Passive components design for UWB-RFID systems

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PASSIVE COMPONENTS DESIGN FOR UWB-RFID SYSTEMS

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

2013

STATEMENT OF ORIGINALITY

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

Date

SHEN Yizhu

DEDICATION

TO

My supervisor

And

My grandparents, my parents and my husband

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SUMMARY

Recently, radio-frequency identification (RFID) systems emerge in many daily applications. However, their operation frequencies are narrowband which inherently limit their capabilities. To break this bottleneck, impulse radio based ultra-wideband (UWB) technology is a promising candidate. It enables the next-generation RFID system with powerful functions such as real-time location tracking with high accuracy. This PhD thesis presents some key passive components for UWB enabled RFID (UWB-RFID) systems.

Filter is one of the passive components in a UWB-RFID system. This thesis presents novel UWB filters which employ composite right/left-handed transmission line (CRLH-TL) structures. To facilitate the filter designs, their circuit models are introduced and analyzed. Following a comprehensive theoretical analysis, design guidelines are given. Two different filters are then numerically simulated, experimentally measured, and practically integrated in UWB-RFID transmitter and receiver, respectively. These filters feature compact size, low in-band insertion loss, and also high out-of-band suppression to avoid interference. In addition, an absorptive filter is proposed to absorb the stop-band power rather than reflect it towards the UWB source. It therefore makes the UWB source operate with high stability.

Antenna is another key component. This PhD thesis presents several planar antennas for different UWB-RFID tags. The first antenna is a two-port design for wireless-powered UWB-RFID tag. In the same volume, it physically integrates two antennas, i.e., a narrowband slot for energy harvesting and a circularly polarized (CP) quasi-spiral for UWB signal transmission. This two-port antenna features compact size and high electrical isolation. The second antenna is a CP square slot for active UWB-RFID tag. After simulating and measuring this UWB CP antenna in the frequency and time domains, it is then integrated with active circuits for demonstration. Moreover, this square slot is optimized to be embedded in Concrete for debris tracking.

Following the above filters and antennas, this thesis also demonstrates a UWB-RFID system using chipless CP tag for the first time. The chipless tag adopts the above designed square slot. Its CP characteristic significantly benefits the UWB-RFID system by reducing amplitude ratio between structural and antenna modes, and also alleviates the mutual coupling between transmitter and receiver. In addition, the chipless tag uses short delay line which makes the chipless tag more compact, resulting in overlapped backscattered pulses. To avoid the latter side-effect, a novel four-step time-domain processing scheme is developed to extract both structural and antenna modes from the overlapped pulses. The proposed scheme also enables the system to operate at lower frequency and extends its detection range.

Overall, this thesis presents filters and CP antennas for several different UWB-RFID systems. Following these designs, this thesis also demonstrates a UWB-RFID system using CP chipless tag and novel processing scheme. This thesis gives some new and broad insights for next-generation UWB-RFID systems.

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LIST OF ABBREVIATIONS

AR	axial ratio
BPF	bandpass filter
BSF	bandstop filter
CBCPW	conductor-backed co-planar waveguide
CP	circularly polarized
CPW	co-planar waveguide
CRLH-TL	composite right/left-handed transmission
EIRP	equivalent isotropically radiated power
FCC	Federal Communications Commission
FD	frequency-domain
IR-UWB	impulse radio ultra-wideband
IC	integrated circuit
LH	left-hand
MIM	metal-insulator-metal
OFDM	orthogonal frequency-division
PSD	power spectral density
RF	radio frequency
RFID	radio-frequency identification
RH	right-hand

TD	time-domain
TSA	tapered slot antenna
UWB	ultra-wideband
UWB-RFID	ultra-wideband radio-frequency
VNA	vector network analyzer
WLAN	wireless local area network

Chapter 1

Introduction

1.1 Background

In 2002, the Federal Communication Commission (FCC) released the legal commercial usage of the frequency spectrum covering 3.1-10.6 GHz, with an average mask emission limit of -41.3 dBm/MHz [1]. FCC also defined ultra-wideband (UWB) as its fractional bandwidth is more than 20% referred to the center carrier frequency. In the following years, several countries or regions also proposed their own regulations. For example, in 2007, the Info-communications Development Authority (IDA) of Singapore released its UWB regulation, and

limited the emission power as -41.3 dBm/MHz in the ranges of $3.4\text{-}4.8$ GHz and $6\text{-}8.5$ GHz [2].

Following these regulations, two approaches have been widely investigated to efficiently use the frequency spectrum. One is dividing the wide frequency band into several sub-bands. This method is usually categorized as multiband orthogonal frequency-division multiplexing (OFDM). Combining multiple-input multiple-output (MIMO) and proper coding techniques, multiband-OFDM based UWB shows its great potential for high-speed short-range wireless communications. The other approach is using short pulses which occupy the whole UWB frequency range, to transfer information. It is therefore named as impulse radio based ultra-wideband (IR-UWB). Benefiting from the short pulses, IR-UWB features multi-path immunity and high resolution. Therefore, IR-UWB is very useful for real-time localization and tracking. In addition to these two approaches which are mainly for high-speed communication and high-resolution localization, respectively, UWB technology has also been explored to meet modern wireless trends. For example, today's radio-frequency spectrum becomes very crowded. To smartly use this limited resource, the emerging cognitive radio attracts worldwide attention. A cognitive radio efficiently adopts UWB to sense, access and optimize the operation frequency, aiming to avoid data traffic and provide high-quality service.

Paralleling with the rapid development of UWB technology, radio-frequency identification (RFID) technology has also been widely investigated and applied in our daily life in the past years. A typical RFID system consists of a reader and several tags carrying different information, and relies on backscattered signals for identification. Currently, the operation frequencies of most commercial RFID systems are 125 kHz, 134 kHz, 13.56 MHz, 433 MHz, 860-960 MHz, 2.4 GHz and 5.8 GHz. Their narrow bandwidths seriously limit the indoor localization accuracy. As the demand for localization of assets in indoor environment increases, there is a need to introduce a novel technology into the conventional RFID system to boost its capability. The above introduced UWB technology, which is widely employed in positioning and localization [3], is a good candidate to be merged with RFID to form a UWB enabled RFID (UWB-RFID) system.

1.2 Motivation and Objectives

Since the FCC released the commercial use of UWB technology [1], many UWB modules have been designed and implemented for practical applications. Meanwhile, narrowband RFID systems are also widely applied for commercial applications. To further booster the capabilities and enrich the functionalities of RFID systems, our researcher group have being investigate UWB technology for RFID system, i.e., UWB-RFID system, especially for real-time localization with high resolution.

This thesis focuses on passive components for these UWB-RFID systems. Filter is an essential one. Many reported UWB bandpass filters features good in-band performance. However, the out-of-band characteristic, which is also very important, is seldom studied. In our project, the UWB-RFID systems require bandpass filters with high out-of-band suppression to alleviate interference. To facilitate the system demonstration, this thesis presents novel bandpass filters, by modifying the conventional composite right/left-handed (CRLH) transmission line (TL) [4]. Moreover, to maintain the UWB source with high stability, an absorptive filter is required to absorb the out-of-band reflected power.

Antenna is another important block for UWB-RFID systems. However, most published UWB antennas are with linear polarization, whereas the UWB-RFID systems prefer circularly polarized (CP) UWB antennas to track the moving or rotated RFID tags. To meet these requirements from system point of view, this thesis presents several CP UWB antennas for different UWB-RFID systems.

In addition, only chipless tags with linear polarization have been demonstrated for UWB-RFID system recently. In this thesis, circularly polarized chipless tag is designed and employed to enable a UWB-RFID system with additional advantages, such as reducing the signal ratio of structural to antenna mode, and enabling orientation determination. Furthermore, long delay line is adopted in conventional passive tag design, aiming to generate enough time intervals to physically separate the backscattered structural and antenna modes in the time-

domain. To simplify the passive tag design with short delay line and compact size, this thesis presents a novel four-step processing scheme to detect overlapped structural and antenna modes. A UWB-RFID system with center frequency of 4.1 GHz is also built up to validate the designed chipless CP tag and the proposed four-step scheme.

1.3 Major Contributions

Aligned with the motivation mentioned above, the research outcomes achieved in this thesis include UWB filters, CP UWB antennas, and a UWB-RFID system using chipless CP tags. Some components are integrated for different UWB-RFID systems, i.e., wireless-powered, active, embedded in Concrete, and chipless.

The major contributions of this thesis are listed as follows:

- 1) Two novel CRLH-TL unit cells are proposed for bandpass filter design. The thorough theoretical analysis provides a guideline to design bandpass filters with high attenuation at desired out-of-band frequencies. Based on the proposed unit cell, this thesis also presents an absorptive filter, which reshapes the source pulses and also absorbs the reflected power. This unique characteristic enhances the stability of UWB source.
- 2) A quasi-spiral UWB antenna with wideband circular polarization is designed for UWB-RFID system. Compared with conventional spiral

antenna with complicated feeding network, the proposed quasi-spiral antenna has simple structure and compact size. The two arms are offset to reduce the input impedance, and the bottom arm is used as ground of Microstrip feeding line. Moreover, optimized axial ratio is achieved by tuning the feeding point.

- 3) A two-port antenna for wireless-powered UWB-RFID tag is proposed. Conventionally, the wireless-powered tag consists of two separate antennas which occupy a large area. To make the tag compact, a two-port antenna is designed by physically integrating the receiving and transmitting antennas in the same volume. Although they are physically integrated together, the receiving and transmitting parts have high electrical isolation in simulation and measurement.
- 4) Square-slot antennas for active UWB-RFID tag in free space and also embedded in Concrete are presented in the thesis. Conventionally, a single-fed antenna has simple structure but narrow axial-ratio bandwidth, whereas dual-fed antenna achieves wide axial-ratio bandwidth but needs a complicated feeding network. In this thesis, a planar antenna working in lower UWB band is proposed using single-fed CPW. In addition, using the measured dielectric properties of Concrete, this antenna is optimized as an implantable antenna for a UWB-RFID active tag which is embedded in Concrete.

5) A UWB-RFID system using chipless CP tag is proposed and demonstrated for the first time. The conventional linearly polarized system has high signal ratio between structural and antenna modes, and requires chipless tag with long delay line. Using chipless CP tags, the proposed system significantly reduces the signal ratio to unburden other circuits. Moreover, a novel four-step processing scheme is proposed to separate the overlapped structural and antenna modes. This scheme simplifies the passive tag design with short delay line.

1.4 Organization of the Thesis

Chapter 2 gives brief introductions on three UWB-RFID systems, UWB bandpass filters and circularly polarized UWB antennas. Recent research progress on active, chipless, and wireless-powered UWB-RFID systems is investigated and compared. Bandpass filter designs based on three structures: hybrid transmission line, multiple-mode resonator, and CRLH-TL are reviewed in details. Moreover, single- and dual-fed UWB antennas with circular polarization are reviewed. Finally, embedded UWB antenna designs and applications are briefly introduced.

Chapter 3 investigates a CRLH-TL unit cell for bandpass filter design. Compared with the conventional structure, extra shunt inductor and capacitor are added in the series arm in the conventional unit cell, which are used to generate high

attenuation at resonant frequencies. Theoretical analysis on the proposed unit cell is given in details, and the derivation of the shunt lumped component value as well as cutoff frequencies are also shown in this chapter. Then, a three-layer based bandpass filter is designed by employing this unit cell. Good agreement among the theoretical, simulated, and measured results validates the proposed structures. Following that, a filter with both bandpass and bandstop characteristic is designed to absorb the reflected power and improve the emitted pulse shape.

Chapter 4 introduces a quasi-spiral UWB antenna and a two-port antenna for wireless-powered UWB-RFID system. This two-port antenna physically integrates two antennas in the same volume, and the receiving antenna is a linearly polarized narrowband slot for energy harvesting, while the transmitting one is a circularly polarized UWB quasi-spiral for signal radiation. The isolation between these two ports is higher than 20 dB although they are in the same volume.

Chapter 5 introduces a square-slot antenna for active UWB-RFID tag in free space and embedded in Concrete. Both frequency- and time-domain results show that this design can realize wideband axial ratio and impedance matching in the lower UWB band. Far-field radiation of the integrated active UWB-RFID tag is measured. The transfer function of the receiving antenna is characterized and de-embedded from the received pulse, and the radiation is then compared with FCC indoor regulation. Following that, the square-slot antenna is re-designed for

active tag embedded in Concrete. This antenna takes into account of the measured dielectric properties of Concrete. The fabricated antenna is integrated with active circuits to form an active tag. This tag is then covered by Silicon gel, embedded in Concrete, and used for a UWB-RFID system demonstration. Measurement results show the good reading distance between the active tag and four UWB sensors.

Chapter 6 demonstrates a UWB-RFID system with circular polarization, employing the square-slot antenna introduced in Chapter 5 as chipless tag. By using a horizontally polarized receiving antenna, which is cross-polarized with the transmitting antenna, the signal amplitude ratio of structural to antenna mode is reduced. On the other hand, the antenna mode amplitude almost remains the same because the passive tag realizes circular polarization. Meanwhile, a four-step time-domain processing scheme is proposed to extract the structural and antenna modes, even if they are overlapped. This scheme brings many advantages. For example, it enables the chipless UWB-RFID system to work at lower operating frequency, with less propagation loss and stronger antenna mode signal. Moreover, this scheme simplifies the passive tag, because shorter delay line can be employed to integrate with the antenna, instead of using long transmission line to physically separate the structural and antenna modes. In order to verify this CP UWB-RFID system and the proposed four-step processing scheme, a chipless UWB-RFID system is built up. This demonstration system consists of an IR-UWB transmitter, a receiver, and two high-gain UWB antennas

for transmitting and receiving. Measurement results show that the received signal amplitude of the structural mode response is reduced by 10.9 dB, and the overlapped structural and antenna modes can be effectively separated.

Finally, Chapter 7 gives the conclusions and recommendations for future work.

Chapter 2

Literature Review

2.1 UWB-RFID System

Radio-frequency identification (RFID) technology has been widely employed for object identification in many applications, such as supply chain management, retailing, and so on. A typical RFID system consists of a reader and several tags. In some RFID applications, reliable identification and high-accuracy localization are required simultaneously. However, most current RFID systems cannot meet this requirement because their operating frequencies are narrowband. Narrowband signal is sensitive to interference and multi-path, which will reduce

the range resolution in real environment. Therefore, a new technology should be adopted for system improvement, which brings the birth of UWB-RFID.

An IR-UWB system uses very short time-duration pulses as information carriers. IR-UWB technology is expected to bring many advantages to RFID systems. Firstly, UWB has high resolution of multi-path and time-of-flight of the signal. This characteristic results in high-accuracy localization. Moreover, its power consumption is much lower than that of a narrowband system. With these advantages, UWB enabled RFID (UWB-RFID) system is highly preferred for next-generation identification and localization applications [5]. As illustrated in Table 2-1, from the perspective of the tag, UWB-RFID systems can mainly be catalogued into three kinds: active, chipless, and wireless-powered.

Table 2-1 Category of UWB-RFID systems

System Types	Active	Chipless	Wireless-powered
Tag Components	a IC chip, a battery, and a UWB antenna	no IC chip, no battery, only a UWB antenna	a IC chip, no battery, a narrowband antenna, and a UWB antenna
Advantages	long reading range	unlimited lifetime, low cost	long reading range, long lifetime
Disadvantages	limited lifetime, high cost	short reading range	high cost, bulky size
References	[6-10]	[11-15]	[16, 17]

A tag in an active UWB-RFID system radiates UWB pulses generated by clock generator and pulse generator. In the active UWB-RFID tag for indoor localization in [6], the clock signal is used to generate a Gaussian pulse which is then up-converted by a mixer. The UWB signal is then transmitted by a UWB antenna, propagated through the free space, and then captured by many receivers. All these readers are synchronized, thus the obtained time-difference-of-arrival (TDOA) can be utilized in the localization. Similarly, reference [7] introduces an active UWB-RFID system consisting of four receivers, a referenced tag, and an active tag under test. The real measurement in a lab with free-space environment shows high position resolution. A novel UWB-RFID active system is also developed in our research group [8], which employs an pulse generator with pulse forming network in the active tag to generate UWB signals. There is no mixer used like in [6], which brings the benefit of compact structure with improved power efficiency. Moreover, some commercial designs in [9, 10] are also available for active UWB-RFID applications.

Compared with active system, chipless one is with simple structure because there is no power supply and integrated circuits in the tag. It relies on the backscattered UWB pulses with information bits coded in the tags. The backscattered signal of each tag is designed to be unique, so that the tag ID information can be extracted from the backscattered signal for identification and localization. The chipless UWB-RFID tag has many potential applications because of its compact size, low cost, long life-time, and battery-free configuration.

The general scattering properties of antenna show that, there are two parts in the backscattered signals: structural and antenna modes [11]. The structural mode is dependent on the antenna's shape, while the antenna mode is related to the load termination. In the idea case, there is no antenna mode if a perfect matched load is employed for termination. On the other hand, antenna modes will have 180° phase difference when it is terminated with open-/shorted- loads. Reference [12] validates this theoretical analysis by measuring UWB signals backscattered from an antenna with different loads. In addition to the UWB antenna, the tag prototype in [12] requires integrated circuit boards to realize pulse amplitude modulation, which increases the complexity. A novel chipless tag is proposed in our research group [13-15]. The tag has only a UWB antenna with terminated feeding line. The idea is to integrate transmission lines with different length with the UWB antenna for tag identification. By detecting the different delayed time due to the transmission line and different backscattered pulse shapes due to the load termination (matched-load, open-circuit, and short-circuit), a set of tag ID information can be generated. This chipless tag is not only with simple structure and low cost, but also with long life time, because there is no battery required.

Wireless-powered UWB-RFID system adopts both narrowband and UWB signals. The narrowband signal is employed for energy harvesting to power IC chip in the tag, and UWB signal is for communications between tag and readers. In the system proposed in [16], UHF signal is generated at the reader and transmitted to the tag, and the tag is then excited by this wireless transferred power. Following

that, UWB signal is generated by the powered tag, and transmitted to the reader. Reference [17] also proposes a wireless-powered system. The tag can be powered up by wireless power at three different frequencies, namely 915 MHz, 2.4 GHz, and 5.8 GHz. In all, in this wireless-powered system, no battery is needed in the tag for power supply.

Generally, active UWB-RFID system is with high position resolution and long detection range, but its operation time is limited by the battery. Chipless system solves the life time issue, with compact and low-cost tag. Nevertheless, its operation range is short. Wireless-powered system does not have battery and operation range issues, however, the IC chip contained in the tag is with high complexity due to the additional narrowband subsystem.

2.2 UWB Bandpass Filter

Microwave filter has been widely investigated in the past decades. To design a microwave filter, there are mainly two methods, namely, image parameter and insertion loss methods [18], which both provide lumped-element circuits to facilitate design.

The image parameter theory was developed in the late 1930s, mainly for low-frequency filters in radio and telephony. It is based on cascading many orders of two-port networks, and the desired characteristics in both the in- and out-of-bands can then be achieved. However, the disadvantage is that the working

frequency range cannot be freely specified, and many orders of iterations are needed for the design.

In early 1950s, the insertion loss method is therefore proposed to solve the abovementioned problem. Different from the image parameter method, this method has the feasibility to design filter with specified frequency response, including both the in- and out-of-bands. The method is defined by insertion loss of a filter. There are many implementation types including maximally flat, equal ripple, elliptic function, and linear phase. For example, the maximally flat type is employed for low and flat in-band insertion loss filter designs; the equal ripple Chebyshev type is preferred for filter with sharp cutoff at out-of-band; an elliptic functional filter can realize sharp attenuation in the stopband; and the linear phase filter can achieve linear phase response in the passband.

Generally, the insertion loss method provides a way to synthesize and flexibly design a filter with required performance for the in- and out-of-bands. It is therefore widely employed in the filter design. Its design procedure is straightforward. Any kind of filters, i.e., highpass, bandpass, and bandstop design, starts from the lowpass filter with normalized impedance and frequency. Synthesis is given for the lowpass prototype design, and then it is transformed to the desired frequency and impedance. Then Richard's transformation and Kuroda's identities [18] are used to help in the implementation of lumped elements by using transmission lines.

For UWB bandpass filter design, there are three popular methods. The first method is based on cascading shunt and series stubs with main transmission line, and theoretical synthesis is proposed to analyze this structure and predict the performance. Another popular topology for bandpass filter design is multiple-mode resonator (MMR), with simple structure and good performance, which can cover UWB frequency range by generating a few of resonant frequencies. The third one employs the composite right/left-handed transmission line, which is a periodic structure with compact size.

A. Hybrid transmission line

A common structure adopted in bandpass filter design is a highpass prototype [19], which is composed of a transmission line together with many shunted short-circuited arms. The length of the main transmission line and shunt stubs can be carefully chosen according to the cutoff frequency, then the passband will occur periodically in the frequency range. Based on this structure, a 11th order bandpass filter is proposed in [20] to work in UWB band covering 3.1-10.6 GHz.

However, this filter needs via holes for the short-circuited stubs, which increases the fabrication complexity and cost. Therefore, transmission line with series open-stubs is proposed in [21] for bandpass filter, which avoids via holes. Based on this, non-uniform transmission line with both shunt short-circuit and series open-circuit is proposed in [22]. Compared with above two methods, this

fabricated UWB filter achieves high suppression at the out-of-band frequencies, but remains the same passband ripples.

B. Multiple-mode resonator (MMR)

MMR structure is firstly introduced in [23]. It is composed of a MMR line with half-wavelength and two parallel-coupled lines with roughly quarter-wavelength at the two ports. Impedance ratio of MMR structure can be adjusted by its width. Three resonant modes can be generated to cover the desired frequency range at the lower, middle, and upper bands. Another two resonant modes are induced by the added parallel-coupled lines. Therefore, in all, [23] proposes five transmission poles filter. However, the upper stopband of this MMR structure is narrow. This problem has been solved in [24-26] by embedding electromagnetic bandgap elements. Measurement results show that it can achieve high suppression at the stopband, which improves the performance of the conventional MMR structure.

C. Composite right/left-handed transmission line (CRLH-TL)

Recently, CRLH-TL structure has been widely employed in the bandpass filter design, due to its unique characteristics [4]. As plotted in Fig. 2-1, a CRLH-TL unit cell is composed of series and shunt arms. In the series arm, a right-hand (RH) inductor L_R is in series with a left-handed (LH) capacitor C_L . Meanwhile, in the shunt stub, a LH inductor L_L is shunted with a RH capacitor C_R . At low frequencies, this CRLH unit cell can be simplified as a LH highpass filter with C_L

and L_L . On the other hand, at high frequencies, it works as a RH lowpass filter with L_R and C_R . Therefore, this unit cell features bandpass characteristic. Reference [4] demonstrated an implemented bandpass filter. In this design, inductor L_L is realized by a short-circuited stub, C_L is implemented by an inter-digital capacitor, and L_R and C_R are for parasitic reactance, respectively.

Fig. 2-1 Circuit model of a CRLH unit cell.

Although at the first sight, the CRLH-TL filter is very similar to the conventional bandpass filters, there are extinct differences as follows [4]:

- 1) As discussed above, the unit cell has a LH range (L_L and C_L) at low frequencies, featuring highpass characteristic. However, a conventional filter only has the RH range.
- 2) When the CRLH structure is used as transmission line, only its passband is useful. Therefore, in a CRLH structure, the filtering performance is not used due to the parasitic effects at the stopband. Nevertheless, the out-of-band rejection characteristic is one of the key performances in conventional filter design.

- 3) CRLH-TL filter is a cascade of several unit cells, and each unit cell should satisfy the homogeneity condition, i.e., its electrical length is much smaller than quarter-wavelength, to describe the transmission line with numbers of unit cells. On the contrary, this homogeneity condition is not necessary for conventional filter.
- 4) CRLH-TL filter can be implemented using numbers of unit cells with the same LC values; while for the non-uniform conventional filter, each order generally has different LC values.

2.3 Circularly Polarized UWB Antennas

As one of the most significant components of UWB systems, UWB antennas have triggered considerable research work in this field. Nevertheless, most of the reported UWB antennas are linearly polarized. On the other hand, circularly polarized antennas are widely employed in radar and communication systems because they allow flexible orientation of the transmitter and receiver, whereas most of them cannot be used in UWB application due to their narrow impedance bandwidth. Antennas combining both UWB and circular polarization characteristics are seldom studied but highly expected in future radar and communication applications.

Circular polarization is obtained by exciting two orthogonal linearly polarized modes, which are with the same amplitude, but with 90° phase difference. There

are two feeding structures to generate circular polarization: single-fed and dual-fed. These two types will be discussed in details in the following.

Table 2-2 Two feeding types of circularly polarized antennas

Type	10-dB Return Loss Bandwidth	3-dB Axial Ratio Bandwidth	Antenna Size	Reference
Single-fed	1.604-2.45 GHz (39.6%)	1.84-2.08 GHz (12.4%)	70×70 mm ²	[27]
	2.1-3.5 GHz (43%)	2.18-2.58 GHz (17%)	100×100 mm ²	[28]
	1.75-2.624 GHz (39.9%)	1.78-2.15 GHz (18.8%)	70×70 mm ²	[29]
	1.25-1.75 GHz (33.3%)	1.67-2.2 GHz (27%)	185×185 mm ²	[30]
Dual-fed	2.8-5.2 GHz (60%)	2.8-5.2 GHz (60%)	131×88.7 mm ²	[31]
	1.21-2.55 GHz (71.28%)	1.03-2.45 GHz (81.6%)	292×292 mm ²	[32]

A. Single-fed type

The prototype of single-fed type is given in Fig. 2-2. It can excite circular polarization by a so-called perturbation segment, which will disturb the current distributions on the surface of antenna. Thus two orthogonal components with the same amplitude can be generated.

Fig. 2-2 Single-fed type.

These perturbation structures can be as follows: [27] embeds a cross-patch with the same arm length in the square slot center, a widened L-type strip in [28] is connected with CPW fed and perturbed into the slot center, and a strip protrudes from CPW ground to the center of antenna slot in [29]. Besides, some modifications are also done to the conventional dipole [30], to widen the strip arm and optimize the overlapped area of these two widen arms. Compared with conventional dipole only having z -direction current distribution, circular polarization can be induced with widening arm width and asymmetrical feed structure. Thus two orthogonal currents are along in x -direction and y -direction.

However, as summarized in Table 2-2, the above mentioned circularly polarized antennas are with wide axial ratio bandwidth, but they are still not large enough to be used in UWB applications. Therefore, research work is still needed to improve the axial ratio bandwidth of the single-fed antenna.

B. Dual-fed type

Compared with single-fed type, Fig. 2-3 shows the prototype of dual-fed one. With usage of external polarizer, better circularly polarized characteristics can be achieved, as well as wider impedance bandwidth. However, it is at the cost of complicated structure and large size. As demonstrated in Fig. 2-3, the external polarizer normally is the power divider or coupler, and its two arms feed two contiguous sides of a square slot, or orthogonal angle of a circular slot. In this case, circular polarization will be induced by two modes with the same magnitude and 90° phase difference.

Fig. 2-3 Dual-fed type.

Up to now, only a few UWB circular polarization antennas have been proposed based on dual-fed type. The proposed antenna in [31] works in the lower UWB

band. Its external polarizer is a three-stub hybrid coupler, which generates 90° phase difference at the output of the network and feeds into two wide radiator slots. The axial ratio measured result shows that this dual-fed antenna can realize axial ratio bandwidth up to 60%. Another dual-fed antenna proposed in [32] can also realize wide circular polarization bandwidth. A Wilkinson power divider integrated with phase shift is employed as its external polarizer. Same power is obtained at the two outputs of the power divider, and then propagates into phase shifter to obtain 90° phase difference. Moreover, the usage of the power divider improves not only the impedance bandwidth, but also the axial ration bandwidth. The radiation component of this proposed antenna is a circular slot. By employing two above mentioned external polarizers, the four outputs are connected to probes to feed the circular patch orthogonally. Its measured axial ratio is as high as 81.6 %. However, its working range is 1.03-2.45 GHz, which is not suitable for UWB band applications.

However, as summarized in Table 2-2, the size of dual-fed type is generally larger due to the complicated feeding structures compared to single-fed, which will also incur more loss and low the power efficiency of the whole system. Therefore, single-fed UWB circularly polarized antennas with improved axial ratio bandwidth are studied and presented in Chapter 5. Those antennas significantly enable the flexible orientation of the transmitter and receiver in UWB system.

2.4 Embedded UWB Antennas

A book by Yahya and Kim [33] introduces many design cases for the medical embedded antennas. However, the working frequency is narrowband, from 402 to 405 MHz, not suitable for UWB applications. Recently, the IR-UWB technology for embedded antennas has been widely investigated for medical therapy [34-37]. For example, they can be used for diagnosing breast tumor, medical wireless communication, and used as implantable medical devices. There are many kinds of human tissue materials, such as blood, kidney, muscle, stomach, and so on. Some of the dielectric constant and conductivity of those frequency dependent materials in a UWB frequency band are given in Table I of [38]. Dielectric properties of all those human tissues can also be measured according to different applications. Because of the application that the antenna will be embedded inside human body, different antenna designs discussed in [34-37] and [38-41] are all based on the dielectric properties of body tissue and the propagation characteristics in human body. Generally, for the embedded antenna design, material with dielectric properties similar with human tissue is considered in simulation as the surrounding material of the antenna. Models that mimic human head or body are also employed in simulation. After fabrication, measurement setup is built by using a tissue-simulating material such as liquid. The impedance matching and radiation performance are then measured to verify the designed antennas. Moreover, safety measurements are also done to satisfy the regulations of implantable device inside the human body.

Conventionally, X-ray is a common way to detect tumor. Nevertheless, X-ray is non-healthy. In addition, due to the fact that the tumor is very similar with its surrounding human tissue, mistake happens in the practical diagnosis. Alternatively, UWB technology is healthy and can improve the detection accuracy. Moreover, the multiply reflections are also mitigated to make it possible to detect deep tumor. Embedded antennas for tumor detection are generally reviewed as follows. An antenna proposed in [34] is a patch antenna with stacked layers. Compared with conventional stacked-patch antenna radiating in free space using low permittivity substrate, this proposed antenna employs material with high dielectric constant to separate the upper and lower patches. This is due to the fact that, this antenna is used to radiate into a breast tissue, with high dielectric property. Similarly, [35] also designs a slot antenna for detecting breast tumor, which is embedded in liquid. Therefore, the permittivity of the liquid is used in the simulation, and this antenna is designed for 3.1-10.6 GHz UWB applications. Another Microstrip slot antenna is fabricated in [36] operating at 2-6 GHz. It radiates UWB waves into human body to detect breast tumor, so the dielectric property of human tissue is employed in the antenna design. By adopting substrate with high permittivity, the antenna size is comparably small. The compact and planar prototype makes it a good candidate for breast cancer detection applications.

UWB embedded antenna is also widely applied in implantable instruments. For this application, it is necessary to overcome the high attenuation in human tissue

to detect deep objectives or propagate into tissue with high efficiency. Reference [37] is a good example about the cochlear implant design, which introduces UWB wireless communication path through human tissue. It has two parts: an external part called the speech processor to convert sound to electrical signals, and an interior implanted part to receive the transmitted signals. Considering the effect of human tissue, simple monopole and dipole antennas are designed for this implantable device, which is small enough and radiates with high efficiency. In addition to this design, there are some other implanted embedded UWB antenna applications [38-41].

Another research topic is to use UWB embedded antenna for the detection of interior structure of a collapsed building [42]. Let's assume this situation that a building collapses after earthquake. Rescuing survivors is very challenging and time-consuming without knowing the detailed internal collapsed building. Moreover, another collapse may occur if rescuers move some debris. Therefore, in order to improve the rescuing efficiency and protect the life safety of both rescuers and survivors, UWB sensors embedded in building during construction is proposed. In this case, when the building collapses in some undesirable situations, all these sensors will get the information of the collapsed wall such as composition, thickness, dimension and location using UWB radios, and then transmit all data to the rescuing center. By collecting all this information, a three-dimensional view of the interior structure can be obtained.

Generally, embedded UWB antennas have many applications. And an active UWB-RFID tag integrated with UWB antenna and embedded in Concrete for debris tracking in explosion is designed and system demonstrated in this thesis.

Chapter 3

Bandpass Filters for UWB-RFID System

This chapter investigates three UWB filters which are required for UWB transceiver working in lower UWB band (3.4 – 4.8 GHz). They feature compact size, low insertion loss, wide bandwidth, and high suppression of the interference between UWB and wireless local area network (WLAN) systems. The first bandpass filter is designed for UWB transmitter. It has high attenuation at 5.8 GHz to alleviate our UWB-RFID system from interfering with WLAN. The second bandpass filter is designed for UWB receiver. It therefore has high suppression at both 2.4 GHz and 5.8 GHz to reduce external WLAN signals. Furthermore, an

absorptive filter integrating both bandpass and bandstop performances is also designed for UWB transmitter. It absorbs the out-of-band reflected signal. Parts of the research work in this Chapter are published in [43, 44].

As introduced in Section 2.2, a CRLH transmission line can be realized by cascading a number of unit cells, which is shown in Fig. 2-1. The series and shunt arms are actually impedance and admittance branch, respectively:

$$Z = j(\omega L_R - \frac{1}{\omega C_L}) = j \frac{(\omega / \omega_{series})^2 - 1}{\omega C_L}, \quad (3.1)$$

$$Y = j(\omega C_R - \frac{1}{\omega L_L}) = j \frac{(\omega / \omega_{shunt})^2 - 1}{\omega L_L}, \quad (3.2)$$

where $\omega_{series} = 1/\sqrt{L_R C_L}$ and $\omega_{shunt} = 1/\sqrt{L_L C_R}$ represent the resonant frequencies.

Here, same definition is chosen according to [4]: when the series and shunt resonant frequencies are equal to each other ($\omega_{series} = \omega_{shunt} = \omega_0$), it is defined as balanced case in which there is no gap between the LH and RH range. It should be mentioned that balanced case is chosen for all the CRLH-TL filter designs in the Chapter.

3.1 5.8-GHz Suppressed UWB Bandpass Filter for Transmitter

3.1.1 Circuit model and analysis

In Section 3.31 of Chapter 3 in reference book [4], design procedures are introduced with detailed derivations of the LC values in a CRLH unit cell.

Following those procedures and setting two required cutoff frequencies according to our desired passband which is from 3.2 GHz to 5.27 GHz, the values of the CRLH unit cell are derived as $L_R = 7.5$ nH, $L_L = 0.47$ nH, $C_R = 3.18$ pF, and $C_L = 0.2$ pF.

Fig. 3-1 Circuit schematic of the proposed CRLH unit cell with C_S [43].

Compared with the conventional unit cell shown in Fig. 2-1, symmetric T-network is employed in the proposed modified structure as shown in Fig. 3-1. Two capacitors C_S are added in shunt with the series branch to introduce a resonant frequency at 5.8 GHz with high suppression. Different from equation (3.1), the proposed series branch is with impedance:

$$Z' = \left(j\omega C_S + \frac{1}{j\omega \frac{L_R}{2} + \frac{1}{j\omega 2C_L}} \right)^{-1}. \quad (3.3)$$

Theoretically, when $Z' = \infty$, this modified branch can be used to obtain a high

rejection at a specified angular frequency ω_{rej} . The required C_S can be evaluated follows

$$j\omega_{rej}C_S = - \frac{1}{j\omega_{rej} \frac{L_R}{2} + \frac{1}{j\omega_{rej} 2C_L}}. \quad (3.4)$$

Therefore, there is

$$C_S = \frac{2C_L}{\omega_{rej}^2 L_R C_L - 1} = \frac{2C_L}{(\omega_{rej}/\omega_{series})^2 - 1}. \quad (3.5)$$

Letting $\omega_{rej} = 2\pi \times 5.8 \times 10^9$ rad/s (5.8 GHz for WLAN), and still keeping the same values of L_R and C_L , we can obtain $C_S = 0.4$ pF from (3.5).

In the following, the balanced case is discussed. From equation (3.3), the impedance of $Z' = 0$ occurs at $\omega_{series} = 1/\sqrt{L_R C_L}$ which is independent of C_S . Therefore, the modified CRLH unit cell is still a balanced structure and retains flat frequency response characteristics covering the pass band.

Next, the cutoff frequencies of this modified CRLH unit cell are studied and derived as follows. Referring to the illustrations given in Section 3.2.5 of Chapter 3 in reference book [4], the cutoff frequencies can be derived under the condition of $Z \times Y = -4$ (Z and Y are the impedance and admittance of series and shunt arms, respectively). From (3.5), in order to simplify the derivation, we let the value of C_S in the modified unit cell to be n times that of C_L , namely $C_S = nC_L$, where

$$n = \frac{2}{(\omega_{rej}/\omega_{series})^2 - 1}. \quad (3.6)$$

The Z impedance matrix and Y admittance matrix of the proposed modified unit cell in Fig. 3-1 are then derived:

$$Z = 2 \times \left[\left(j\omega \frac{L_R}{2} + \frac{1}{j\omega 2C_L} \right) \parallel \frac{1}{j\omega n C_L} \right] = -2j \frac{(\omega / \omega_{series})^2 - 1}{\omega C_L [n(\omega / \omega_{series})^2 - n - 2]}, \quad (3.7)$$

$$Y = j(\omega C_R - \frac{1}{\omega L_L}) = j \frac{(\omega / \omega_{shunt})^2 - 1}{\omega L_L}. \quad (3.8)$$

Then employing the above-mentioned condition $Z \times Y = -4$, the two new cutoff frequencies of the modified unit cell are derived as follows:

$$ZY = \frac{[(\omega / \omega_{series})^2 - 1][(\omega / \omega_{shunt})^2 - 1]}{\omega^2 C_L L_L [\frac{n}{2} (\frac{\omega}{\omega_{series}})^2 - \frac{n+2}{2}]} = -4, \quad (3.9)$$

$$(\omega_L^2 + 2n\omega_{shunt}^2)\omega^4 - [\omega_{shunt}^2 \omega_L^2 + \omega_{series}^2 \omega_L^2 + (2n+4)\omega_{shunt}^2 \omega_{series}^2] \omega^2 + \omega_{series}^2 \omega_{shunt}^2 \omega_L^2 = 0, \quad (3.10)$$

where $\omega_L = 1/\sqrt{L_L C_L}$. And for the balanced case, there is $\omega_{series} = \omega_{shunt} = \omega_0$. Therefore,

$$\omega_c = \omega_0 \sqrt{\frac{(n+2)\omega_0^2 + \omega_L^2 \pm \omega_0 \sqrt{(n+2)^2 \omega_0^2 + 4\omega_L^2}}{(\omega_L^2 + 2n\omega_0^2)}}. \quad (3.11)$$

It should be mentioned that the equation of cutoff frequency for conventional unit cell derived in [4] is the special case of (3.11), when $n = 0$, i.e., $C_s = 0$, $\omega_{rej} = \infty$.

Using (3.11) and letting desired rejection frequency as 5.8 GHz, cutoff frequencies of the proposed unit cell are calculated as 3.06 GHz and 4.95 GHz.

Fig. 3-2 Simulated frequency response of the CRLH unit cell with and without C_s [43].

Fig. 3-3 Dispersion and attenuation diagram of the proposed modified structure [43].

To verify the above theoretical analysis, circuit schematic simulation is carried out using Agilent ADS. S-parameters of the CRLH unit cell with and without C_S are obtained and shown in Fig. 3-2. The curves clearly demonstrate that, by adding the capacitors C_S , a transmission pole and high rejection are achieved at 5.8 GHz.

The balanced case and calculated cutoff frequencies of the proposed modified unit cell can be verified by propagation constant ($\gamma = \alpha + j\beta$). Fig. 3-3 presents the dispersion and attenuation behavior retrieved from $ABCD$ matrix[4] based on Bloch theory, where p is the unit cell length. Detailed illustrations of the derivations can be found in Chapter 8 of reference book [18]. As demonstrated in Fig. 3-3, LH and RH regions of the modified CRLH structure are 3.06-4.1 GHz and 4.1-4.95 GHz, respectively. There is no frequency gap between these two regions, i.e., the proposed one is a balanced case. This simulated result agrees well with the above theoretical analysis. In addition, the two simulated cutoff frequencies are 3.06 GHz and 4.95 GHz, respectively. They are the same as the theoretical results predicted by (3.11).

3.1.2 Multilayered configuration

The proposed CRLH unit cell is implemented by a three-layered structure for PCB fabrication and illustrated in Fig. 3-4. The layout is symmetrical along line AA'. The series inductor $L_R/2$ is realized by a Microstrip line on the top layer (metal #1 layer). Its length and width is 4.7 mm and 0.1 mm, respectively. The

shunt inductor L_L is represented by a stub in the middle layer (metal #2 layer) shorted to ground. Its length is around 1.2 mm, and width of 0.1 mm.

(a)

(b)

(c)

Fig. 3-4 Illustration of the unit cell of proposed CRLH UWB bandpass filter: (a) top view (it is symmetrical along the AA' line), (b) cross Sectional view, and (c) 3-D explored view (Unit: mm) [43].

It should be noted that, to reduce the fabrication cost, a through-hole via (from metal #1 to metal #3 layers) rather than a blind via (from metal #2 to metal #3 layers) is used to form the short-circuited stub. The capacitors are implemented by MIM structures which are with higher Q factor compared with the inter-digital ones [45]. It therefore reduces the insertion loss of the CRLH bandpass filter. The series capacitor $2C_L$ is realized by the top and middle overlapped layers with a size of $1.2 \times 1.2 \text{ mm}^2$. Similarly, shunt C_R is implemented by the overlapped area of middle and bottom metal layers, with a rectangle of $1.3 \times 4.6 \text{ mm}^2$ and two squares of $1.2 \times 1.2 \text{ mm}^2$. Meanwhile, C_S is presented by part of the feedline, which is a rectangle overlapped between top and middle layers with size of $1.1 \times 2.1 \text{ mm}^2$.

Fig. 3-4 (b) shows the cross Sectional view of the three-layered configuration. Two different substrates, i.e., Rogers RO4003C (dielectric constant is 3.38) and RO4403 (dielectric constant is 3.2), with same thickness of 0.2032 mm are employed in the design. This configuration facilitates the PCB fabrications. Its explored 3D view is depicted in Fig. 3-4 (c) for easy understanding.

Fig. 3-5 Simulated (HFSS and ADS) S parameters of the modified unit cell [43].

Ansoft HFSS is employed to validate the three-layered structure. As illustrated in Fig. 3-5, simulated results using ADS circuit model (in Fig. 3-1) match well with that using HFSS 3D full-wave simulation, except in the frequency band above the rejection frequency of 5.8 GHz. The discrepancy between HFSS and ADS simulation results above 5.8 GHz is mainly due to the middle conductor strip. In the high order band, the implementation of C_S and C_R are no longer lumped. Therefore, in the middle layer, the conducting strip for C_R acts as a transmission

line with tight coupling through C_S at the input and output ports. This results in signals passing through at high order band, which introduce the difference between HFSS and ADS simulation at high frequencies. On the other hand, the difference below 5.8 GHz is because of the inductors' parasitic effect and through-hole via, which are not accounted for in ADS simulation.

3.1.3 Measured results and discussion

The above studied CRLH unit cell has inherent bandpass characteristics. Identical unit cells can be periodically cascaded to form a UWB bandpass filter.

Based on the proposed CRLH unit cell, two CRLH UWB bandpass filters with two and three unit cells are optimized, fabricated, and shown in Fig. 3-6. To facilitate the on-wafer measurement, the Microstrip feed line is transformed to a conductor-backed co-planar waveguide (CBCPW) structure. Its width is 0.3 mm and the gap is 0.15 mm.

Fig. 3-6 Photograph of bandpass filters employing two (top) and three (bottom) CRLH unit cells [43].

Fig. 3-7 S-parameters of UWB bandpass filters with two unit cells in simulation (HFSS) and measurement [43].

Fig. 3-8 S-parameters of UWB bandpass filters with three unit cells in simulation (HFSS) and measurement [43].

Fig. 3-9 Simulated (HFSS) and measured group delay of UWB bandpass filters [43].

Table 3-1 Performance comparison of UWB bandpass filters [43]

Reference	[46]	[47]	This Work	
Filter Type	SIR	CRLH	CRLH	
Frequency (GHz)	3.4-4.8	2.465-5.222	3.09-4.79	3.22-4.61
Insertion Loss (dB)	~2.8	~2	~1.1	~1.2
Size (mm ³)	42×42 ×0.6	33×19.8 ×0.8	16.8 × 8 ×0.6	23.8 × 8 ×0.6
Suppression at 5.8 GHz (dB)	~35	~15	~70	~70

The UWB filter frequency performances are demonstrated in Fig. 3-7 with two unit cells and in Fig. 3-8 with three unit cells, respectively, including HFSS simulated and on-wafer measured results. For the two unit cells in Fig. 3-7, the measured 3-dB bandwidth is from 3.09 GHz to 4.79 GHz, which is with fractional bandwidth of 43.15% referring to the 3.95 GHz center frequency. The in-band insertion loss is approximately 1.1 dB.

Fig. 3-8 shows two-port S parameters of a bandpass filter with three unit cells. It has a measured 3-dB bandwidth between 3.22 and 4.61 GHz (35.51%), with insertion loss of around 1.2 dB. Fig. 3-9 presents the group delay of these two UWB bandpass filters. Both the simulated and measured are flat covering the 3-dB pass band.

Table 3-1 tabulates and summarizes the performances of the proposed filter and other reported ones. From the comparisons, it is found that the proposed CRLH-TL based bandpass filter is not only with small size and good passband performance, but also with sharp rejection skirt at 5.8 GHz. Therefore, it is useful to employ the proposed filter in UWB transmitter design, to alleviate the interference between UWB and WLAN system.

3.2 2.4-GHz and 5.8-GHz Suppressed UWB Bandpass Filter for Receiver

The most widespread WLAN operates at the 2.4 GHz and 5.8 GHz frequencies. These signals introduce strong interference to the UWB receivers. Therefore, it is highly required to integrate a bandpass filter in the receiver side to suppress the strong and narrowband signals from WLAN systems. In this Section, a modified CRLH unit cell with two additional inductors and capacitors is proposed to achieve high out-of-band suppression at two different resonant frequencies. The modified circuit model is theoretically analyzed and numerically simulated first, and then a bandpass filter is designed in the lower UWB band (3-5 GHz) for verification. The fabricated filter is with compact size of $10 \times 9.8 \text{ mm}^2$. Measured results show that it can achieve rejection at 2.4 GHz and 5.8 GHz with 40 dB and 30 dB, respectively, while maintaining wide passband and low insertion loss simultaneously.

3.2.1 Circuit model and analysis

As shown in Fig. 3-10, the symmetrical T-network without C_S and L_S is a typical bandpass circuit. Here, we use the fixed values ($L_R = 7.5 \text{ nH}$, $C_R = 3.18 \text{ pF}$, $L_L = 0.47 \text{ nH}$, and $C_L = 0.2 \text{ pF}$) in Section 3.1.1 for a bandpass filter in low UWB band. However, in the simulated results depicted in Fig. 3-2, there is no high suppression at 2.4 and 5.8 GHz.

Fig. 3-10 Circuit schematic of the proposed UWB BPF with high suppression at two different resonant frequencies [44].

Compared with Fig. 3-1, Fig. 3-10 introduces two additional L_S to achieve another resonant frequency at 2.4 GHz. Therefore, the admittance for series arm is derived as:

$$Y = j \left(\omega C_S - \frac{1}{\omega L_S} + \left(\frac{1}{2\omega C_L} - \frac{\omega L_R}{2} \right)^{-1} \right). \quad (3.12)$$

Theoretically, $Y = 0$ can be used to obtain two transmission poles at desired rejection angular frequency ω_{rej} , where

$$\omega_{rej}^2 C_S - \frac{1}{L_S} = \frac{2C_L \omega_{rej}^2}{\omega_{rej}^2 L_R C_L - 1}. \quad (3.13)$$

To introduce high suppression at 2.4 and 5.8 GHz, i.e., let $\omega_{rej1} = 2\pi \times 2.4 \times 10^9$ rad/s and $\omega_{rej2} = 2\pi \times 5.8 \times 10^9$ rad/s, respectively, and without changing L_R and

C_L values obtained above, L_S and C_S are calculated using equation (3.14) as 3.6 nH and 0.615 pF, respectively.

$$\begin{aligned}\omega_{rej1}^2 C_S - \frac{1}{L_S} &= \frac{2C_L \omega_{rej1}^2}{\omega_{rej1}^2 L_R C_L - 1}, \\ \omega_{rej2}^2 C_S - \frac{1}{L_S} &= \frac{2C_L \omega_{rej2}^2}{\omega_{rej2}^2 L_R C_L - 1}.\end{aligned}\quad (3.14)$$

Fig. 3-11 ADS simulated frequency response of the circuit model [44].

To validate the above theoretical analysis of the proposed circuit model, circuit simulator ADS is employed again for schematic simulation. As shown in Fig. 3-11, by adding capacitors C_S and inductors L_S , two transmission poles with high rejection are achieved at WLAN frequencies (2.4 & 5.8 GHz), simultaneously, whereas the pass band frequency response keeps the same.

3.2.2 Bandpass filter implementation and measured results

To verify the proposed circuit model, the bandpass filter is implemented with symmetrical layout in Fig. 3-12 (a) on a three-layered substrate with size of $10 \times 9.8 \text{ mm}^2$. The three layer substrate is employed as same as that introduced in Section 3.1. The input and output ports are tapered from 50Ω Microstrip line with width 1.1 mm, to a CBCPW structure which is suitable for on-wafer testing.

Compared with structure shown in Fig. 3-6 (a), there are two additional inductors representing by the lines added at the two ports and then connected to the middle layer. Implementation details are illustrated as follows. As shown in Fig. 3-12 (b), all inductors are realized by Microstrip line with width of 0.1 mm. $L_R/2$ is on metal #1 layer with total length of 5.7 mm, L_L is the 1.3 mm short-circuited stub on metal #2 layer, and L_S begins from metal #1 and then goes through metal #2 layer with total length of 6.3 mm. On the other hand, metal-insulator-metal (MIM) structure is chosen to realize the capacitors. $2C_L$ uses the $1.5 \times 1.5 \text{ mm}^2$ top-middle overlapped layer, C_R is introduced by the middle-bottom overlapped layers with a rectangle of $4.4 \times 1.1 \text{ mm}^2$ and two squares of $1.5 \times 1.5 \text{ mm}^2$, and C_S uses partial of feeding line with top-middle overlapped area of $2 \times 0.9 \text{ mm}^2$. Finally, the designed filter is fabricated and shown in Fig. 3-12 (c). It is noted that through-hole via (from metal #1 to ground) is used in fabrication to reduce fabrication cost.

(a)

(b)

(c)

Fig. 3-12 The proposed UWB bandpass filter: (a) top view, (b) 3-D view, and (c) photograph of fabricated filter [44].

To include parasitic effect in the three-layered structure, the 3-D full-wave simulator Ansoft HFSS is adopted in simulation. Fig. 3-13 illustrates the HFSS simulated and on-wafer measured S-parameters. Its 3-dB bandwidth is between 3 and 4.5 GHz, about 40%. In addition, its out-of-band suppressions are around 40 dB at 2.4 GHz and 30 dB at 5.8 GHz. These suppressions effectively reduce interference from WLAN signals. The insertion loss shown in Fig. 3-13 (a) is around 1 dB and the group delay difference over the pass band in Fig. 3-13 (b) is within 0.25 ns, which is within 1 ns deviation.

(a)

(b)

Fig. 3-13 Simulated and measured S-parameters (a) and group delay (b) [44].

3.3 Absorptive Filter for UWB Transmitter

A bandpass filter, which is usually connected between a pulse generator and a transmitting antenna, is required for a UWB transmitter to shape its time-domain pulse. In a conventional bandpass filter design, out-of-band signal is reflected back to the nearby active circuit blocks. For a UWB transmitter, this characteristic, i.e., the out-of-band reflection from a bandpass filter, will result in the pulse generator being unstable, and therefore seriously affect the performance of the UWB transmitter. To alleviate this issue, an absorptive UWB filter, which combines both bandpass and bandstop performance in the same physical volume, is investigated in this subsection.

This absorptive UWB filter has bandpass characteristics in 3-5 GHz and bandstop characteristics in the out-of-band. Based on CRLH-TL theory, the bandpass filtering part has a high suppression at WLAN frequencies using the structure proposed in Section 3.1. On the other hand, the three-order bandstop filter, which can absorb the out-of-band power and pass the desired signals to the following bandpass filter, is added at the input port and terminated with a 50- Ω resistor at the other port. Circuit model, physical layout, simulated and measured results of this proposed filter are described in details as follows.

3.3.1 Filter design and implementation

The circuit model of the proposed filter is shown in Fig. 3-14. For the bandpass filter operating in low UWB band, we use the same values ($L_R = 7.5$ nH, $C_R = 3.18$ pF, $L_L = 0.47$ nH, and $C_L = 0.2$ pF) in Section 3.1.1. This filter introduces a high attenuation at out-of-band 5.8 GHz by adding C_s . Additionally, a third-order bandstop filter is designed following [18] and connected to the input port (port 1) shown in Fig. 3-14 of the bandpass filter. It consists of two series arms of inductor shunted with capacitor, and a shunt arm in the middle of inductor L_2 in series with C_2 to ground. The values of this bandstop part are $L_1 = 1.87$ nH, $C_1 = 1$ pF, $L_2 = 2.078$ nH, $C_2 = 0.9$ pF, $L_3 = 1.29$ nH, and $C_3 = 0.69$ pF.

Fig. 3-14 Circuit model of the proposed absorptive UWB filter.

When employing ADS simulation of the circuit model, the 50- Ω resistor is considered as the third port, and thus S-parameters of this three-port circuit

network are plotted in Fig. 3-15. Compared with Fig. 3-12 of the bandpass filter, it is interesting to note that $|S_{11}|$ changes a lot, especially in the low frequency range. For practical applications, port 1 of this filter is connected with pulse generator. By adding the stopband structure, the reflection coefficient at port 1 is under -10 dB for all frequencies below 2.8 GHz. The results shown in Fig. 3-15 mean that, (1) this absorptive filter can pass signals of 3-5 GHz with low insertion loss and high suppression at 5.8 GHz, and (2) in case of reflecting back to the pulse generator, this absorptive filter can absorb the out-of-band signals by a resistor terminating at the bandstop filter output.

Fig. 3-15 Circuit simulated frequency response of the proposed prototype.

Fig. 3-16 illustrates the physical layout of this filter. The bandpass part between port 1 and port 2 is similar with Fig. 3-4. As shown in Fig. 3-16 (b), the complete

absorptive filter is based on a three-layer substrate with total size of $13 \times 11 \text{ mm}^2$. All inductors are implemented by lines with width of 0.1 mm. To save space, inductor L_1 is realized by line on the metal #1 and #2 layers with same path. The total length is around 4 mm. Similarly, both of inductors L_2 and L_3 are on the metal #2 layer, whose lengths are 7.5 mm and 1.4 mm, respectively. Here, capacitors in the circuit model are implemented by MIM structure. C_1 is realized by the overlapped square area of $1.1 \times 1.1 \text{ mm}^2$ between top and middle layers, and both C_2 and C_3 are with area of $2 \times 2 \text{ mm}^2$ between middle layer and ground. At the bottom of the ground, there is a slot for soldering the $50\text{-}\Omega$ resistor. The 3D view is shown in Fig. 3-16 (c) for easy reading.

(a)

(b)

(c)

Fig. 3-16 The proposed filter with 50- Ω : (a) top, (b) cross-sectional, and (c) 3-D explored view (Unit: mm).

3.3.2 Measured results and discussion

Following the above numerical study, the proposed filter is fabricated using a three-layer PCB and shown in Fig. 3-17. Two ports of this filter are soldered with

SMA connectors, and a 50- Ω resistor is soldered at the ground layer at the backside.

Fig. 3-18 compares the HFSS simulation and measurement results. Out of our expectation, there is serious difference between these two results. The measured impedance bandwidth becomes narrower and shifts to lower frequency.

(a)

(b)

Fig. 3-17 Photographs of the absorptive filter: (a) top view, and (b) bottom view.

To explore and explain the above phenomena, we carried out step-by-step trouble shooting and checked the design in the following aspects:

- **Simulation:**

The same simulator Ansoft HFSS with the same setting is used for the bandpass filters presented in Sections 3.1 and 3.2. The simulated and measured results of these two filters are in good agreement. These results validate the accuracy of the simulator.

- **Measurement:**

The same vector network analyzer is used for the two-port S parameter measurements for the two bandpass filters and this absorptive filter. The only difference is the SMA connector. Different samples are also re-measured. Similar results were obtained.

- **Fabrication:**

Following the confirmation of the above simulation and measurements, the fabrication becomes the most possible root cause for the differences. It should be noted that, the two bandpass filters and the absorptive filter are fabricated by different companies in two-round fabrications. The filters presented in this Chapter are implemented in three-layer PCB. To fabricate a three-layer or multi-layer PCB, two-layer PCBs are etched and then stacked together. Therefore, the thickness of the middle layer substrate (i.e., RO4403), which is critical for the MIM capacitor values and the filter performance, may be not accurately controlled in the stacking process.

(a)

(b)

Fig. 3-18 Frequency response of the filter in simulation (HFSS) and measurement (considering the effect of thickness of middle substrate layer in fabrication)

To investigate the above assumption on PCB fabrication, HFSS simulator is employed to numerically study the S-parameters variation of the absorptive filter

by changing the thickness of middle layer substrate (RO4403) while remaining all other parameters. As shown in Fig. 3-18, when the thickness is chosen as 0.1 mm, the simulated results agrees well with the measured results, except at higher frequency as explained in 3.1.3. It is also noted that, as shown in Fig. 3-18 (b), the notch frequency of around 5.8 GHz does not change much with the thickness of the middle layer RO4403. This interesting phenomena can be explained by the fact that, as shown in Fig. 3-1 and discussed in Sections 3.1, C_S is employed to introduce the 5.8-GHz suppression. On the other hand, as shown in Fig. 3-4, C_S is realized by the top substrate, i.e., the RO4003C substrate, and is independent of the middle substrate (RO4403). Therefore, we can conclude that the thickness of middle layer, which has significant impact on the filter performance, has not been well controlled and implemented in the three-layer PCB fabrication. Although the above-discussed problem exists in this fabrication, the measured $|S_{11}|$ still demonstrates the desired good impedance matching below 2.5 GHz.

Chapter 4

Quasi-Spiral CP Antennas for Wireless-Powered UWB-RFID Tag

4.1 Wireless-Powered UWB-RFID Tags

As discussed in Section 2.1, an active UWB-RFID system can operate over long distance, but with high tag cost due to the need of an IC chip with a battery, and with limited operation time without recharging the battery. On the other hand, a chipless tag has long lifetime but limited reading range. To overcome the bottlenecks of these two kinds of UWB-RFID architectures, more recently, some

research work is done on the wireless-powered (or remotely-powered) UWB-RFID tags [16, 17].

As illustrated in Fig. 4-1, many readers are involved for real-time localization and tracking. Different from the active and chipless UWB-RFID systems, the wireless-powered system consists of two subsystems: a narrowband one and a UWB one. In Fig. 4-1, 5.8 GHz is chosen as an example for the narrowband system. The 5.8 GHz subsystems are used to power on the UWB tags with modulated signals. Therefore, in this kind of system, the tag does not require a battery but can still work as an active tag, i.e., this tag is similar to an active one but battery-free. These wireless-powered UWB-RFID tags have many inherent advantages but also a defect: as illustrated in Fig. 4-2 (a), the tag contains two separate antennas which occupy a large area.

Fig. 4-1 Conceptual illustration of a wireless powered UWB-RFID system [48].

Fig. 4-2 The wireless-powered UWB-RFID tags with different antenna topologies: (a) two separate antennas in [16, 17], and (b) the proposed two-port antenna [48].

In this chapter, as demonstrated in Fig. 4-2 (b), a two-port antenna is proposed to replace the two separate antennas for wireless-powered UWB-RFID tags. The narrowband antenna is used to receive signal for energy harvesting and data

reception, whereas the UWB antenna provides short pulse radiation for UWB communication. It is also useful for cognitive radios [49].

In the remainder of this chapter, firstly, a single-fed UWB antenna with improved axial ratio bandwidth is illustrated in Section 4.2. Based on this structure, a slot is introduced to perform as a linearly polarized narrowband antenna sharing the same volume with the UWB antenna. The details of the proposed two-port antenna are provided in the following Sections. The research work in this chapter is published in [48, 50].

4.2 One-Port Quasi-Spiral Antenna

A spiral antenna is theoretically frequency-independent. Moreover, it is especially suitable for circularly polarized applications, because it can generate perfectly circularly polarized radiation covering a wide frequency range.

Nevertheless, a spiral antenna has several disadvantages [51-53]. Firstly, as a self-complementary antenna, the input impedance of the spiral antenna is $\eta_0 / 2 = 188.5\Omega$ (η_0 is characteristic impedance of the free space). Therefore, it is difficult to match a spiral antenna with standard $50\text{-}\Omega$ systems. Secondly, a wideband balun [54-57] is needed to feed a spiral antenna. A balun transferring Microstrip to coplanar Stripline is usually employed to vertically feed the spiral.

This structure not only increases complexity and size, but also makes it bulky and incompatible with two-dimensional planar circuits.

In this Section, a quasi-spiral antenna is developed for UWB-RFID applications operating at the upper UWB frequency range of 7 to 10 GHz. Its measured return loss and axial-ratio bandwidths are around 72.4% (5.15 to 11 GHz) and 44.3% (6.37 to 10 GHz), respectively. Details of the proposed wideband CP antenna are presented and discussed in the following subSections.

4.2.1 Antenna geometry

Fig. 4-3 Geometry of the quasi-spiral antenna design [50].

A compact planar quasi-spiral antenna is shown in Fig. 4-3. Rogers RO4003C is chosen as antenna substrate. Compared with conventional Archimedean spiral antenna using wideband Balun, this proposed antenna employs a simple tapered Microstrip line to directly feed the two offset arms on the top and bottom of substrate. The bottom arm is used as a partial ground of the feeding line.

Different from a conventional Archimedean spiral antenna, the left arm is shifted down by G_y and then moved to the right-hand side by G_x . These two symmetrical quasi-spiral arms are separately located on two different sides of the substrate. A third-order Chebyshev impedance transformer [18], which is tapered from $50\ \Omega$ to $100\ \Omega$, is used to feed one arm.

As shown in Fig. 4-3, after optimization, the final parameters of the proposed antenna are $W_1 = 1\ \text{mm}$, $L_1 = L_2 = L_3 = 6.67\ \text{mm}$, $W_2 = 0.65\ \text{mm}$, $L_4 = 5\ \text{mm}$, $W_3 = 0.4\ \text{mm}$, $W = 37\ \text{mm}$, $W_4 = 7\ \text{mm}$, $W_5 = 12\ \text{mm}$, $G_x = 7\ \text{mm}$, $G_y = 0.8\ \text{mm}$, and $L = 36.7\ \text{mm}$. The total antenna size is $37 \times 36 \times 0.508\ \text{mm}^3$.

The simulated real part of the input impedance is plotted in Fig. 4-4, with various offset distance (G_x) between two arms when L_4 is fixed as $5\ \text{mm}$. It shows that, by increasing the offset distance from $3\ \text{mm}$ to $9\ \text{mm}$, the real part of Z_{11} can be tuned down to around $50\ \Omega$. After optimization and trade-off consideration, G_x is chosen as $7\ \text{mm}$. Antenna input impedance is then transformed from $100\ \Omega$ to $50\ \Omega$ by a tapered Microstrip line as shown in Fig. 4-3.

Fig. 4-4 Simulated real part of Z_{11} with different G_x [50].

Fig. 4-5 Simulated axial ratio with different L_4 [50].

This proposed quasi-spiral antenna is not a self-complementary structure anymore. The feeding structure is asymmetric and unbalanced. Its circular polarization performances may become frequency-dependent. The axial-ratio bandwidth is therefore numerically investigated and illustrated in Fig. 4-5. For the simulation, the offset distance of G_x is 7 mm, and all the other parameters are fixed, only L_4 changes. The simulated results show that, the distance L_4 significantly affects circular polarization. The axial-ratio bandwidth is widened as L_4 increases. In our design, L_4 is chosen as 5 mm, which is the largest value for the bottom arm of the spiral antenna to be used as the ground for the Microstrip feeding line.

4.2.2 Measurement results and discussion

The antenna is designed using simulator Ansoft HFSS and measured by a vector network analyzer (Agilent N5244A). Fig. 4-6 shows the fabricated prototype. Fig. 4-7 shows the simulated and measured return losses of the designed antenna. The measured 10-dB return loss bandwidth is 5.15-11 GHz (72.4% referenced to 8.075 GHz), while the simulated one is 5-10.6 GHz. Good input impedance matching of this proposed antenna is achieved covering the upper UWB frequency range. The difference between these two results may be due to the fabrication tolerance and the SMA connector for coaxial to Microstrip transition.

Fig. 4-6 The fabricated quasi-spiral antenna: top view (left) and bottom view (right) [50].

Fig. 4-7 Simulated and measured $|S_{11}|$ curves of the developed quasi-spiral antenna [50].

Fig. 4-8 demonstrates that this quasi-spiral antenna achieves a measured 3-dB axial-ratio covering the frequency range of 6.37-10 GHz, or about 44.3 %, which is highly improved compared with previous published single-fed circularly polarized antenna designs [27-29, 58, 59].

Fig. 4-8 Simulated and measured bore-sight axial ratio of the quasi-spiral antenna [50].

The radiation patterns of the proposed quasi-spiral antenna in the x-z plane are also shown in Fig. 4-9 at four frequencies of interest within the bandwidth: 6.5 GHz, 7 GHz, 7.5 GHz, and 8 GHz. The measured antenna gain is about 3 dBi. The measured results match well with the simulation. Moreover, in all angles of the x-z plane, the difference is within 5 dB, which demonstrate the proposed antenna can realize almost omni-directional radiation.

(a)

(b)

(c)

(d)

Fig. 4-9 Simulated and measured radiation pattern in x - z plane: (a) 6.5 GHz, (b) 7 GHz, (c) 7.5 GHz, and (d) 8 GHz [50].

In all, the proposed quasi-spiral antenna can realize circular polarization with good impedance matching. It modifies the conventional Archimedean spiral antenna by locating two arms on both sides of substrate with reasonable offset distance between each other. By carefully choosing the offset feeding point, a Microstrip tapered line can be employed to directly feed the antenna, which is simpler compared with conventional Balun structure.

4.3 Two-Port Quasi-Spiral Antenna

4.3.1 Antenna geometry

Based on the one-port spiral antenna presented in the above Section, a two-port antenna is proposed and shown in Fig. 4-10. It includes a narrowband slot for energy harvesting and a circularly polarized quasi-spiral for UWB signal radiation. These two antennas are physically integrated on a Rogers (RO4003C, $\epsilon_r = 3.38$) substrate in the same volume of $99 \times 92 \times 0.508 \text{ mm}^3$. The optimized antenna parameters are listed as follows: $w1 = 13 \text{ mm}$, $L1 = 48 \text{ mm}$, $w2 = 1 \text{ mm}$, $L2 = 30.45 \text{ mm}$, $w3 = 1 \text{ mm}$, $L3 = 21.05 \text{ mm}$, $w4 = 30.5 \text{ mm}$, $w5 = 32 \text{ mm}$, $s = 0.4 \text{ mm}$, $g = 0.8 \text{ mm}$, $Ls = 20 \text{ mm}$, $g1 = 10 \text{ mm}$, and $shift = 18 \text{ mm}$. Detailed design procedures of this antenna are described in the following parts.

The designed antenna prototype is shown in Fig. 4-10 (a). Similar with the antenna studied in Section 4.2, by shifting the bottom arm along the x and $-y$ axis by 21.6 mm and 0.8 mm , respectively, the quasi-spiral antenna can be easily

matched by a planar tapered Microstrip line. Another important part of the two-port antenna is a narrowband antenna for energy harvesting. As shown in Fig. 4-10, a $20 \times 0.4 \text{ mm}^2$ slot is etched on the top arm of the quasi-spiral antenna. At port 2, a $50\text{-}\Omega$ Microstrip line with size of $30.45 \times 1 \text{ mm}^2$ is then employed to feed the narrow slot resonant at 5.8 GHz . The top arm is also used as the ground of the Microstrip feeding line. This narrow band slot antenna is designed to operate with linear polarization.

(a)

(b)

Fig. 4-10 Geometry and prototype of the proposed two-port antenna [48].

(a)

(b)

Fig. 4-11 Simulated 3-D radiation patterns: (a) UWB quasi-spiral antenna, and (b) 5.8 GHz narrowband slot antenna [48].

Fig. 4-10 shows the two-port antenna consisted of the above designed quasi-spiral and narrowband slot antenna. When two ports of this proposed structure are terminated with $50\text{-}\Omega$ loads, respectively, Fig. 4-11 illustrates 3-D radiation patterns of the two individual antennas. It shows that, the main radiation lobes of the quasi-spiral and the narrowband slot are along the $-x$ axis and the $\pm z$ axis, respectively. This radiation characteristic is useful to reduce the mutual interference of the two physically integrated antennas.

4.3.2 Parametric studies

In this Section, some important parameters of this proposed two-port antenna are investigated to see their influence on the antenna performance including the impedance matching and axial ratio bandwidth. Its parameter studies are carried out by using Ansoft HFSS software. All other parameters that have not been mentioned are kept the same as stated above.

A. The effect of the feeding position *shift*

A planar tapered Microstrip line is employed to feed the top arm, while using the bottom arm as the ground. It is found that the feeding position has significant influence on the impedance and axial-ratio bandwidths. As shown in Fig. 4-12, when the feeding position is shifted to the right-hand side, both the impedance and axial-ratio bandwidths become wider. However, it should be mentioned that *shift* = 18 mm is the largest shifting distance, because the bottom arm is used as ground of the feeding line. The shifting effect is summarized in Table 4-1.

(a)

(b)

Fig. 4-12 The simulated characteristics with varying the arm gap difference *shift*:
(a) $|S_{11}|$, and (b) axial ratio [48].

Table 4-1 Summary of varying *shift* on the antenna performance [48].

<i>Shift</i>	10 mm	14 mm	18 mm
10-dB Return Loss Bandwidth	2-4.1 GHz (68.9%)	2-4.2 GHz (70.9%)	1.95-5.15 GHz (90.14%)
3-dB Axial Ratio Bandwidth	3.4-4.2 GHz (21.1%)	3.32-4.38 GHz (27.5%)	3.35-4.66 GHz (32.7%)

B. The effect of gap difference between two arms g

Compared with conventional antenna, the bottom arm of the proposed antenna is shifted along the $-y$ axis with distance g . When g increases, it has large effect on the input impedance as shown in Fig. 4-13 (a). Meanwhile, the axial ratio does not change much (see Fig. 4-13 (b)) except for a slight shift to a higher frequency. The details are shown in Table 4-2. Finally, $g = 0.8$ mm is chosen for the wide bandwidth.

(a)

Fig. 4-13 Simulated characteristics with varying the arm gap difference g : (a) $|S_{11}|$, and (b) axial ratio [48].

Table 4-2 Summary of varying g on the antenna performance [48].

g	0.4 mm	0.6 mm	0.8 mm
10-dB Return Loss Bandwidth	4.41-5.2 GHz (16.4%)	2.8-5.2 GHz (60%)	1.95-5.15 GHz (90.14%)
3-dB Axial Ratio Bandwidth	3.28-4.61 GHz (33.7%)	3.31-4.67 GHz (34.1%)	3.35-4.66 GHz (32.7%)

C. The effect of the slot width s

A $20 \text{ mm} \times 0.4 \text{ mm}^2$ Microstrip-fed slot is etched on the top radiation arm and used as a narrowband antenna resonant at 5.8 GHz for harvesting energy in the UWB-RFID system. The slot width effect is also investigated in Fig. 4-14 and

summarized in Table 4-3. As s increases, the impedance bandwidth becomes wider, and the resonant frequency moves to higher frequency. $s = 0.4$ mm is employed for the desired frequency.

Fig. 4-14 The simulated two-port antenna with varying the slot size s [48].

Table 4-3 Summary of varying s on the antenna performance [48].

s	0.2 mm	0.4 mm	0.8 mm
10-dB Return Loss Bandwidth	5.4-5.7 GHz (5.4%)	5.55-5.9 GHz (6.1%)	5.71-6.1 GHz (6.8%)
Center Frequency	5.55 GHz	5.725 GHz	5.9 GHz

D. The effect of the slot length L_s

It is also important to investigate the length of the slot. As illustrated in Fig. 4-15 and Table 4-4, the center frequency shifts to lower frequency when L_s increases, while the impedance bandwidth does not change much. Therefore, when $L_s = 20$ mm, 5.8 GHz resonant frequency can be obtained for energy harvesting.

Fig. 4-15 Simulated two-port antenna with varying the slot size L_s [48].

Table 4-4 Summary of varying L_s on the antenna performance [48].

L_s	18 mm	20 mm	22 mm
10-dB Return Loss Bandwidth	6.3-6.6 GHz (4.65%)	5.65-6 GHz (6.1%)	4.95-5.35 GHz (7.8%)
Center Frequency	6.45 GHz	5.825 GHz	5.15 GHz

E. The effect of the slot position g_1

Fig. 4-16 indicates that, changing the feeding position of the Microstrip line with respect to the slot results in variation of the input impedance. The center frequency shifts to higher frequency and the impedance bandwidth is widened by increasing g_1 . Comparison of the effect is given in Table 4-5 “NA” in the table stands for “not applicable” and there is no 10-dB return-loss bandwidth, since all $|S_{22}|$ values are higher than -10 dB when $g_1 = 9$ mm.

Fig. 4-16 Simulated $|S_{22}|$ with varying the slot size g_1 [48].

Table 4-5 Summary of varying g_1 on the antenna performance [48].

g_1	9 mm	10 mm	11 mm
10-dB Return Loss Bandwidth	NA (NA)	5.65-6 GHz (6.1%)	5.8-6.25 GHz (7.5%)
Center Frequency	NA	5.825 GHz	6.025 GHz

4.3.3 Measured results and discussions

After the optimization process discussed in 4.3.2, a prototype as illustrated in Fig. 4-10 (b) is in-house fabricated and measured.

Fig. 4-17 Comparisons of simulated and measured S parameters at two ports [48].

S-parameters of the two-port antenna are measured using a VNA (Agilent N5244A). Fig. 4-17 shows that, the measured return loss at port 1 showing higher than 10 dB between 2.85 GHz and 5.16 GHz, which is around 57.7% of the center frequency of 4 GHz. The wideband operation ensures that this antenna can transmit the lower band UWB signal for the wireless-powered UWB-RFID system.

Meanwhile, at port 2, the measured results verify that the narrowband slot antenna resonates at 5.8 GHz. It can be used to receive a modulated 5.8 GHz signal for energy harvesting and clock generation.

Fig. 4-18 Simulated and measured isolation $|S_{21}|$ between two ports [48].

Fig. 4-18 plots the isolation coefficient between the two ports. Simulated and measured $|S_{21}|$ results are smaller than -20 dB covering the UWB band of interest. The slight difference between simulation and measurement may be mainly due to the fabrication variation. Fig. 4-19 shows that the measured 3-dB axial ratio covers the frequency range of 3.05-4.43 GHz with less than 3 dB variation in bore-sight, corresponding to a fractional bandwidth of 36.9%. This result indicates that the proposed antenna can generate circularly polarized radiation at desired frequency range.

Fig. 4-19 Simulated and measured axial ratio [48].

Fig. 4-20 xz -plane antenna radiation pattern when port 2 is terminated with a $50\text{-}\Omega$ load: (a) 3.5 GHz, (b) 4 GHz, and (c) 4.5 GHz [48].

Fig. 4-21 xz -plane antenna radiation pattern at 5.8 GHz when port 1 is terminated with a $50\text{-}\Omega$ load [48].

Moreover, radiation patterns in xz -plane of the UWB quasi-spiral antenna at port 1 and the 5.8 GHz narrowband slot antenna at port 2 are illustrated in Fig. 4-20. Three frequencies, namely 3.5, 4, and 4.5 GHz, are chosen according to the impedance bandwidth of the UWB quasi-spiral antenna. At 4 GHz, the results show that the quasi-spiral antenna radiates mainly towards the $-x$ axis, with 3 dB beamwidth of around 140° . The peak antenna gain is around 3.2 dBi. Similarly, Fig. 4-21 also plots the simulated and measured radiation pattern at 5.8 GHz. Its maximum radiation is in the boresight of xz -plane, or both the positive and negative directions of the z -axis.

Table 4-6 compares the proposed two-port antenna in the circularly polarized performance with other reported one-port antennas. Based on the single-fed

method, the proposed antenna achieves the widest impedance and axial ratio bandwidths.

Table 4-6 Performance comparison of circularly polarized antennas [48].

Reference	Antenna Size	10-dB Return Loss Bandwidth	3-dB Axial Ratio Bandwidth
[30]	185×185 mm ²	1.25-1.75 GHz (33.3%)	1.67-2.2 GHz (27%)
[28]	100×100 mm ²	2.1-3.5 GHz (43%)	2.18-2.58 GHz (17%)
[27]	70×70 mm ²	1.604-2.45 GHz (39.6%)	1.84-2.08 GHz (12.4%)
This work	99×92 mm ²	2.85-5.16 GHz (57.7%)	3.05-4.43 GHz (36.9%)

As conclusion, for wireless-powered UWB-RFID systems, the proposed antenna is composed of two different functional antennas, i.e., a narrowband slot antenna for energy harvesting and a circularly polarized quasi-spiral antenna for UWB signal radiation, in the same physical volume and with high electrical isolation.

Chapter 5

Square-Slot CP Antennas for Active UWB-RFID Application

In this Chapter, UWB antennas with circular polarization are designed for detection and tracking of debris throw during explosion. UWB technology is employed in this application due to its special characteristics such as high positioning accuracy, multipath immunity, and low power consumption. In this application, the whole UWB transmitter is embedded in Concrete. Therefore, the propagation of UWB signals is significantly affected by the characteristics of

materials surrounding the transmitter. In this Chapter, a square-slot antenna for free-space application is designed and presented first in the Section 5.1. Following that, in Section 5.2, this square-slot antenna is re-designed to be embedded in Concrete, based on the measured dielectric properties of Concrete slab. The proposed antenna is integrated with active circuit to form a UWB-RFID tag, and it is aimed for debris tracking of Concrete in explosion. The research work in this Chapter is published in [60-62].

5.1 Square-Slot Antenna for UWB-RFID Transmitter

A CPW-fed square-slot antenna for circularly polarized UWB application is designed and investigated both in frequency- and time-domains. Measured results show that it has a 10-dB impedance bandwidth from 2.7 to 5.0 GHz, and a 3-dB axial ratio bandwidth from 3.5 to 4.8 GHz. The proposed antenna can be used in lower UWB system and easily integrated with active circuits. The design details are presented as follows.

5.1.1 Antenna geometry

The proposed circularly polarized UWB antenna is shown in Fig. 5-1. It is printed on a Rogers (RO4003C) substrate with size of $35 \times 35 \times 0.508 \text{ mm}^3$ and dielectric constant of $\epsilon_r = 3.38$. It has a square slot with the side length L . A 50- Ω CPW feeding line is with a signal strip of width W_f and two identical gaps of

width g between the signal strip and the coplanar ground plane. The signal strip is widened to a width of W_f and protruded into the square slot with a length of $L_t + g_2$. The length of the truncated corner is m and the inverted-L strip has a length of L_x and L_y with a width of n .

Fig. 5-1 Prototype and geometry of the proposed UWB circularly polarized antenna [61].

The simulated surface current distribution is shown in Fig. 5-2. By using the truncated corner and a grounded inverted-L strip at the two opposite corners, the two elements work as perturbation structures to distort the surface current and help to realize circular polarization. The proposed antenna has simpler geometry, smaller size, and wider fractional impedance and axial-ratio bandwidths. What is more important, it can be used in lower UWB system with uni-planar CPW feeding structure which is simpler compared with the feeding networks in [31, 32]. In general, this novel design is with several advantages such as easy for

fabrication, low cost, and easy integration with microwave integrated circuits (MIC).

Fig. 5-2 Simulated surface current distribution of the proposed antenna [61].

Some crucial parameters of the signal strip which largely influence the impedance matching of the antenna are investigated. Fig. 5-3 demonstrates the effect of changing of width on performance while the length of signal strip L_t is fixed to 15 mm. When the width of signal strip is increased, better impedance matching can be obtained. As to the length of the signal strip shown in Fig. 5-4 with a fixed width of 9 mm, it is observed that the first resonant frequency shifts to lower frequency. After optimizing, an antenna design is fabricated with $G = 35$ mm, $L = 25$ mm, $W_t = 9$ mm, $L_t = 15$ mm, $W_f = 4$ mm, $L_x = 4$ mm, $L_y = 4$ mm, $g = 0.4$ mm, $g_2 = 0.6$ mm, $m = 13$ mm, and $n = 1$ mm.

Fig. 5-3 Simulated return loss curves with different signal strip widths (W_t) (Unit: mm) [61].

Fig. 5-4 Simulated return loss curves with different signal strip lengths (L_t) (Unit: mm) [61].

5.1.2 Frequency-domain measurement results

As shown in Fig. 5-5, the simulated 10-dB return loss bandwidth of the proposed antenna is from 2.6 to 6.4 GHz, whereas the measured one is from 2.7 to 5.0 GHz, about 59.7% referenced to the center frequency of 3.85 GHz. In addition, Fig. 5-6 presents that the antenna achieves a measured 3 dB axial-ratio bandwidth of 31.3 % from 3.5 to 4.8 GHz in bore-sight, and has a wide angle coverage from -50° to 40° in the horizontal plane while facing the direction of $\phi = 0^\circ$. Those results show that the designed antenna meets the desired requirement for lower UWB application and can generate circular polarization radiation efficiently.

Fig. 5-5 Simulated and measured return loss results of the proposed antenna [61].

Fig. 5-6 Measured axial ratio with respect to frequency in bore-sight, and with respect to $\theta(\phi = 0^\circ)$ at 4 GHz [61].

Fig. 5-7 presents the measured radiation patterns in x - z plane at three different frequencies selected in the impedance bandwidth mentioned above. The measured results are normalized to the absolute antenna gain. It shows that bi-directional radiation is obtained as expected, because there is no reflecting plate against the antenna. That is to say, maximum radiation value can be obtained at two bore-sights of $(\phi = 0^\circ, \theta = 0^\circ)$ and $(\phi = 0^\circ, \theta = 180^\circ)$, respectively. Moreover, it is also demonstrated that the proposed antenna can realize wide angle of omni-direction about 80° in x - z plane.

Fig. 5-7 Normalized measured radiation patterns to absolute gain of antenna in x - z plane [61].

5.1.3 Time-domain measurement results

For impulse-radio UWB (IR-UWB) systems, one of the most important characteristic is time-domain waveform, especially in ranging applications where Time-of-Arrival (TOA) measurement is dependent on pulse shape and phase center of an antenna. Therefore, the proposed antenna is also measured in time domain of the UWB pulse response to demonstrate its circular polarization and radiation in x - z plane. In the measurement setup, the transmitter is the proposed antenna fed by a pulse generator with the operating frequency from 3.5 to 5 GHz, while a linearly polarized tapered slot antenna [63] (2.8 to 12.0 GHz) is employed as the receiver. The distance between the transmitting and receiving antennas is large enough to satisfy the far-field requirement.

Fig. 5-8 Fidelity of the proposed antenna according to different angles of ϕ [61].

Firstly, time-domain circular polarization characteristics are measured. Due to the linear polarization of the receiver, by adjusting the different angles (ϕ) of the receiver, the output waveforms due to different observation angles can be measured. In ranging detection, since the time responses of antenna vary in different directions, the parameter of fidelity is proposed to show the capability of pulse detection [64].

$$Fidelity(\theta, \phi) = \max_{\tau} \left[\frac{\int_{-\infty}^{+\infty} S_{tx}(t) S_{rx}(t + \tau, \theta, \phi) dt}{\sqrt{\int_{-\infty}^{+\infty} S_{tx}^2(t) dt \int_{-\infty}^{+\infty} S_{rx}^2(t, \theta, \phi) dt}} \right]. \quad (5.1)$$

‘Fidelity’ in Eq. (5.1) refers to the correlation coefficient between the transmitted pulse $S_{tx}(t)$ and received pulse $S_{rx}(t)$, with delay time of τ . It quantitatively describes the similarity between the two pulses. The fidelity value will reach unity when the two pulses have identical shapes after normalization, which is the case that the receiving antenna does not distort the incident pulse at all. From system point of view, high fidelity reduces the decision error, and therefore increases ranging accuracy. Results shown in Fig. 5-8 demonstrate the correlation of the received pulse and transmitted one at the output port of pulse generator at $\theta = 0^\circ$. The fidelity of the proposed antenna is mostly larger than 0.8 in the range of ϕ from 0° to 300° , and it is almost flat in the whole angle range, which means that the transmitted pulse shape is not seriously distorted by the proposed antenna.

Fig. 5-9 Measured time-domain waveforms of the proposed antenna [61]. (Note: waveforms of $\phi = 200^\circ$, and $\phi = 110^\circ$ are shifted up and down for easy comparison.)

Among all the output waveforms with the angle ϕ varying from 0° to 360° , Fig. 5-9 shows the maximum and minimum peak-to-peak amplitude, which are 8.5 mV at 200° and 11.9 mV at 110° , respectively. The time shift is less than 0.1 ns, and the minimum peak-to-peak amplitude only decreases to 71.4% (-2.9 dB) compared with the maximum one. These results agree with the measurement results of axial ratio and demonstrate that the proposed antenna can realize circular polarization covering the lower UWB frequency range.

Fig. 5-10 Fidelity of the proposed antenna according to different angles of θ [61].

Following that, by rotating the transmitter, the radiations of the proposed antenna on the x - z plane are also investigated. In Fig. 5-10, the fidelity of the proposed antenna is larger than 0.85 in the whole angle range of θ , which ensures the

similarity between two pulses mentioned earlier. As shown in Fig. 5-11, for the three different observation angles of θ , the peak-to-peak normalized value is 1.05, 1.5, and 2, respectively. The value decreases to 52.5% (-5.6 dB) at $\theta = 90^\circ$, and 75% (-2.5 dB) at $\theta = 45^\circ$, compared with the maximum one at $\theta = 0^\circ$.

Fig. 5-11 Radiation waveforms in the x - z plane of the proposed antenna [61].
(Note: waveforms of $\theta = 0^\circ$, and $\theta = 90^\circ$ are shifted up and down for easy comparison.)

5.1.4 Far-field radiation calculation

The proposed antenna is then integrated in UWB active tag as shown in Fig. 5-12, and its far-field radiation is a key parameter to verify the performance of transmitter. It should meet the FCC frequency regulation, not to exceed the average and peak power of emission limit. The transmitted signal in the far-field is calculated in this Section, in the maximum antenna gain direction.

(a)

(b)

Fig. 5-12 The integrated UWB transmitter: (a) front-side, and (b) back-side.

(a)

(b)

Fig. 5-13 Transmitting-receiving antenna system [65].

The detailed derivation of the transfer function of antenna is introduced in [65]. As shown in Fig. 5-13, it has transmitting and receiving antennas with same polarization. According to [65], the transmitting antenna transfer function $H_{TA}(\omega, \theta, \varphi)$ is defined as the ratio of normalized electric field $E(\omega, \theta, \varphi)$ at the spatial test point to the input signal $V_{in}(\omega)$ at the input port of transmitting antenna. While receiving antenna transfer function $H_{RA}(\omega, \theta, \varphi)$ is the ratio of the output signal of the receiving antenna $V_{oc}(\omega, \theta, \varphi, R)$ to the incident field $E(\omega, \theta, \varphi, R)$. Here, the normalized electric field $E(\omega, \theta, \varphi) = E(\omega, \theta, \varphi, R)Re^{jkR}$, and $V_{oc}(\omega, \theta, \varphi, R) = V_{out}(\omega, \theta, \varphi, R)(Z_l + Z_{in2})/Z_l$, where ω is the angular frequency, k is the free space wave number, θ and φ are the orientation, R is the distance between the transmitting and receiving antenna, and other definitions are shown in Fig. 5-13 (a).

This system is considered as a two-port network as in Fig. 5-13 (b) as $ABCD$ matrix. Because the $ABCD$ parameters describe the relationship of the input and output voltage and current, it can be related to the antenna transfer function mentioned above. However $ABCD$ matrix is difficult to be measured with high accuracy. Therefore instead of using $ABCD$ matrix, S parameters of this two-port network is measured and then converted back to the required $ABCD$ matrix using the conversion relations in [18]. Finally, as derived in [65], the antenna transfer function is related with S parameters as follows:

$$H_{TA}(\omega, \theta, \varphi) = \sqrt{\frac{2S_{21}}{(1+S_{11})(1-S_{22})}} \frac{j\omega}{2\pi C_0} Re^{jkR}, \quad (5.2)$$

$$H_{RA}(\omega, \theta, \varphi) = \sqrt{\frac{2S_{21}}{(1+S_{11})(1-S_{22})} \frac{2\pi C_0}{j\omega} \text{Re}^{jkR}}. \quad (5.3)$$

Where S parameters are referred to the two-port network in Fig. 5-13 (b), and C_0 is the light velocity in free space.

Recall the aim of this subsection is to check the compliance of the active tag with FCC regulation. In the following the transmitting and receiving antennas are aligned along their main lobe with maximum gain. In practical measurement, it is difficult to directly measure the radiated signals in the far-field. However, as discussed above, it can be related with the measured signals $V_{out}(\omega, \theta, \varphi, R)$ at the load of receiving antenna, which generally is the output pulse measured by an oscilloscope, as shown in (5.4) and (5.5),

$$E(\omega, \theta, \varphi, R) \times H_{RA}(\omega, \theta, \varphi, R) = V_{oc}(\omega, \theta, \varphi, R), \quad (5.4)$$

$$V_{out}(\omega, \theta, \varphi, R) = \frac{Z_l}{Z_l + Z_{in2}} V_{oc}(\omega, \theta, \varphi, R) = \frac{1-S_{22}}{2} V_{oc}(\omega, \theta, \varphi, R). \quad (5.5)$$

Therefore, the far-field radiation can be calculated as

$$E(\omega, \theta, \varphi, R) = \frac{V_{oc}(\omega, \theta, \varphi, R)}{H_{RA}(\omega)} = \frac{V_{out}(\omega, \theta, \varphi, R)}{H_{RA}(\omega)} \times \left[\frac{2}{1-S_{22}} \right]. \quad (5.6)$$

In the following, far-field radiation of a compact IR-UWB transmitter demonstrated in Fig. 5-12 is measured, by employing a transmitting-receiving time-domain measurement system, which is the same with prototype demonstrated in Fig. 5-13.

(a)

(b)

Fig. 5-14 Received signal: (a) time-domain response, and (b) frequency response.

The transmitter in Fig. 5-12 includes a pulse generator with a low-power microcontroller, and battery for power supply, with details in [8]. It is integrated with the above proposed CP antenna. Following the system setup in Fig. 5-13, a high-gain TSA antenna [63] shown in Fig. 5-15 (a) is employed as a receiver at 1 meter away (in the far field range) to capture UWB signals to the oscilloscope, which mean the load Z_L in Fig. 5-13 is 50 ohms.

(a)

(b)

Fig. 5-15 Setup for antenna transfer function measurement: (a) TSA prototype, and (b) system setup.

According to the FCC regulation, UWB systems must comply with stringent EIRP which is defined as $EIRP = P_T(f) \cdot G_T(f)$, where $P_T(f)$ is output power of the transmitter just before antenna (i.e., the input power of transmitting antenna under perfect matching), and $G_T(f)$ is the maximum gain of the transmitting antenna over all possible directions. Therefore, it is worth to mention that, in the measurement setup, the transmitting and receiving antennas are aligned along their main lobe with maximum gain. The measured pulse waveform at the oscilloscope $V_{out}(\omega, \theta, \phi, R)$ in (5.6) is obtained and plotted in Fig. 5-14 (a), with its frequency response in Fig. 5-14 (b).

The second step is to obtain the $H_{RA}(\omega, \theta, \phi)$ and S_{22} in equation (5.6). As shown in Fig. 5-15 (b), two identical TSA antennas [63] in Fig. 5-15 (a) with the same polarization are connected to two ports on Agilent N5244A network analyzer sweeping up to 8 GHz, and used as transmitter and receiver with distance of 1 meter between each other. Using equation (5.3), the transfer function of receiving antenna is then calculated from the measured S-parameters.

Fig. 5-16 plots both the amplitude and phase performance of antenna transfer function up to 8 GHz. The measured results in Fig. 5-16 show flat response and linear phase variation in the pass band, showing that the generated signals are distorted slightly.

(a)

(b)

Fig. 5-16 The TSA antenna transfer functions: (a) amplitude, and (b) phase.

Fig. 5-17 Power spectral density (PSD) of the integrated IR-UWB transmitter.

With equation (5.6) and all values obtained above, the far-field radiation of the IR-UWB transmitter is calculated and compared with FCC mask regulation in Fig. 5-17. It is clear that this UWB transmitter integrating pulse generator and CP antenna is sufficient to meet the indoor FCC regulation in the far field. Different from conventional transmitter employing bandpass filter for shaping frequency response, this transmitter just utilizes the Gaussian pulse generator and an antenna with bandpass filtering characteristic. Therefore, it is a good candidate to form a highly-integrated transmitter with compact size and high power for IR-UWB applications.

5.2 Square-Slot Antenna for UWB-RFID Tag Embedded in Concrete

Following the square-slot antenna for free-space application, this antenna is re-designed for UWB-RFID embedded in Concrete. The aim is to employ the embedded tags for Concrete debris tracking under high explosions.

Traditionally, high-speed digital video cameras are employed to record the explosion process. Several of them are used during the tests for different functions: some are dedicated to register the break-up process of the Concrete wall, while others record the launch velocity and angles, or any possible bounce and roll of debris. All images are used to compose a break-up pattern. Finally, post processing is made to determine debris mass size distribution and location. However, the usage of cameras has many limits as it cannot capture debris tracking in smoke, or behind some obstacles. Furthermore, because of the large amount of data registered in cameras, it becomes difficult to evaluate comparisons like launch angles versus debris size and loading density, or debris velocity versus debris size and travelling distance. Therefore, we try to employ UWB technology in this real-time location tracking of debris. Hundreds of compact UWB tags (transmitters) are embedded in Concrete before explosion. Instead of using high speed camera to capture images, UWB sensors (receivers) at hundreds of meters away are employed to receive signals transmitted by the tags. Subsequently, the received raw data are processed by positioning algorithm

for debris identification and trajectory tracking. Compared with conventional method, there are a lot of advantages, as it is immune to multipath effect and independent of weather and lighting conditions.

However, there are many challenges. Firstly, the tags integrated with antennas are buried in Concrete, which has some effect on the antenna performance, as it is in antenna's near-field region. Therefore, the dielectric property of Concrete and its effect on the antenna performance should be investigated. Secondly, the active tags should be compact to enhance the chance that the fragmented Concrete wall in explosion will have debris bigger than the tag size, so the complete tag can be intact in debris. Thirdly, the transmitted power should be high to enhance the debris detection range, and the low power is necessary due to the limited battery size in tags.

To focus on the topic of this thesis on UWB passive component designs, this section introduces Concrete dielectric property measurement and the design procedure of embedded UWB antennas. For the developed debris tracking systems, detailed information can be found in [62, 66].

5.2.1 Dielectric properties of concrete

As shown in Fig. 5-18, the measurement is carried out in an enclosed room using a large moving metallic screen with a transmission window of $40 \times 40 \text{ cm}^2$ in the

center. The metallic screen is covered with electromagnetic wave absorbing materials to reduce scattering and reflections from areas outside the transmission window.

Fig. 5-18 The measurement setup [60].

In the test, a Concrete slab with various thicknesses of 4 cm, 7 cm, and 10 cm to simulate the Concrete wall is used to cover the transmission window. A vector network analyzer (VNA) is used to sweep the frequency range and measure S-parameter to obtain the frequency response. The measurement frequency sweeps from 2 to 6 GHz which covers the desired investigation frequency of 3-5 GHz in our project. Two transversal electromagnetic (TEM) horn antennas are fixed at each side of a screen as the transmitter and receiver, respectively. The distance

between the two horn antennas is 2.1 m, large enough to make the screen to be located in the far-field of each antenna and make the incident electromagnetic signal on the Concrete slab to be a plane wave.

The details about the measurement and signal processing are illustrated in [60]. Finally, an insertion transfer function is obtained through two free-space measurement results, which are $S_{21}(f)$ in the presence of Concrete slab to cover the window and $S_{21}^{fs}(f)$ without the Concrete slab. The dielectric constant (ϵ_r) and the loss tangent ($\tan \delta$) of three Concrete slabs are then calculated using a single-pass analysis method. The results are shown in Fig. 5-19 and Fig. 5-20.

Fig. 5-19 Dielectric constant of Concrete slabs [60].

Fig. 5-20 Loss tangent of Concrete slabs [60].

There are mainly two reasons for these variations of different concrete thickness. Firstly, there are many materials used to fabricate a Concrete slab, such as sand, stone, water, air bubbles, and so on. It is very difficult (if possible) to fabricate three identical Concrete slabs. The inhomogeneous density of materials inside each part of the Concrete slab will also incur the measured difference. In addition, the standard dielectric constant of water is up to 78.5, the dielectric property is quite sensitive to water content in the Concrete slab.

5.2.2 Antenna geometry and effect of Concrete

The proposed antenna with compact size of $25 \times 25 \text{ mm}^2$ is printed on a square RO4003C substrate, with the similar structure as that one in free space discussed in Section 5.1, except a two-step impedance transformer shown in Fig. 5-21.

The dimensions of the final antenna design are: $G = 25\text{mm}$, $L = 17\text{mm}$, $W_t = 7\text{mm}$, $L_t = 10\text{mm}$, $L_x = L_y = 3\text{mm}$, $n = 0.5\text{mm}$, $W_f = 3\text{mm}$, $g = g_2 = 0.3\text{mm}$, $m = 7\text{mm}$, and $p = 1.4\text{mm}$. It is fabricated and embedded in Concrete with thickness of 4 cm as shown in Fig. 5-21.

Using the measured dielectric property of Concrete in Section 5.2.1, the effect of Concrete on the antenna performances is studied through simulations assuming another dielectric layer covering the antenna with dielectric constant of 8.16 and loss tangent of 0.33, which are the averaged values obtained from three Concrete samples. However, as discussed above, the dielectric constant and loss tangent values may vary according to frequency and different fabrication. Therefore, it is important to ensure that the performance of the antenna is acceptable over the range of variation. Keeping this in mind, the sensitivity of the antenna performance on dielectric constant and loss tangent variations is also studied by simulation as follows.

(a)

(b)

Fig. 5-21 Geometry (a) of the slot antenna and photo (b) after embedded in Concrete [62].

Fig. 5-22 Simulated return loss of the UWB slot antenna embedded in Concrete with different complex dielectric constant [62].

Fig. 5-23 Simulated return loss of the UWB slot antenna embedded in Concrete with different loss tangent [62].

As shown in Fig. 5-22, the resonant frequency shifts to lower frequency by increasing dielectric constant. The resonant frequency changes by 20% when the dielectric constant changes by 38%. Meanwhile, in Fig. 5-23, the return loss of the proposed antenna is not very sensitive to loss tangent variations, except at the resonant frequency. The results also show that the effect of Concrete is two-fold. On one hand, due to its high $\tan \delta$, the Concrete will obviously attenuate the antenna radiated signal strength. On the other hand, compared with the free-space antenna design, the antenna embedded in Concrete can be smaller due to larger ϵ_r . The Concrete works as a superstrate in the antenna design.

5.2.3 Antenna measurement results

As shown in Fig. 5-24, the measured 10-dB return loss bandwidth is similar to simulated one which is from 2.7 to 4.5 GHz, or about 50% referenced to the center frequency of 3.6 GHz. The discrepancies between simulated and measured results are due to the fabrication tolerance, the additional coaxial feeding line shown in Fig. 5-21, and the abovementioned fact that the surrounding Concrete are not exactly with $\tan \delta$ of 0.33 and ϵ_r of 8.16.

Similar setup as shown in Section 5.1.3 is established in Fig. 5-25, with the proposed antenna as the receiver and the tapered slot antenna [63] as the transmitter.

Fig. 5-24 Measured and simulated return loss of the slot UWB antenna embedded in Concrete [62].

Fig. 5-25 Time domain measurement setup [62].

(a)

(b)

Fig. 5-26 Measured waveforms received by the antenna embedded in Concrete:
(a) varying ϕ whereas $\theta = 0^\circ$, (b) varying θ whereas $\phi = 90^\circ$ [62].

The transmitting antenna is rotated in the x - y plane (varying ϕ) to determine the circular polarization properties of the receiving antenna. Subsequently, the receiving antenna is rotated in the x - z plane (varying θ) to determine the azimuth plane radiation pattern of the receiving antenna. In the x - y plane of the proposed antenna, the measured results in Fig. 5-26 (a) show that the amplitude and waveform vary with ϕ from 0° to 180° . The maximum normalized peak-to-peak amplitude is 2 at $\phi = 90^\circ$, while the minimum one is 0.986 (-6.143 dB) at $\phi = 0^\circ$. Fig. 5-26 (b) shows the measured radiation variation in the x - z plane of the embedded antenna. The largest normalized peak-to-peak amplitude of 2 occurs at $\theta = 0^\circ$, decreasing to 1.502 (-2.487 dB) at $\theta = 40^\circ$, and finally reaches the minimum one of 0.923 (-6.717 dB) at $\theta = 90^\circ$.

These time domain measurement results demonstrate that the proposed antenna can receive signals both in the x - y and x - z planes, which will improve the signal reception for different orientation and direction of the debris with respect to the receivers.

5.2.4 Square-slot antenna integrated with active tag in Concrete

As demonstrated in Fig. 5-27 and Fig. 5-28, the developed antenna is integrated with active UWB tag and then embedded in Concrete block of $7 \times 7 \times 7 \text{ cm}^3$. The whole UWB active tag consists of the microcontroller unit (MCU), multi-vibrator, UWB impulse generator, pulse shaping network, two amplifiers, together with

the proposed antenna. One most distinct advantage of the active UWB tag is its compact size, which is achieved by employing multilayer technology. Moreover, Strip line rather than Microstrip line or CPW is adopted to dramatically reduce the layout size of the radio frequency (RF) front-end, especially the pulse shaping network and two amplifiers parts. Another key advantage of the tag is its low power consumption with high output power. In order to power up the active tag from battery with small size, the pulse repetition frequency (PRF) of the active tag is chosen to be as low as 200 kHz. Reduction in PRF significantly results in low power consumption. Furthermore, both the pulse width and pulse delay are carefully controlled to maximize the final output amplitude of the active tag and minimize the power consumption.

The received signal of the tag embedded in Concrete is shown in Fig. 5-29 using the same measurement setup in Fig. 5-25 and the distance between the tag and the receiver is 1 meter away. Fig. 5-29 (a) shows the received pulse characteristics with the tag oriented for maximum peak-to-peak amplitude of around 116 mV. Fig. 5-29 (b) shows the received pulse characteristics with the tag oriented for minimum peak-to-peak amplitude of around 56 mV. This decrease in amplitude of 48.27% (-6.32 dB) due to tag orientation matches with the measurement results of the proposed antenna in Section 5.2.3. In addition, the measurement results show that in the maximum tag antenna gain direction, the total loss at 1 meter distance is 50.3 dB. This loss is 4.8 dB more than the free space loss of 44.5 dB. This extra loss is reasonable due to the loss in Concrete.

Fig. 5-27 The prototype of active UWB tag [62].

(a)

(b)

Fig. 5-28 The fabricated active UWB-RFID tag: (a) integrated design covered by Silicone rubber, and (b) an active tag embedded in Concrete [62].

(a)

(b)

Fig. 5-29 Received pulses when the transmitter is embedded in Concrete: (a) the active tag is co-polarized with TSA receiver antenna, (b) the active tag is cross-polarized with TSA receiver antenna [62].

Compared with the time-domain measured results shown in Fig. 5-26, the pulse duration shown in Fig. 5-29 is lengthened quite a lot. There are mainly two

reasons. Firstly, the antenna is designed embedded in Concrete directly with no other cover. This design is validated by the measured results shown in Fig. 5-26. However, when the antenna is integrated with active tag shown in Fig. 5-28 (a), this tag is covered by Silicone rubber first and then embedded in Concrete. Therefore, the characteristics of this antenna will be changed a lot and poor impedance matching is induced. Secondly, due to the fabrication tolerance, it is not very easy to obtain the desired results compared with simulation ones with such a small size constrain to integrate the whole tag. Therefore, the multiple reflections of interior function blocks of active tag may also increase the pulse duration. More work need to be done to optimize the output pulse in our future work.

The embedded UWB tag is integrated with active tag of whole size $25 \times 25 \times 25 \text{ mm}^3$ and then embedded in Concrete. The developed debris tracking system is introduced in [62, 66], which include four UWB sensors, one locator, and tracking algorithm. Measurement results show that the maximum detection range between the active tag and four UWB sensors can be more than 190 meters.

Chapter 6

A UWB-RFID System Using Circularly Polarized Chipless Tag

In this Chapter, a UWB-RFID system using circularly polarized chipless tag is proposed, analyzed, and demonstrated. The research work in this chapter is published in [67]. The chipless tag employs the square-slot UWB antenna introduced in Chapter 5. Different from some related research done by other team members in our group [13-15], this Chapter presents a novel UWB-RFID system with two main innovations listed as follows:

- 1) It is the first time to use CP chipless tags for UWB-RFID demonstration. The CP characteristics significantly reduce the mutual coupling between transmitter and receiver, by placing the antennas with different orientation. The proposed method also reduces the signal ratio between backscattered structural and antenna modes, which improves range and localization accuracy.
- 2) This Chapter presents a novel processing scheme to distinguish the overlapped structural and antenna modes. This scheme enables the system to operate at lower frequencies resulting in reduced propagation loss. This in turn helps to increase the range and localization accuracy. It also facilitates using shorter delayed lines to design more compact tags.

6.1 Backscattering Characteristics of UWB Antennas

The underlying theory of the chipless UWB-RFID system is based on the backscattering characteristics of UWB antennas. As investigated in [68], the backscattered E-field signal $\vec{E}_s(Z_l, f)$ of an antenna can be expressed by

$$\vec{E}_s(Z_l, f) = \vec{E}_{st}(Z_c^*, f) + \frac{Z_c^*(f) - Z_l(f)}{Z_c^*(f) + Z_l(f)} I(Z_c^*, f) \vec{E}_r(f), \quad (6.1)$$

where f is the operating frequency, $Z_c(f)$ and $Z_l(f)$ are the characteristic impedance and load termination of the antenna, respectively. $\vec{E}_{st}(Z_c^*, f)$ and $I(Z_c^*, f)$ are the backscattered structural mode E-field and the induced current

when the antenna is terminated with its conjugate characteristic impedance $Z_c^*(f)$. $\vec{E}_r(f)$ is the radiated E-field in the case of unit current input. Equation (6.1) clearly shows that the backscattered signals consist of two parts: the first part is the radiated field of a match-loaded antenna, which is called structural mode (σ_{str}); and the second part is a function of the load impedance and known as antenna mode (σ_{ant}). As shown in (6.1) and defined in [68], the structural mode (σ_{str}) is only dependent on the physical shape, size, and material of an antenna. This mode is an inherent feature of a physical object, whether it is an antenna or not. On the other hand, antenna mode (σ_{ant}) indicates the fact that, an antenna is a special design to transmit and/or receive RF energy. It is directly determined by the load termination Z_l of an antenna. In addition, the antenna mode is also dependent on the antenna gain (G) [69, 70] and is given by:

$$\sigma_{ant}(f) = \frac{|\Gamma(f)|^2 \lambda^2 G^2(f)}{4\pi} = |\Gamma(f)|^2 \frac{c_0^2 G^2(f)}{4\pi f^2}, \quad (6.2)$$

where $\Gamma(f) = (Z_c^* - Z_l) / (Z_c^* + Z_l)$ is the reflection coefficient at antenna termination, λ is the wavelength according to specified operating frequency f , and c_0 is the light velocity in free space. Therefore, the signal amplitude of antenna mode can be increased by enlarging the antenna gain and/or reducing the operating frequency, which helps to reduce the signal ratio of structural to antenna mode to improve the localization range and accuracy.

As investigated in [14, 67], the structural and antenna modes can be described

and measured both in the frequency and time domains. As shown in Eq. (6.1), these two modes are linearly super-positioned in the frequency domain. It can be converted to the time domain using inverse fast Fourier transform (IFFT). Since IFFT is a linear operator, these two modes are still linearly super-positioned in the time domain. Therefore, frequency-domain equations (6.1) and (6.2) are directly related to the UWB time domain signals. Based on these two classic frequency-domain equations, some predictions are listed below:

- 1) Theoretically, the backscattered signal of a UWB antenna consists of two pulses, i.e., structural mode and antenna mode;
- 2) Covering the whole UWB band, when the antenna is terminated with a conjugate-match, i.e., $Z_l = Z_c^*$, then $\Gamma = 0$, which means there is no antenna mode, $\sigma_{ant} = 0$. On the other hand, when the antenna is terminated with the open-circuited ($Z_l = \infty$, then $\Gamma = 1$) or short-circuited ($Z_l = 0$, then $\Gamma = -1$) loads, respectively, the antenna mode reaches its maximum amplitude;
- 3) In time domain, there is a 180° phase difference between the open-circuited and short-circuited cases.

The above predictions are the theoretical cornerstones of the chipless UWB-RFID system. Based on the interesting backscattering characteristics of an antenna, it becomes feasible to use a single UWB antenna as a chipless tag. Following (1), it is feasible to control the time interval between the UWB pulses of the structural and antenna modes, i.e., employ the pulse position modulation (PPM) to generate more ID codes. Introducing antenna feeding line with different

length is a viable option. Meanwhile, based on (2) and (3), a set of tag ID information can also be carried by the UWB antenna with different terminations, and then be extracted from the detected backscattering signals.

6.2 System Building Blocks

Fig. 6-1 UWB-RFID systems using chipless tags [67]: (a) conventional linearly polarized system, (b) the proposed circularly polarized system, and (c) prototype of a CP system.

Fig. 6-1 (a) shows a typical UWB-RFID system investigated in [14]. In this system, linearly polarized chipless tags are adopted. Therefore, in the UWB-RFID reader, there is strong coupling between the transmitting and receiving antennas that are adjacent and co-polarized. This strong coupling may block the reader from receiving the weak signal backscattered from a tag. Furthermore, a UWB antenna features a backscattered signal with two different pulses, i.e., structural and antenna modes. This characteristic is very useful for identification and localization by using the structural mode pulse as a reference signal. However, the amplitude difference between these two pulses is usually very high, for example, up to 26.1 dB for a slot antenna and 25.6 dB for a monopole in [13]. This amplitude difference brings another challenge for receiver design.

To circumvent the above two challenges, a UWB-RFID system is developed and shown in Fig. 6-1 (b) and (c), which highlights the main differences between the typical and our developed UWB-RFID systems. Firstly, CP tags with short delay lines rather than LP tags with long delay lines are employed. Secondly, the receiving antenna in the UWB-RFID reader is rotated by 90° to reduce the coupling between the transmitting and receiving antennas with linear polarization. More details of the UWB-RFID system including reader, and chipless tags are described below.

6.2.1 UWB-RFID reader

The UWB-RFID reader operates in lower UWB band with center frequency of 4.1 GHz. It consists of several functional blocks developed by researchers in our group [8, 63], such as an impulse radio UWB transmitter, a receiver, and two identical reader antennas. The transmitter consists of an IC chip packaged on a printed circuit board. Its output is an eighth-order UWB pulse with large peak-to-peak amplitude up to 6.4 V and pulse duration of around 1 ns. The time-domain pulse has a clean waveform with little ripple at the end. This characteristic is very important to distinguish clutter and background noise existing in the backscattered signals. The receiver includes three low noise amplifiers with total gain of 36 dB, and two filters for interference rejection. Reader antennas are two identical linearly-polarized tapered slot antennas for transmitting and receiving, respectively. The antenna is with size of $126.2 \times 153 \times 0.508 \text{ mm}^3$, and 10-dB return loss bandwidth covering 2.65-12.9 GHz. It has a high and flat gain of around 10 dBi.

6.2.2 Chipless tag

Square-slot antenna introduced in Section 5.1 is used here as chipless tag with circular polarization. Compared with chipless tags in [14], there is no meandrous transmission line integrated at the output port of the antenna.

To validate the proposed UWB-RFID system, as shown in Fig. 6-2, two types of tags are chosen in the measurement for comparison. They are described as follows:

Type A: square-slot CP antenna with SMA male-to-male right-angle adaptor whose length is $L=3.5$ cm. It is then terminated with VNA calibration kits, i.e., matched-load, open-circuit, and short-circuit.

Type B: square-slot CP antenna with SMA-to-SMA cable whose length $L=9.5$ cm. It is then terminated with VNA calibration kits, i.e., matched-load, open-circuit, and short-circuit.

(a)

(b)

Fig. 6-2 Chipless CP tag: (a) Type A, and (b) Type B.

Here, VNA calibration kits are employed as the three popular loads (matched-load, open-circuit, and short-circuit load). Obviously, the only difference between

these two type tags is the transmission line length which can be used for pulse-position modulation.

Moreover, from system point of view, the backscattered UWB pulses of structural and antenna modes are preferred with similar amplitude. This square-slot antenna is suitable to reduce the amplitude ratio of structural to antenna mode, because its radiation element is a slot with copper only in the center and surrounding the square slot. The absence of metal in the relatively large slot gap helps to reduce the signal amplitude of the structural mode.

6.3 Measurement Results and Discussion

6.3.1 Four-step time-domain processing scheme

In the following, a platform shown in Fig. 6-1 is set up to demonstrate the proposed chipless UWB-RFID system. To mimic the practical application environment, this demonstration system is build up in our lab instead of an anechoic chamber. In this case, all the clutter and background noise in the surroundings are taken into consideration.

The transmitter is powered up by 2 batteries with AA size, and the receiver board uses the Instek GPS-3303 as DC power supply. Two TSA antennas are connected with the transmitter and receiver, and placed in the vertical polarization and

horizontal polarization planes respectively, which are aligned with the bore-sight of the chipless tag attached to a large surface of foam. A gap of $g = 0.105$ m is chosen to separate the transmitting and receiving antennas, and the distance between the reader and passive tag is $R = 0.5$ m. The backscattered signals are displayed on Agilent DSO81204B oscilloscope.

Fig. 6-3 shows measured boresight backscattered pulses of the two types of CP passive tags with different loading condition. These pulses are overlapped together, and the pulse shape is quite different from the waveforms obtained in [14]. In [14], there are two separate pulses (i.e., structural and antenna modes) which are easy to distinguish.

(a)

(b)

Fig. 6-3 Measured bore-sight backscattered pulses using 4.1 GHz transmitter: (a) type A tag, and (b) type B tag [14].

Different operating frequency and chipless tags with different delay line length, are the two main reasons. Take the type A CP tag as an example. Theoretically, the return time for antenna mode pulse travelling in the adaptor is:

$$\Delta t = \frac{2L}{C / \sqrt{\epsilon_r}} = \frac{2 \times 0.035}{3 \times 10^8 / \sqrt{2.1}} \approx 0.34 \text{ ns}, \quad (6.3)$$

where L is the length of adaptor, C is the light velocity in free space, and ϵ_r is the dielectric constant of inner material of the adaptor. This return time (0.34 ns) is shorter than the transmitted pulse duration (around 1 ns as introduced in Section 6.2). Therefore, the structural mode signal, which does not travel through the

right-angle connector, is caught up by the antenna mode pulse (which travel through the connector), and then are overlapped together in time domain. Similarly, tag B in Fig. 6-2 with delay length of 9.5 cm introduces 0.92 ns time delay, which is also shorter than the transmitted pulse duration. On the contrary, the UWB transmitter employed in [14] is with a higher operating frequency of 7.5 GHz and its output pulse is narrow at around 0.35 ns. Moreover, the two chipless tags designed in [14] are with delay line length of 37.6 mm and 41.6 mm. Employing equation (6.3), the corresponding delay time is 0.36 ns and 0.398 ns, which are longer than the transmitted pulse duration.

Fig. 6-4 The proposed processing scheme to distinguish structural and antenna modes in backscattered pulses [14].

The purpose to employ UWB transmitter with lower center frequency is to reduce propagation loss and increase the tag antenna mode response. Hence it improves the range and localization accuracy. In order to separate the backscattered structural and antenna modes for identification, two straightforward methods are employing a long delay line and shortening the incident pulse [14]. However, the former makes the tag bulky whereas the latter is limited by frequency regulation.

In this thesis, a time-domain processing scheme is proposed in the following as shown in Fig. 6-4, where R represents the received time-domain response. Step 1 is used to measure the background clutter for background calibration. In this step, there is no tag in the test environment. The received pulse is a combination of clutter ($S_{clutter}$) and coupling ($S_{coupling}$) pulses between the transmitter and receiver in the reader, which remain the same in the four steps. The following three steps are used to measure passive tags with different loads, i.e., matched-load, open-circuit, and short-circuit. In these 3 steps, the structural mode is independent of the antenna termination, whereas the antenna mode varies according to different loads.

Based on the theoretical analysis in Section 6.1, the antenna mode does not exist in the second matched-load step and has 180° phase difference in the open-circuit and short-circuit steps. Therefore, ideally, we can get two separate pulses as illustrated in Fig. 6-4, i.e., structural mode (S_{str}) and antenna mode ($S_{ant_o/c}$) for

open-circuit, and $S_{ant_s/c}$ for sort-circuit). In addition, these two antenna modes should have a 180° phase difference as analyzed in Section 6.1.

This proposed scheme is not only useful in detecting the overlapped structural and antenna modes, but also simplifies the chipless tag designs. In previous work [14], long transmission line is necessary to generate enough time delay to separate the structural and antenna modes, as least to distinguish the two peak positions, in order to use the pulse-position modulation. This creates design difficulties and enlarges the tag size. Moreover, the signal amplitude of the antenna mode is further reduced after going through the long delay line, thus the amplitude ratio of structural to antenna mode becomes worse. Therefore, with this proposed time-domain measurement method, simple and compact passive tags can be identified and localized, and more ID information can be embedded in the backscattered signals.

6.3.2 Comparisons and discussion

Following the four-step processing scheme, measurement is done to extract both structural and antenna modes from backscattered signals in Fig. 6-3. The subtracted structural and antenna modes are given in Fig. 6-5 and Fig. 6-6 for these two types of passive tags respectively. They are plotted in two separate figures for easy reading and comparison. It clearly shows that the overlapped time-domain waveforms in Fig. 6-3 are distinguished as two separate pulses. The

time interval between these two pulses is around 0.5 ns for type A tag in Fig. 6-5, and around 1.17 ns for type B tag in Fig. 6-6.

The measured time interval is longer than the theoretical value calculated in equation (6.3). The difference is mainly contributed by the feed line of the chip-less CP tag as well as the inherent phase delay between structural and antenna modes. Fig. 6-5 (b) and Fig. 6-6 (b) validate the expected 180° phase difference between the antenna modes when the chip-less CP tag is terminated by open-circuit and short-circuit.

As mentioned before, the system configuration shown in Fig. 6-1 (b) can reduce the amplitude ratio between structural and antenna modes. The backscattered structural mode should be lower than that in Fig. 6-1 (a). Meanwhile, the antenna mode will remain the same due to the CP performance of the chip-less tags. In order to verify this advantage, the same setup in Fig. 6-1 (b) is adopted but just changing the Rx antenna to be cross-polarized with Tx antenna. The extracted waveforms are shown in Fig. 6-7 and Fig. 6-8. Compared with Fig. 6-5 and Fig. 6-6, obviously, the main difference is the increased signal amplitude of structural mode. That is because that the polarization between the transmitting and receiving antennas is the same. Therefore, more backscattered structural mode is received with less loss.

(a)

(b)

Fig. 6-5 Structural (a) and antenna (b) modes of type A tag with horizontally polarized receiving antenna [67].

(a)

(b)

Fig. 6-6 Structural (a) and antenna (b) modes of type B tag with horizontally polarized receiving antenna.

(a)

(b)

Fig. 6-7 Structural (a) and antenna (b) modes of type A tag with vertically polarized receiving antenna [67].

(a)

(b)

Fig. 6-8 Structural (a) and antenna (b) modes of type B tag with vertically polarized receiving antenna.

Table 6-1 Comparisons of signal amplitude

Type (passive tag)	Polarization	V1 (mV)	V2 (mV)	V3 (mV)
Type A (Fig. 6-5 & Fig. 6-7)	Tx and Rx (cross-polarized)	30.2	19.4	26.4
	Tx and Rx (co-polarized)	106.2	28.6	35.6
Type B (Fig. 6-6 & Fig. 6-8)	Tx and Rx (cross-polarized)	39.1	24.8	25.3
	Tx and Rx (co-polarized)	104.2	32.1	31.2

Notes: V1 is the peak-to-peak signal amplitude of structural mode

V2 is the peak-to-peak signal amplitude of antenna mode (open-circuit load)

V3 is the peak-to-peak signal amplitude of antenna mode (short-circuit load)

As summarized in Table 6-1, for type A tag, by employing the cross-polarized Tx and Rx antennas, the peak-to-peak signal amplitude of structural mode is reduced from 106.2 mV to 30.2 mV (around 10.9 dB). On the other hand, the signal strength of the antenna mode does not change so much. The value reduces from 28.6 mV to 19.4 mV (3.4 dB) with open-circuit load, and reduces from 35.6 mV to 26.4 mV (2.6 dB) with short-circuit load. Similar phenomenon is observed for type B tag. Its peak-to-peak structural mode amplitude is reduced from 104.2 mV to 39.1 mV (around 8.5 dB), while the antenna mode amplitude reduces from 32.1 mV to 24.8 mV (2.2 dB) with open-circuit load, and from 31.2 mV to 25.3 mV (1.8 dB) with short-circuit load. Therefore, by adopting the CP configuration,

the amplitude ratio between structural and antenna modes is greatly decreased to alleviate the difficulty of a receiver design in a UWB-RFID reader.

The abovementioned reduced signal ratio is very useful and important in practical application. It is well known that the structural mode usually occupies most of the energy in the backscattering waves, typically two times of the signal strength of the antenna mode. It is even worse when the chipless tag is attached to large reflecting objects. Therefore, if the signal ratio of structural to antenna mode can be suppressed by using a cross polarized receiving antenna, it will improve the positioning and detection accuracy, and also alleviate the complexity of the receiver design.

Overall, the proposed novel chipless UWB-RFID system has more advantages over previous systems. By employing the CP passive tags and cross-polarized transmitting and receiving antennas, the signal strength of structural mode in the backscattered signal can be significantly reduced, while that of the antenna mode is little affected. Meanwhile, with the proposed four-step time-domain measurement based processing scheme, it is possible to distinguish the structural and antenna modes with close time spacing. It therefore simplifies the passive tag with shorter delay lines, and increases the signal amplitude of the antenna mode. This proposed chipless UWB-RFID system with CP tags provides a good option for the next-generation RFID applications.

Chapter 7

Conclusions and Recommendations

7.1 Conclusions

For UWB-RFID systems, this thesis investigates novel passive components including filters and CP antennas. Following these designs, for the first time, this thesis demonstrates a UWB-RFID system using chipless CP tag and time-domain processing scheme.

Bandpass filters are proposed for UWB-RFID reader operating within 3.4 – 4.8 GHz which follows Singapore's IDA regulation. Compared with conventional

CRLH-TL filters, the proposed unit cell introduces shunt inductor and capacitor to achieve a resonance at WLAN frequencies. Following theoretical analysis in Section 3.1.1 of this thesis, we show that the proposed unit cell is a special case of the conventional CRLH unit cell. The theoretical analysis also provides a design guideline. The proposed compact filters are fabricated on three-layered PCB. Measured results demonstrate that the rejection at 2.4 and/or 5.8 GHz is up to 60 dB, and the passband insertion loss is around 1.1 dB. Following that, an absorptive filter integrating both the bandpass and bandstop characteristics is designed to pass the desired in-band signals, and also absorb the reflected out-of-band signals. This absorptive filter improves the pulse shape of the transmitter and also maintains the stability of the UWB source.

UWB antenna with circular polarization is very important to increase system flexibility. However, most published circularly polarized antennas are narrowband. This thesis investigates a spiral-inspired UWB antenna. Different from conventional spiral antennas which require complicated feeding networks, the proposed antenna is planar and directly fed by a simple Microstrip line. It features wide bandwidths for both impedance and axial ratio, covering 6 to 10 GHz for the upper UWB band. Based on this one-port quasi-spiral antenna, a two-port antenna is also proposed for wireless-powered UWB-RFID system. Conventionally, two separate antennas are required: a narrowband antenna for harvesting RF energy to power up the tag, and a UWB antenna for signal transmission. This thesis presents a two-port antenna which physically integrates these two separate antennas in the

same volume. The measured electrical isolation between these two ports is higher than 20 dB covering 1-8 GHz.

Another UWB CP antenna, i.e., a square-slot antenna, is also investigated for the UWB-RFID active tags. The antenna has compact size of $35 \times 35 \times 0.508 \text{ mm}^3$ and operates in 3-5 GHz. It is integrated with active circuits to form an active UWB-RFID tag. This tag satisfies FCC indoor frequency regulation according to the far-field radiation measurement. Apart for free-space application, the square-slot antenna is also investigated for an active tag embedded in Concrete. The dielectric properties of Concrete including dielectric constant and loss tangent are measured and employed in the embedded antenna design. After integrated with active circuits, the embedded tag features a total size of $25 \times 25 \times 25 \text{ mm}^3$, and with good reading distance.

Using the square-slot antenna as a chipless CP tag for the first time, a novel UWB-RFID system is proposed and built up for demonstration. The CP chipless tag facilitates reducing the signal ratio between structural and antenna modes to improve system performances. In addition, a four-step measurement scheme is proposed to separate the overlapped structural and antenna modes. This scheme not only simplifies the tag design by using short delay line, but also enables the system to operate at lower center frequency. Measured results show that the structural mode can be reduced up to 10.9 dB, and the overlapped structural and antenna modes can be effectively detected and successfully separated.

7.2 Recommendations for Future Work

Wireless-powered UWB-RFID tag is a promising candidate for many applications especially when battery is not allowed in a tag. More research work is worthwhile to improve the reported wireless-powered systems. It would be interesting to integrate the two-port antenna proposed in this thesis with active circuits for system level demonstration.

Single-fed UWB CP antenna has been realized in this thesis. The measured results demonstrate that it achieves circular polarization in some specific angles in the 3-D domain only, whereas UWB antennas with 3-D omni-directional circularly polarized radiation are highly preferred for many applications. To the author's knowledge, there is little research work on this point. It would be a good research topic.

Last but not least, chipless UWB-RFID system with circular polarization is a very interesting research topic. The whole system as well as the positioning algorithm requires more research work. Many challenging and interesting issues should be carefully considered, especially when a UWB-RFID system targets to identify and locate many chipless tags simultaneously.

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