2.5).

**Step 2:** Overall $ABCD$ matrices of $Z_{1,2}(x)$ are utilized in (5.17) and (5.18) to obtain the even-odd mode impedances $Z_{in}^{e_{1,2}}$, $Z_{in}^{o_{1,2}}$.

**Step 3:** Resulting impedances from **Step 2** and the $ABCD$ matrix of $Z_3(x)$ are used in (5.19) to calculate the total even-odd impedances $Z_{in}^{ee, eo, oe, oo}$ at each port of the BLC.

**Step 4:** A $2 \times 2$ reflection coefficient matrix representing the reflection coefficients of the four analysis modes is calculated in (5.20) utilizing the impedances obtained from **Step 3**. The reflection coefficient matrix is incorporated to formulate the scattering parameters according to (5.21–5.24).

**Step 5:** Scattering parameters in **Step 4** at each frequency $f$ within $[f_l, f_h]$ along with the desired coupling level ($C$) are conveyed in the error function given in (5.25).

**Step 6:** Sum of errors, expressed by (5.27), at all frequencies within the design bandwidth are minimized such that Fourier series coefficients of $Z_{1,2,3}(x)$ are set as optimization variables subject to constraints (2.8), (2.9), and (5.26).

Figure 5-8 shows a pseudocode that describes the broadband BLC design procedure.
Algorithm: Broadband Impedance-Varying BLC Design

Given: \([\varepsilon_r, h]\) - Substrate Parameters; 
\([d_1, d_2, d_3]\) - \(Z_{1,2,3}(x)\) Lengths; 
\([Z_1, Z_2, Z_3]\) - Reference Impedances; 
\([Z_{min}, Z_{max}]\) - \(Z_{1,2,3}(x)\) Constraints; 
\([f_l, f_h, \Delta f]\) - Frequency Range and Step; 
\(C\) - Coupling Level; 
\(K\) - No. of Uniform Segments; 
\(N\) - No. of Fourier Series Terms;

1: **Procedure** Broadband\_NonUniform\_BLC()
2: **Loop**: for each frequency do
3: for each \(i\) impedance do
4: \(\Delta x_i = d_i / K;\)
5: for each \(j\) segment do
6: \([A B; C D] = ABCD\_Matrix();\) // initial coefficients assumed
7: end for
8: \([A B; C D]_i = Overall\_ABCD\_Matrix();\) // \([\cdot]\) denotes the \(ABCD\)
9: // matrix of the \(i^{th}\) impedance
10: \([Z_{in}^{(1,2)}, Z_{in}^{(1,2)}]\) = EvenOdd\_Imped\_of\_Z\_1,2(x)([A B; C D]_1,[A B; C D]_2);
11: \([Z_{in}^{ee,oe,oo,oo}]\) = \(Z_{in}^{ee,oe,oo,oo}(1) / Z_{in}^{ee,oe,oo,oo}(2)\)\([Z_{in}^{(1,2)}, Z_{in}^{(1,2)}]\));
12: \([Z_{in}^{T}][ee,oe,oo,oo]\) = Total\_EvenOdd\_Imped([Z_{in}^{ee,oe,oe,oo}], [A B; C D]_3);
13: \([\Gamma_{ee,oe,oe,oo}]\) = Reflection\_Coefficients([Z_{in}^{(T)}][ee,oe,oe,oo], Z_0);
14: \([S_{11}, S_{21}, S_{31}, S_{41}]\) = \(S\)-Parameters([\(\Gamma_{ee,oe,oe,oo}\)]);
15: \([E]\) = Set\_Error\_Value([S_{11}, S_{21}, S_{31}, S_{41}],[f_l, f_h, \Delta f], C);
16: end for
17: \([c_0, [a_1, \ldots, a_N], [b_1, \ldots, b_N], Objective] =\) Minimize\_Sum\_of\_Errors([E]);
18: **Repeat Loop** until optimal \([c_0, [a_1, \ldots, a_N], [b_1, \ldots, b_N]]\);
19: **end Procedure**

Figure 5-8: Pseudocode of the proposed broadband impedance-varying BLC.
5.2.2 Analytical Results

Three design examples of 3-dB, 6-dB, and 9-dB broadband BLCs built with 0.813-mm-thick Rogers RO4003C substrate with a relative permittivity of 3.55 and dielectric loss tangent of 0.0027 are presented. The operating band is selected such that $f_l = 2.15$ GHz, $f_h = 3.85$ GHz, and $\Delta f = 0.1$ GHz. For a compact BLC design, the lengths $d_{1,2}$ are chosen to be $\lambda/8$; while $d_3 = \lambda/4$, all at center frequency of 3 GHz. The reference impedance $Z_1$ is set to 50 Ω, and $Z_2 = Z_3 = 35$ Ω. Such values lead to predefined reasonable width terminations for $Z_{1,2,3}(x)$. Otherwise, $Z_i(0)$ and $Z_i(d)$ will arbitrarily be allocated by the optimization process, which may cause impractical widths at both ends of $Z_i(x)$. $Z_3$ and $Z_2$ are given equal values to avoid discontinuities at the four junctions of the proposed BLC. The numbers of the uniform segments $K$ and Fourier terms $N$ for $Z_{1,2,3}(x)$ are set to 25 and 5, respectively, which are sufficient to achieve the required optimization goals. $Z_{1,2,3}(x)$ are bounded by minimum and maximum impedance values to ensure physical realization within fabrication limits. In other words, $Z_i(x) \in [21, 128]$ Ω, which correspond to widths variations between 0.2 and 6 mm. The minimization of the objective function in (5.27) is performed in 3000 iterations using Matlab.

Figure 5-9 shows the resulting impedance profiles for the proposed designs. All non-uniform profiles in each BLC example are constrained by the previously mentioned impedance interval. Furthermore, all optimized transmission lines follow (2.9). In other words, $Z_1(0) = Z_1(d) = 50$ Ω; whereas $Z_2(0) = Z_2(d) = Z_3(0) = Z_3(d) = 35$ Ω. It is paramount to point out that the almost flat variation of $Z_2(x)$ across $d_2$ shows that it has the lowest effect on the overall performance in similar structures (i.e., quadrature BLCs with extended output ports) as compared with $Z_{1,3}(x)$. 

60
Figure 5-9: Variations as a function of length: (a) 3-dB; (b) 6-dB; (c) 9-dB broadband BLCs. Solid, dotted, and dashed lines represent $Z_1(x)$, $Z_2(x)$, and $Z_3(x)$, respectively.

Figure 5-10 represents the analytical response of the proposed 3-dB, 6-dB, and 9-dB broadband BLCs over a frequency range normalized to 3 GHz. We begin our discussions taking into account a maximum of ±1 dB and ±5° amplitude and phase imbalances, respectively, and –10 dB impedance matching and isolation [47]. Theoretical benchmarks of the magnitude of the through ($S_{21}$) and coupled ($S_{31}$) parameters are indicated in Table 5.3 for comparison purposes.
Figure 5-10: Analytical response of the proposed broadband BLCs with different values of $C$. Magnitudes of $S$-parameters for: (a) $C = 3$-dB; (b) $C = 6$-dB; (c) $C = 9$-dB. Phase difference between the through and coupled ports for: (d) $C = 3$-dB; (e) $C = 6$-dB; (f) $C = 9$-dB.

Table 5.3: Theoretical values of the through and coupled parameters.

| Coupling Level | $-|S_{21}|$, $-|S_{31}|$ (dB) |
|----------------|-------------------------------|
| 3-dB           | 3, 3                          |
| 6-dB           | 1.25, 6                       |
| 9-dB           | 0.58, 9                       |

The three designs have the magnitude of $S_{11}$ and $S_{41}$ better than $-10$ dB over a fractional bandwidth of 57%, that is, from $f/f_c = 0.72$ to 1.29. Furthermore, the magnitudes of $S_{21}$ and $S_{31}$ equal to $-3 \pm 0.5$ dB for the 3-dB BLC (Figure 5-10(a)), and are $-1.25 \pm 0.5$ dB and $-6 \pm 0.5$ dB, respectively, for the 6-dB BLC (Figure 5-10(b)), and equal to $-0.58 \pm 0.5$ dB and $-9 \pm 0.5$ dB, respectively, for the 9-dB BLC (Figure 5-10(c)) across the design frequency spectrum.
The phase difference, $\angle S_{21} - \angle S_{31}$, of the proposed couplers is plotted in Figure 5-10(d-f). As shown in such responses, the difference is almost constant and equals $90^\circ$ over the predefined bandwidth with $\pm3^\circ$ phase imbalance in all design examples. Hence, the proposed optimization-driven framework demonstrates an excellent performance over a broad frequency range. The proposed methodology differs from other previous efforts in the following aspects: 1) Unlike [39], [43]-[46], and [50], all BLC designs presented in this work are planar, with controllable minimum and maximum impedances. Impedance variations lead to flexible bandwidth allocation and coupling levels within the allowable fabrication tolerances. 2) Based on what was presented in [49, Fig. 3], it is not possible to realize the impedances of a broadband 3-dB BLC with conventional uniform transmission lines, and the maximum ‘theoretical’ bandwidth this approach can achieve is around 30% for 3-dB coupling [49, Fig. 4]. Our technique, however, achieves almost twice the bandwidth by modulating the impedances of all BLC branches. 3) Higher order harmonics are suppressed in the proposed technique, since the electrical characteristics of the broadband BLC are enforced to match the required performance only across a specific frequency band. In contrast, other designs that depend on port extensions via uniform $\lambda/4$, $\lambda/2$, or their equivalent coupled transmission lines suffer from the presence of harmonics at each odd multiple of the center frequency [42]-[43], [48]-[49]. Thus, more area is needed for broadband clean-up filters.

Figure 5-11 shows the calculated response across a wide frequency range for a 6-dB BLC based on the equations presented in [49] and the counterpart proposed non-uniform coupler. Both designs have the same center frequency and transmission line lengths. Thus, an occupied circuit area (in terms of $\lambda$) of $\lambda/4 \times 3\lambda/4$ is obtained.
As shown in Figure 5-11, the adoption of uniform microstrip lines results in $S$-parameters exhibiting broadband characteristics at 3 GHz and 9 GHz (Figure 5-11(a)). However, the proposed methodology shows that harmonics at 9 GHz are completely suppressed (Figure 5-11(b)). Furthermore, the phase difference between the coupled and through ports maintained a quadrature response (i.e., $90^\circ$ and $270^\circ$) at 3 GHz and 9 GHz in the case of utilizing conventional microstrip lines (Figure 5-11(c)). On the other hand, the proposed technique shows a constant phase difference of $90^\circ$ only at the predefined design band. Table 5.4 shows the resulting Fourier series coefficients of the non-uniform impedances $Z_i(x)$ for the proposed 3-dB, 6-dB, and 9-dB broadband BLCs.
Table 5.4: Fourier coefficients of the impedances of the three couplers.

<table>
<thead>
<tr>
<th>Coupling level</th>
<th>$Z_i(x)$</th>
<th>$c_0$</th>
<th>$a_1$</th>
<th>$a_2$</th>
<th>$a_3$</th>
<th>$a_4$</th>
<th>$b_1$</th>
<th>$b_2$</th>
<th>$b_3$</th>
<th>$b_4$</th>
<th>$b_5$</th>
<th>Error in (5.29)</th>
</tr>
</thead>
<tbody>
<tr>
<td>3-dB</td>
<td>$Z_1$</td>
<td>-0.0101</td>
<td>0.1239</td>
<td>-0.0536</td>
<td>0.1129</td>
<td>0.0240</td>
<td>-0.0475</td>
<td>0.0083</td>
<td>-0.2035</td>
<td>0.1679</td>
<td>-0.0708</td>
<td>0.4252</td>
</tr>
<tr>
<td></td>
<td>$Z_2$</td>
<td>-0.1790</td>
<td>0.0648</td>
<td>-0.0082</td>
<td>0.0511</td>
<td>-0.0134</td>
<td>0.0353</td>
<td>-0.0139</td>
<td>0.0201</td>
<td>-0.0110</td>
<td>0.0077</td>
<td>-0.0056</td>
</tr>
<tr>
<td></td>
<td>$Z_3$</td>
<td>0.4464</td>
<td>0.6381</td>
<td>0.0517</td>
<td>-0.6419</td>
<td>-0.0110</td>
<td>-0.2684</td>
<td>-0.0102</td>
<td>-0.0432</td>
<td>0.0342</td>
<td>-0.1310</td>
<td>0.0119</td>
</tr>
<tr>
<td>6-dB</td>
<td>$Z_1$</td>
<td>0.3495</td>
<td>-0.1677</td>
<td>0.1348</td>
<td>-0.1179</td>
<td>-0.2295</td>
<td>-0.0511</td>
<td>0.3302</td>
<td>0.2269</td>
<td>0.0798</td>
<td>-0.2396</td>
<td>-0.0014</td>
</tr>
<tr>
<td></td>
<td>$Z_2$</td>
<td>-0.1739</td>
<td>0.0600</td>
<td>-0.0161</td>
<td>0.0458</td>
<td>-0.0171</td>
<td>0.0326</td>
<td>-0.0227</td>
<td>0.0263</td>
<td>-0.0207</td>
<td>0.0091</td>
<td>-0.0181</td>
</tr>
<tr>
<td></td>
<td>$Z_3$</td>
<td>0.3104</td>
<td>0.5142</td>
<td>0.0107</td>
<td>-0.0772</td>
<td>0.0002</td>
<td>-0.3391</td>
<td>-0.0467</td>
<td>-0.0766</td>
<td>0.0204</td>
<td>-0.3317</td>
<td>-0.1218</td>
</tr>
<tr>
<td>9-dB</td>
<td>$Z_1$</td>
<td>0.6116</td>
<td>0.0426</td>
<td>0.1852</td>
<td>-0.2627</td>
<td>-0.2905</td>
<td>-0.2752</td>
<td>0.1603</td>
<td>0.0439</td>
<td>-0.0754</td>
<td>-0.1601</td>
<td>-0.0341</td>
</tr>
<tr>
<td></td>
<td>$Z_2$</td>
<td>-0.1791</td>
<td>0.0622</td>
<td>-0.0039</td>
<td>0.0518</td>
<td>-0.0049</td>
<td>0.0351</td>
<td>-0.0074</td>
<td>0.0199</td>
<td>-0.0036</td>
<td>0.0101</td>
<td>-0.0025</td>
</tr>
<tr>
<td></td>
<td>$Z_3$</td>
<td>0.2421</td>
<td>0.4240</td>
<td>0.2570</td>
<td>-0.3540</td>
<td>-0.2568</td>
<td>-0.1643</td>
<td>-0.3107</td>
<td>-0.0951</td>
<td>-0.2302</td>
<td>-0.0526</td>
<td>-0.2672</td>
</tr>
</tbody>
</table>
All coefficients are within the interval \([-1,1]\). It is noteworthy to point out that there is no unique solution for the unknown Fourier series coefficients. In other words, each optimization attempt results in different sets of coefficients \(c_0, a_n, \) and \(b_n\). However, the optimum electrical response adjoined with an impedance profile \(Z_i(x)\) that follows the constraints is considered in the next design steps. The optimization error reduces as the coupling level increases; indicating that better characteristics (matching, isolation, etc.) are achieved for BLCs with higher coupling levels.

5.2.3 Simulations and Measurements

Full-wave EM simulated and measured results are presented and discussed for three BLCs: 3-dB, 6-dB, and 9-dB designed in Section 5.2.2. Lengths \(d_1 = d_2 = 7.5\) mm and \(d_3 = 15\) mm. Figure 5-12 shows photographs of the fabricated designs built with the Rogers RO4003C substrate mentioned earlier.

![Photographs of the fabricated BLCs](a) 3-dB; (b) 6-dB; (c) 9-dB.

Figure 5-12: Photographs of the fabricated BLCs: (a) 3-dB; (b) 6-dB; (c) 9-dB.

Figure 5-13 shows the simulated and measured results. Simulations were done with the finite element method-based tool ANSYS HFSS. Measurements were made with an HP 8720B VNA. Simulated and measured results showed a positive frequency shift of 50 MHz in the overall response, which could be due to discontinuity effects.
Figure 5-13: Magnitude response of: (a) 3-dB; (b) 6-dB; (c) 9-dB BLCs. Dashed, dotted, solid, and dashed-dotted lines represent the simulated $S_{21}$, $S_{31}$, $S_{11}$, and $S_{41}$, respectively; whereas the plus, star, circle, and cross markers represent the measured $S_{21}$, $S_{31}$, $S_{11}$, and $S_{41}$, respectively.
Figure 5-14: Simulated and measured phase difference between the through and coupled ports: (a) $C = 3$-dB; (b) $C = 6$-dB; (c) $C = 9$-dB.
As seen in Figure 5-13(a), simulated and measured $S$-parameters of the 3-dB BLC are in an excellent agreement, and show that $|S_{21}|$ and $|S_{31}|$ are around $-3 \pm 1$ dB; whereas the input port matching and isolation are below $-10$ dB across 2.2–3.9 GHz. Figure 5-13(b) shows the simulated and measured response of the 6-dB coupler. $|S_{21}|$ and $|S_{31}|$ are close to their theoretical values of $-1.25$ dB and $-6$ dB, respectively; whereas both $|S_{11}|$ and $|S_{41}|$ are below $-10$ dB over the band of interest. Finally, Figure 5-13(c) shows that the simulated and measured $|S_{21}|$ and $|S_{31}|$ of the 9-dB coupler are in proximity to $-0.5$ dB and $-9$ dB, respectively; while $|S_{11}|$ and $|S_{41}|$ are below $-10$ dB across the design band. Better matching/isolation is achieved with the increase in coupling level. The small discrepancies between the analytical response and simulated (or measured) results are thought to be due to conductor and dielectric losses.

Simulated and measured phase differences in all examples, shown in Figure 5-14, are in a very good agreement, and show a constant phase difference of $90^\circ \pm 5^\circ$. Hence, a broadband frequency performance, described by a fractional bandwidth of 57%, is obtained. An efficient suppression of higher-order harmonics is also observed in the simulated and measured three coupler examples.

The concept of impedance-varying transmission lines is further investigated by fixing the coupling level while varying the frequency band $[f_l,f_h]$ to obtain different fractional bandwidths. New optimizations are carried out with $C = 0.5012$, which correspond to 6-dB coupling. All other design parameters mentioned earlier are kept unchanged. Figure 5-15 shows the analytical response of impedance-varying broadband 6-dB BLCs optimized over frequency ranges of 2.5–3.5 GHz, 2.15–3.85 GHz, and 1.9–4.1 GHz. Non-uniform microstrip lines are used to achieve the required bandwidths.
As seen in Figure 5-15(a-b), tolerances from the theoretical value of $|S_{21}|$ are 0.1 dB, 0.3 dB, and 0.5 dB for the 33%, 57%, and 73% fractional bandwidths, respectively; while those from $|S_{31}|$ are 0.3 dB, 0.8 dB, and 1.5, respectively. Besides, Figure 5-15(c-d) illustrates excellent matching and isolation across all design bandwidths.

The phase difference between the through and coupled ports for the three 6-dB BLCs is presented in Figure 5-16. Each example maintained a constant 90° phase difference across the designed bands, justifying the design methodology. Hence, different fractional bandwidths for a specific coupling level are obtained by properly varying the widths of the non-uniform transmission line profiles.
Figure 5-16: Phase differences between through and coupled ports of the impedance-varying broadband 6-dB BLCs optimized for three different fractional bandwidths.

Table 5.5 provides a comparison between the measured results of the proposed broadband impedance-varying BLCs and other state-of-the-art couplers. The adopted technique, coupling level, fractional bandwidth, scattering parameters, and occupied circuit area are set as benchmarks. Bandwidth definitions for the given techniques are different; and thus, cannot be directly compared.

The capacity of the proposed approach in achieving arbitrary coupling levels is illustrated in three different BLC examples (i.e., 3-, 6-, and 9-dB). Each example is designed and measured considering a 57% fractional bandwidth. The proposed methodology has better input port matching and isolation as compared to other reported techniques. The through and coupled parameters for each coupling level are within acceptable tolerances. Resulting designs are planar, and the associating advantages of the underlined principle come at no expense to the occupied circuit area. Broadband filters are also unrequired since the proposed technique results in suppressing higher harmonics. On the other hand, previous studies that depend on conventional $\lambda/4$, $\lambda/2$, and coupled lines suffer from harmonics at each odd multiple of the center frequency.
Table 5.5: Comparison between electrical and physical characteristics of recent broadband branch-line couplers.

| Applied Technique                                      | C  | $f_c$ | Fractional BW (%) | $-|S_{11}|$, $-|S_{41}|$ at $f_c$ (dB) | $-|S_{21}|$ (dB) | Variation in C (dB) | Area (in terms of $\lambda$) |
|--------------------------------------------------------|----|-------|-------------------|-------------------------------------|-----------------|-------------------|---------------------------|
| [39] $\lambda/4$ port extension with $\lambda/2$ open stubs | 3  | 3.9   | 20                | 20, 20                              | 3.4             | 0.05              | $\lambda/4 \times 5\lambda/4$ |
| [40] $\lambda/2$ port extension with $\lambda/2$ open stubs | 3  | 10.5  | 40                | 16, 18                              | 3.2             | 0.2               | $5\lambda/4 \times 5\lambda/4$ |
| [42] Double $\lambda/4$ port extensions                 | 3  | 1     | 30                | 20, 20                              | 3.8             | 0.8               | $\lambda/4 \times 5\lambda/4$ |
| [43] Suspended $\lambda/4$ coupled lines port extensions | 3  | 6     | 49                | 23, 25                              | 3.6             | 0.5               | $\lambda/4 \times 3\lambda/4$ |
| [45] CPW open and short circuited stubs                 | 3  | 3     | 31                | 20, 16                              | 4               | 1.4               | $\lambda/4 \times 3\lambda/4$ |
| [46] Rectangular-coaxial lines & short stubs            | 3  | 39.5  | 41                | 10, 13                              | 4               | 0.8               | $\lambda/4 \times 3\lambda/4$ |
| [48] Double $\lambda/4$ port extensions                 | 6  | 1     | 20                | 20, 25                              | 1.3             | 0.4               | $\lambda/4 \times 5\lambda/4$ |
| [49] Single $\lambda/4$ port extensions                 | 10 | 3     | 50.9              | 22, 28                              | 1               | 0.5               | $\lambda/4 \times 3\lambda/4$ |
| **This Work**                                           |    |       |                   |                                     |                 |                   |                           |
| $\lambda/4$ impedance-varying branches with $\lambda/4$  | 3  | 3     | 57                | 23, 23                              | 2.6             | 0.8               | $\lambda/4 \times 3\lambda/4$ |
| **port extensions**                                     | 6  | 3     | 57                | 30, 30                              | 1.7             | 0.6               | $\lambda/4 \times 3\lambda/4$ |
| **This Work**                                           | 9  | 3     | 57                | 40, 40                              | 0.9             | 0.7               | $\lambda/4 \times 3\lambda/4$ |
5.3 Conclusions

Based on NTLs theory, a new approach for the design of multi-frequency 90° BLCs is investigated to overcome realization difficulties that are often encountered with conventional single-frequency topologies. Each uniform microstrip transmission line in the conventional BLC branches is replaced with NTLs of Fourier-based impedance profiles. First, proper design equations are derived from the even-/odd-mode analysis. Then, the developed equations are solved by adopting an optimization-driven process to achieve the desired response at the design frequencies.

To justify the design principle, dual- and triple-band BLCs suitable for modern wireless applications (i.e., GSM, WLAN, Wi-Fi, and WiMAX) are designed, fabricated, and experimentally tested. The good agreement between both simulated and measured results proves the validity of the design methodology.

This chapter also proposes a novel design procedure to broaden the bandwidth of a single-section 90° BLC with ports extensions. Uniform impedances in the conventional coupler design are replaced with impedance-varying lines through an optimization-driven process based on the even-/odd-mode circuits. As a result of this process, the optimum Fourier series coefficients that meet the given design requirements (i.e., broadband frequency characteristics) are obtained. Different fractional bandwidths for a specific coupling level are achieved by properly designing the impedance profiles. The proposed methodology is advantageous for applications where BLCs with broadband frequency characteristics and low coupling levels are imposed. For verification purposes, three quadrature BLC examples with arbitrary coupling (i.e., 3-dB, 6-dB, and 9-dB) and 57% fractional bandwidth are designed and built.
Simulated and measured results are in a good agreement and show matching and isolation parameters better than –10 dB, and through and coupling parameters close to their theoretical values across the design band. The proposed broadband BLC design concept is systematic and valid for any coupling level. The underlying principle results in compact and planar (i.e., single-layered) structures with effective higher-order harmonics suppression due to enforcing the multi-/broadband functionality at specific predefined frequencies/fractional bandwidths.
Chapter 6

Dual-Band Notch Antipodal Vivaldi Antenna

Researchers all over the globe are in harmony when it comes to the significance of the AVA in the field of UWB communications, due to its wideband frequency matching and directive radiation. Such desirable electrical characteristics encourage its utilization in several applications, including medical microwave imaging and radar telemetry. However, the UWB matching nature of the AVA induces cross-interference to the existing telecommunication technologies; and thus, negatively impact their functions.

In this chapter, a double narrowband-notch UWB AVA is proposed based on compact mushroom-like EBG structures. First, an AVA is designed and optimized to operate over an UWB spectrum. Then, two pairs of EBG cells are introduced along the antenna feed-line to suppress the frequency components at WiMAX and ISM bands. This simple yet effective approach eliminates the need to disfigure the antenna radiators with slots/parasitic elements or comprise multilayer substrates.

This chapter is organized as follows: Section 6.1 presents the proposed antenna configuration. Then, the carried out performance assessment of the underlined method is discussed in Section 6.2. Simulated and measured results of a fabricated prototype are elaborated in Section 6.3. Finally, conclusions and remarks are provided in Section 6.4.
6.1 Antenna Configuration

A schematic diagram of the proposed antenna layout along with the associating dimensions is illustrated in Figure 6-1. Such dimensions are based on a 0.813-mm-thick Rogers RO4003C substrate with a relative permittivity and loss tangent of 3.55 and 0.0027, respectively. The microstrip-fed input has a characteristic impedance of 50 Ω.

In this design, two pairs of mushroom-like EBG cells surround the antenna feedline. The frequency notches \( f_i = 1/2\pi\sqrt{L_iC_i} \), where \( i = 1, 2 \), are fundamentally due to the inductance \( L_i \) that results from the current flowing through the vias, and the capacitance \( C_i \) established from the gap between the cells’ top patches and the ground plane. The AVA’s flares are of an elliptical taper with design equations derived in [92]. In order to obtain the desired functionality, an AVA that covers the 3.1–10.6 GHz frequency spectrum is first designed. Then, the EBG pairs are incorporated, one at a time, to obtain the notch characteristics (e.g., location, rejection level) based on parametric studies performed with ANSYS HFSS full-wave EM simulation tool.

![Figure 6-1: Proposed dual-band notched AVA; black and gray strips refer to upper and lower flares, respectively.](image)

**Design variables (in mm):**

- \( W \) (substrate width) 66.3
- \( L \) (substrate length) 66.3
- \( w_f \) (feedline width) 2.7
- \( w_{m1} \) (EBG1 width) 9.3
- \( w_{m2} \) (EBG2 width) 7
- \( l_f \) (max. flare width) 47.4
- \( l_g \) (ground length) 18
- \( r_u \) (via radius) 0.4
- \( r_i \) (via radius) 0.3
- \( s_f \) (EBG1 to feedline) 0.2
- \( s_f \) (EBG2 to feedline) 0.4
- \( d_f \) (EBG1 to EBG2) 0.45
6.2 Performance Analysis

The proposed design is analyzed to demonstrate its capability in controlling the notches locations by modifying the parameters of each EBG pair. For the sake of brevity, the lower pair (EBG₁) is considered in this study. Though, the same conclusions hold for the upper pair (EBG₂). Figure 6-2 depicts the notch characteristics in the case of utilizing the lower pair versus a single EBG cell (the one either on right or left).

Incorporating two EBG cells around the feed-line increases the notch bandwidth by 10% as compared to one cell. Moreover, the rejection level in the former is higher. Figure 6-3 shows the effect of changing the radius \( r_l \), width \( w_{m1} \), and separation distance \( s_l \) on the notch location. Increasing \( r_l \) (Figure 6-3(a)) reduces the inductance \( L_1 \) [93, eq. 1]. Thus, a positive shift occurs in \( f_1 \). Similarly, increasing \( r_u \) reduces \( L_2 \), which results in an increase of \( f_2 \). On the other hand, increasing \( w_{m1} \) (Figure 6-3(b)) increases \( C_1 \) which reduces \( f_1 \). Alike \( w_{m1} \), increasing \( w_{m2} \) increases \( C_2 \), which in turn reduces \( f_2 \). Figure 6-3(c) shows the effect of varying \( s_l \) on the antenna response. The closer EBGᵢ to the feed-line, the sharper the notch \( f_i \) will be due to the increased coupling between EBGs’ patches and the feed-line, with no significant effect on the notches positions.
Figure 6-3: Effect of changing EBG₁ (a) radius \( r₁ \); (b) width \( w_{m₁} \); (c) separation \( s₁ \).

Figure 6-4 depicts the minor influence of varying \( d_s \) separating the two EBG pairs on the resulting VSWR, justifying the negligible cross-coupling among both pairs, EBG₁ and EBG₂. The same concept of EBG cells was previously applied to introduce frequency notches in UWB monopole antennas [94]-[96].
Figure 6-4: VSWR simulations for four different $d_s$ values.

Figure 6-5 illustrates the VSWR for four different simulation studies. First, an AVA is optimized to operate over the UWB frequency range. Then, two pairs of EBGs, lower and upper, are incorporated in the design - one pair at a time - to achieve a frequency notch at the 3.6–3.9 and 5.6–5.8 GHz bands, respectively. Finally, the antenna is simulated utilizing two EBG pairs (considering the dimensions reported in Figure 6-1) to obtain the two predefined stopbands.

Figure 6-5: VSWR simulation results for four different scenarios.
As can be noticed, the conventional design, without EBG cells, shows a VSWR < 2 in the frequency range 3.1–10.6 GHz. On the other hand, incorporating only the lower EBG pair results in a VSWR < 2 over the UWB range except for the 3.6–3.9 GHz band (VSWR = 5.8). Similarly, the upper EBG pair produces an UWB response except for the 5.6–5.8 GHz band, which possesses a VSWR of 5.4. Finally, concatenating the two EBG pairs generates two simultaneous notches at the 3.6–3.9 and 5.6–5.8 GHz frequencies with VSWR values of 5.8 and 5.4, respectively, and less than 2 elsewhere. Hence, the easiness of controlling each notch without affecting the other is achieved owing to the low cross-coupling between the incorporated EBG elements. It has to be pointed out that although AVA flares have a bulky size; EBG cells with electrically small dimensions are more than enough to introduce high-reject bands.

The current distribution of the antenna is depicted in Figure 6-6. As can be seen, the lower EBG pair is activated around 3.8 GHz, while the upper one is activated at 5.7 GHz creating band notches (i.e., bandgaps) at these frequencies.

![Figure 6-6: Current distribution of the proposed dual-notch AVA at frequencies: (a) 3.8 GHz; and (b) 5.7 GHz.](image)
6.3 Simulations and Measurements

The measured VSWR, radiation patterns, peak gain, and group delay of a fabricated AVA prototype with band-notch characteristics at 3.6–3.9 and 5.6–5.8 GHz are presented and compared with those obtained by simulations. The VSWR is measured after a two-port calibration to a Rhode & Schwarz ZVB20 VNA, and is illustrated in Figure 6-7. Simulated and measured results are in a good agreement with a clear frequency-reject performance at the intended bands. The discrepancies between both results are thought to be due to fabrication tolerances. Figure 6-8 shows the measured conventional and proposed AVAs gain over the UWB spectrum using two identical antennas separated by a distance of \( d = 1.25 \) meters. The measured transmission coefficient is applied to calculate the antenna gain utilizing the equation [97]:

\[
|S_{21}|^2 = G_T G_R \frac{\lambda}{(4\pi d)^2}
\]  

(6.1)

where \( G_T \) and \( G_R \) are the gains of the transmitter and receiver, respectively, and \( \lambda \) is the free space wavelength in meters. As shown in Figure 6-8, an excellent gain suppression of 7 dB and 5.4 dB is obtained at the first and second notches, respectively.

![Simulated and measured VSWRs of the proposed AVA.](image)
Simulated and measured far-field radiation patterns of the proposed antenna at different frequencies are shown in Figure 6-9. As can be seen in Figure 6-9(a-c), the proposed AVA maintained its directive radiation as their corresponding frequencies (i.e., 5, 7, and 9 GHz) are distant from the notches locations. However, clear pattern distortion and gain attenuation appear in Figure 6-9(d and e) as they express the antenna radiation at 3.8 and 5.7 GHz, respectively, (i.e., within the reject-bands). Figure 6-10 illustrates the measured group delay of the proposed AVA. To measure such a parameter, two identical antennas were placed 1.25 meter apart, and $S_{21}$ is recorded with a suitable frequency step size $f_i$. Finally, the group delay ($\tau$) is calculated by the following equation [98]:

$$\tau = \frac{\Delta \theta}{360\Delta f}$$

(6.2)

where $\Delta \theta = \theta_f - \theta_{f-1}$ is $S_{21}$ phase difference between ($\theta_f, \theta_{f-1}$), and $\Delta f = f_i - f_{i-1}$. Measured $\tau$ is almost flat over the UWB range (around 5 ns), which reflects an acceptable linearity between phase and frequency components for the whole band except the two notches showing $\tau$ of 0.1 ns and 0.9 ns at $f_1$ and $f_2$, respectively. The small $\tau$ fluctuation elsewhere is mainly due to various measurement dispersion mechanisms (e.g., cable dispersion).
Figure 6-9: Proposed dual-notched AVA radiation patterns: (a) 5 GHz, (b) 7 GHz, (c) 9 GHz, (d) 3.8 GHz; center frequency of the 1st notch, and (e) 5.7 GHz; center frequency of the 2nd notch.

Figure 6-10: Group delay of the proposed dual-band notched AVA antenna.
6.4 Conclusions

An AVA with dual-notch bands was presented. Adjoining two pairs of mushroom-like EBG structures to the antenna feed-line lead to two frequency stopbands. For verification purposes, an AVA with notches at 3.8 and 5.7 GHz, which correspond to WiMAX and ISM bands, respectively, was designed, simulated, and measured. The good agreement between simulated and measured results proves the underlined concept. The proposed approach is advantageous for antennas with non-uniform flares, and flexible in terms controlling both the number and locations of the stopband frequency notches. This straightforwardly principle is simple and efficient. It also eliminates the need to deform the antenna radiators with slots/parasitic elements or comprise multilayer substrates. Ease of fabrication and excellent electrical performance provide a competitive design that fits many wireless applications questing more EM immunity.
Chapter 7

Conclusions and Future Work

7.1 Summary

The main goal of this dissertation was to demonstrate systematic approaches for the design of front-end microwave components with an improved frequency response and bandwidth accessibility. Furthermore, the realization of the resulting schematics was considered as a point of concern by avoiding any increase in the structural complexity, circuitry occupation, as well as manufacturing cost.

Chapter 2 presented the mathematical derivations of microstrip NTLs, which were then utilized in various proposed components throughout this dissertation. The concept of such impedance-varying profiles was analytically justified by a proof-of-concept example of a miniaturized NTL transformer matching predefined source/load impedances at a predefined frequency. The results were examined in two different ways; specifically, optimizations and modeling, to demonstrate the merit of impedance-varying lines as a competitive candidate not only in achieving a certain electrical performance, but also in miniaturizing the overall circuitry area. Trust-region-reflective algorithm as well as ANN models were adopted as optimization and modeling tools, respectively.
Chapter 3 illustrated the applications of NTLs in the design of a miniaturized planar (single-layered) UWB in-phase equal-split WPD. The realization of such a divider was performed by replacing the uniform microstrip transmission lines in each arm of the conventional design with impedance-varying profiles. Variations were governed by a truncated Fourier series with coefficients optimized to achieve an UWB frequency matching (i.e., 3.1–10.6 GHz). The design concept was built on a clear mathematical platform inspired by transmission line theory. The even-mode analysis was carried out to optimize the series coefficients according to the intended performance; whereas the odd-mode analysis was utilized to obtain the optimum isolation resistors that guarantee an acceptable isolation and output ports matching. The proposed design procedure resulted in a compact easy-to-fabricate structure. For verification purposes, an optimized equal-split UWB power divider was designed, simulated, and measured. The good agreement between both simulated and measured results over the 3.1–10.6 GHz frequency range proved the validity of the design procedure. The optimization-driven framework was also modeled utilizing a QN-based trained ANN to tackle the burden of optimization time and complexity. The results of the two optimization routines (series coefficients and isolation resistors) were considered in a single-staged model. The achieved accuracy and the superfast modeling of impedance variations were two major advantages of the illustrated model. S-parameters derived from the trained ANN outputs were in a good agreement with those obtained by the time-consuming optimization, and showed an excellent electrical performance across the UWB frequency range. Although modeling examples were based on training data derived from analytical optimizations, the overall design was accurate as justified by EM simulations.
Chapter 4 presented the concept of NTLs optimized to wideband multi-way BPD applications. The soul of the proposed procedure depended on substituting the single-frequency matching quarter-wave sections in the conventional design by impedance-varying transmission lines of flexible bandwidth allocation and matching levels. Based on the equivalent transmission line model, impedance variations were profiled according to a truncated Fourier series with coefficients determined by an optimization procedure. To validate the proposed concept, three 3-way BPDs of different fractional bandwidths were designed. Then, two examples of 3- and 5-way BPDs with bandwidths of 4–10 GHz and 5–9 GHz, respectively, were simulated, fabricated, and measured. Simulations and measurements showed an excellent agreement, with input port matching of below –15 dB and –12.5 dB for the 3- and 5-way dividers, respectively, over the bands of interest. Furthermore, the obtained transmission parameters of the 3- and 5-way dividers were –4.77 ± 1 dB and –7 ± 1 dB, respectively, over the design bands.

Chapter 5 discussed the applications of NTLs in the design of multi-frequency and broadband quadrature hybrid BLCs. In the multi-frequency design, each uniform transmission line branch was replaced with single NTL of the same length, but exhibiting a Fourier-based profile. First, properly formulated design equations were derived from the even-/odd-mode analysis according to a systematic guideline. Then, the resulting equations were solved by adopting an optimization-driven process in order to achieve the desired response at the predefined frequencies. The design principle was justified by simulating, fabricating, and measuring two examples of dual- and triple-band 90° BLCs suitable for GSM, WLAN, Wi-Fi, and WiMAX. The agreement between both simulated and measured data validated the design methodology.
This chapter also proposed a novel design procedure to broaden the bandwidth of a single-section 90° BLC with ports extensions. Uniform impedances of the conventional coupler design were replaced with NTLs through an optimization-driven process based on the even-/odd-mode circuits. Consequently, the optimum Fourier series coefficients that meet given design requirements (i.e., broadband frequency characteristics) were obtained. The proposed methodology was capable of achieving different fractional bandwidths for specific coupling levels by the proper modulation of the incorporated impedance profiles. It showed advantages in applications where BLCs with broadband frequency characteristics and low coupling levels were imposed. The adopted technique was analytically justified by exploiting three examples of 3-dB, 6-dB, and 9-dB BLCs with fractional bandwidth of 57%. Further validations through simulated and measured results were provided. The proposed BLC designs were systematic and valid for any coupling level. The underlying principles resulted in compact and planar (i.e., single-layered) structures with effective higher-order harmonics suppression as for enforcing the multi-/broadband functionality only at predefined frequencies/fractional bandwidth.

Finally, Chapter 6 proposed an AVA with dual-rejection bands by incorporating mushroom-like EBG cells. It was concluded that surrounding the antenna feed-line with two pairs of EBG structures led to two frequency notches (i.e., a notch per EBG pair). For verification purposes, an AVA with notches at 3.6–3.9 and 5.6–5.8 GHz was designed, simulated, and measured. The good agreement between both simulated and measured results proved the concept of utilizing EBG elements, with VSWR greater than 5 at the notches locations, and less than 2 elsewhere. Such frequency notches are of importance in various technologies, especially those operating in the WiMAX and ISM bands. It was
seen that this simple yet efficient approach is advantageous for antennas with non-uniform flares, and flexible in terms controlling both the number and locations of the frequency notches. The proposed antenna design resulted in relaxing the need to disfigure or deform the two antenna radiators with slots/parasitic elements or comprise multilayer substrates. The ease of fabrication and excellent electrical performance, characterized by high rejection levels, provide a competitive design that fits many wireless applications.

7.2 Impact on Different Disciplines

Research outcomes demonstrated in this dissertation have a significant merit in adding values to the existing scientific, educational, and industrial fields. The presented studies complement other interdisciplinary areas of electrical engineering, and equally contribute in the development of futuristic technologies.

7.2.1 Global EARS Community

The embedded research impacts nowadays applications (e.g., computer networks, radars) as the main theme of this effort addresses enhancing access to the radio spectrum (EARS). Underlined investigations directly tackle congestion of the scarce frequency spectrum by proposing front-ends that support emerging mechanisms (e.g., cognitive radios) aiming to exploit the underutilized bandwidth. Compatible front-ends presented in this work enable multi-/wideband functionalities for spectrum scanning, determination of inactive frequency band(s), and transmitting/receiving at unexploited channels. The presented research creates a platform for joint collaborations among different areas in electronics/communications engineering to explore solutions to the impending spectral insufficiency problem. Moreover, the conceptual focus on this avenue furnishes guiding
principles to undergraduate and graduate students seeking more knowledge in EARS philosophy. The state-of-the-art tools applied in this EARS-oriented research, such as computer aided design, modeling, simulation, and testing paradigms provide a foundation for future utilizations of such tools in this ever-growing concept, characterized by the emphasis on cognitive communication schemes and mechanisms.

7.2.2 Academia, Society, and Industry

The studies demonstrated in this work benefit both students and scholars by presenting a comprehensive analysis of widely taught front-ends. The underlined components are communicated in almost any RF/microwave engineering reference. Besides, the proposed schematics illustrate in-depth investigations by manipulating well-known design approaches (e.g., transmission line theory). Clear mathematical guidelines are shown in this effort and are systematically driven from theory to practice through engineering reasoning, professional simulations, and experimentation channels. This research helps paving the way for new researchers in this field toward applying science, technology, engineering, and math (STEM) in their own research activities.

There has been an exponentially growing quest on higher data rates, leading to the congestion of the frequency spectrum. As a result, maintaining a reasonable quality of service to public users is endangered. This research introduced front-ends of various designs engineered to be compatible with the emerging solutions to spectrum congestion (e.g., cognitive radios). Hence, the proposed schematics have the potential to enable more efficient bandwidth use; which in turn benefit publicity by achieving higher transfer rates, welcoming more simultaneous devices to log in, and implementing more convenient wireless communication channels/protocols.
The developed methodologies are also useful to the industrial market in the sense that the proposed designs possess advanced electrical characteristics that are of grave importance to current and future applications. Such designs come at no expense to the circuitry occupation, design complexity, and cost. In other words, all novel topologies herein are compact and planar (i.e., single-layered). Moreover, the proposed schematics utilize microstrip line technology to realize inexpensive custom designed front-ends with flexible redefinition capacities and minimum added fabrication constraints (e.g., extra transmission lines, multi-layer structures, packaging and manufacturing).

7.3 Future Work

Research concepts introduced in this dissertation can be further extended. The exploited methodologies, which led to proof-of-principle designs, have the potential of being redefined to contribute toward futuristic real-word applications.

In Chapter 2, miniaturized impedance-varying transmission lines were proposed as an equivalent to the counterpart uniform lines. Examples to extend this work include:

- Modulating the variations of high impedance lines – Although this study showed a merit in future replacement of conventional PCB traces with compact NTLs, controlling the profile of high impedance microstrip lines remains as a major challenge. There have been remarkable efforts on finding solutions to address the impractical narrow widths of high-impedance lines (e.g., short-ended coupled lines [99]). Non-uniform profiles have not yet been utilized in this investigation. Such a technique may have the potential in tackling this challenge and then be used in the design of microwave front-ends with advanced functionalities that could never be realized with conventional microstrip technology (e.g., dividers with high split ratios, broadband multi-stage couplers).
In Chapter 3, a design procedure of a two-way WPD with UWB frequency characteristic was illustrated utilizing impedance-varying profiles with a wideband matching nature. This concept is valid for other divider topologies. Examples include:

*Wideband multi-way dividers* – Impedances with non-uniform profiles can be utilized in the design of WPDs with wideband matching and multiple output ports. Resistors with optimized values between each two adjacent arms are needed to maintain acceptable isolation and output ports matching conditions. Wideband multiple-output dividers are essential to feed sophisticated antenna arrays, especially in radar applications [100].

*Wideband unequal-split dividers* – Based on how impedances of the arms are profiled, compact wideband WPDs with unequal-split ratios can be designed. Though, a different odd mode analysis than that presented in this dissertation to be carried out to optimize the values of the isolation resistors. Furthermore, extra NTLs with wideband characteristics are required to match the resulting asymmetric output ports to $50 \, \Omega$ (i.e., impedance of the SMA connectors) [101].

*Wideband multi-way unequal-split dividers* – Research described in I and II can be applied to design dividers with integrated functionalities. Such custom designs are beneficial to the microwave community and industry, taking into account planarity, compactness, and compromised complexity/cost as advantages.

In Chapter 4, wideband multi-way BPDs were demonstrated. NTLs were adopted in the design of such dividers with predefined bandwidth and matching levels. Research on BPDs can be extended in many ways. Examples include:

*Output ports’ isolation and matching* – Although BPD design does not incorporate lumped elements (e.g., resistors) and has ports that can conveniently be extended to any number, the output ports’ are neither matched nor isolated at design frequency(s). Thus,
wideband matching techniques (e.g., series/parallel stub networks [75]) with applications to enhance BPD electrical performance worth investigations. *Consideration of even number of outputs* – Except the study reported in [27] by the same author, all BPD designs found in literature were presented for dividers with odd number of outputs. However, it could be the case that some applications may require topologies with even ports. Hence, there is a lack of generalized designs of \( N_o \)-way BPDs (where \( N_o \) is either even or odd) with advanced characteristics (e.g., multi-/wideband operation, arbitrary split ratios). *Physical occupation and realization concerns* – As the number of ports in a BPD increases, more challenges arise (e.g., physical circuit area, ports alignment). Hence, maneuvering wideband miniaturization techniques that can be utilized in BPD structures are of importance to manufacturing and packaging processes.

In Chapter 5, designs of multi-/broadband couplers were demonstrated. NTLs were adopted to design BLCs with predefined frequencies, bandwidth, and coupling levels. Research on 90° hybrids can be extended in many ways. Examples include:

3-\( dB \) couplers with UWB frequency matching – Despite presenting 3-dB BLCs with broadband characteristics in this dissertation, broadening the bandwidth of an equal coupling (i.e., 3-dB) BLC to cover the UWB spectrum remains as a challenge. In order to bring this matching feature into practice, compact broadband 3-dB NTL couplers each at a different center frequency can be multi-staged/cascaded. NTLs prove to be a promising solution in related studies (e.g., bandwidth improvement, circuitry miniaturization, harmonics suppression). Multi-band couplers with custom coupling levels – In this dissertation, multi-frequency (i.e., dual-/triple-operation) 3-dB BLCs were proposed. However, some applications may quest more advanced custom designs, such as multi-
functionality with dissimilar coupling levels at each frequency. Transmission lines with impedance varying profiles could have the merit of realizing this complex scheme by modifying the optimization routine (i.e., objective function described in equation (5.13)) to reflect the design frequencies and their corresponding coupling values. **Couplers for crossover applications** – Quadrature hybrid BLCs can be redesigned for crossover application [102]-[103]. S-parameters of an NTL coupler can be modified according to crossover features. More sections of broadband NTL hybrids can also be incorporated in the composite design for broader bandwidth.

In Chapter 6, a design of an AVA with dual narrow band-notch characteristics was presented. EBG pairs were incorporated to realize band-reject frequency response at commercial bands. Examples of future research on AVAs include:

**Tunable multi-frequency notches** – As concluded in Chapter 6, the obtained notch frequencies are fundamentally due to the capacitance and inductance resulting from deploying the EBG cells around the antenna feed-line (refer to Section 6.1). However, the corresponding locations of the notches are fixed once EBG cells are printed on the substrate. In order to obtain tunable notches, biased/active circuitry are suggested. In other words, incorporating variable capacitors (i.e., varactors) along with the EBG structures to vary the resulting capacitance, and thus, the notch location, is an interesting research topic that worth maneuvering. **Multi-notch AVAs with modified EBG structures** – The incorporated design in this study requires one EBG cell (or pair) for each notch. Parametric simulations can be carried out for modified EBG cells with multiple bandgaps [104]. Finally, notches can also be realized utilizing vialess EBGs to relax the drilling process of those of the conventional type [105].
7.4 Research Publications and Outcomes


Conference Papers


Attended Workshops


References


