Chapter 1

Time Reversal for Ultra-wideband Communications: Architecture and Test-bed

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This chapter† presents a wideband radio system design concept that takes advantage of rich multipath propagation. Depending upon the channel information, each pair of transmitter and receiver in the system chooses a transmit waveform that is optimal in some sense. The transmitters are capable of waveform level preprocessing, while the receivers can be rather simple since they do not need special means (like a RAKE combiner) to capture dispersed energy over time, and even equalizers may not be necessary. In particular, the channel impulse response (CIR) can be utilized to provide physical-layer security enhancement thanks to the spatial focusing property of time reversal preprocessing. This design concept is especially attractive to wireless sensor networks (WSN) because (1) low-complexity and low-cost receivers are strongly demanded for any WSN, (2) security is vulnerable in WSNs and enhancement at physical layer is extremely desirable, and (3) deploying WSNs in harsh and scattering radio frequency (RF) environments is quite meaningful and necessary. A few realistic radio channels are studied based on experimental data, aiming at best designing the preprocessing systems. In addition, design and implementation work of an experimental time reversal radio test-bed is reported, verifying the concept of preprocessing plus simple receivers.

1.1. Background

Recent advances in miniaturization, low-power electronics and wireless communications, stimulated by increasing demands for automation in home and industrial areas, have triggered tremendous interests in the wireless sensor network (WSN) research, development and deployment. There are many challenges in WSN design, and some of the challenges are due to tough constraints and conditions posed by specific applications and environments. Examples of these constraints and conditions include power consumption, node simplicity, node cost, non-line-of-sight propagation, severe multipath, and low signal leakage, etc.

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Mainly due to potentially low implementation complexity, suboptimal reception strategies, such as transmitted reference (TR)\textsuperscript{1–3,5,6,15–17} and its variants\textsuperscript{19–21,27,28} as well as energy detection\textsuperscript{22–25} have gotten increasing attention for complexity and cost constrained wideband applications. These schemes may be enhanced to handle multipath and inter-symbol interference (ISI) at the cost of complexity. One philosophy to simplify the receivers without sacrificing overall performance is to shift part of receiver-side functions to the transmitter side, which is meaningful for a centralized network where one powerful central station communicates with a large number of nodes. Preprocessing at the transmitter can be viewed as a performance enhancement to the system where the receivers are simple but perform relatively poorly.

Sensor nodes are often deployed in open areas, thus more vulnerable than those in a wired network in terms of security. The WSN security is still an open field and most proposed security approaches are based on encryption and authentication mechanisms. In other words, majority of them are implemented at the layers above the physical layers.

In this chapter, a wideband radio system design concept that takes advantage of scattering propagation environments is provided. Depending upon the channel information, each pair of transmitter and receiver in the system chooses a transmit waveform that is optimal in some sense. An example is the use of time reversal preprocessing (pre-filtering) at the transmitter to focus the signal at the receiver\textsuperscript{41–51}. In such a system with time reversal preprocessing, the receivers can be very simple, because they do not need special means (like a RAKE combiner) to capture dispersed energy over time, and even equalizers may not be necessary. In particular, the channel impulse response (CIR) can be utilized to provide physical-layer security enhancement thanks to the spatial focusing property of time reversal preprocessing\textsuperscript{48,53,54}.

### 1.2. Suboptimal Schemes Using Low-Complexity Receivers

The functions of a receiver in a communication system include energy capture and demodulation. It is well known that, in terms of maximum signal-to-noise (SNR) ratio at the decision stage, the best reception strategy over an additive white Gaussian noise (AWGN) channel is to match the received data symbol waveform at the receiver. When the channel is frequency selective (due to multipath propagation), the matching process should take into account the multipath channel characteristics. RAKE receiver is such a matching based reception technique that is able to handle multipath distortion.

To demodulate the data, an optimal receiver employs a local template as a reference to correlate with the incoming information-embedded signal, where the template is a piece of received reference symbol waveform. Of course, the condition for optimal reception is that the demodulation reference can be accurately set. Practically the template is obtained from channel estimation. However, in selective fading environments, such as home and machinery areas, multipath turns to rich as the signal bandwidth increases. For instance, in the ultra-wide bandwidth (UWB) scenario, up to a hundred of resolvable paths may be observed from the CIR. This means a RAKE receiver has to contain a large number of fingers.
to achieve acceptable performance, implying a high-complexity, power hungry and expensive solution. Alternatively, orthogonal frequency division modulation (OFDM) is also an efficient means for optimum performance in frequency selective (multipath) channel cases. An OFDM scheme divides a frequency band into a large number of non-frequency-selective subbands and transmits data over these subbands in a parallel basis. Both of RAKE receiver and OFDM can handle multipath very well, but they are not cheap and low-power-consumption solutions, thus they may not be suitable for WSN applications. Motivated by the need for simple wireless receivers with satisfactory performances, a number of old-fashion reception techniques have regained popularity. These suboptimal schemes do not require channel estimation at all and are not quite sensitive to timing error.

Transmitted reference (TR) is a communication scheme with the reference symbols being transmitted along with the data symbols. In order to eliminate the requirement for channel estimation at the receiver, both the message and the reference are transmitted through two dedicated subchannels separated by time or frequency, where the channel separation is small enough to ensure coherence between the two subchannels. The information bits may be modulated in pulse position modulation (PPM) or antipodal format. The received up-to-date reference signal is directly used to demodulate the data. Fig.1.1 describes the time-separated TR scheme. The core of the receiver in a time-separated TR system is a delay line device followed by a correlator. The simplicity is achieved at the cost of performance penalty in the following two aspects. First of all, half of the total transmit power is spending in transmitting the reference. Secondly, the received reference signal is “dirty”–
noise polluted, which results in a non-Gaussian cross-noise term in addition to a Gaussian noise term in the correlator’s output. The non-Gaussian noise is much stronger than the Gaussian noise, leading to significant performance degradation. Aiming at improving the original TR scheme, a number of TR variants have been proposed.

A straight-forward way to reduce the overhead is to reduce percentage of reference symbols, i.e., only one referenced symbol is transmitted for a block of data symbols (called stored reference in\(^4\)), implying that a piece of received reference waveform is stored at the receiver and serves as a reference for demodulating multiple data symbols. However, the waveform storage can be very expensive. To remove the overhead completely, the reference can be provided through differentially encoding, so that a modulated data bit can be demodulated by using the previous symbol as reference. This differentially encoded TR is called autocorrelation demodulation in Simon’s book\(^18\) and DTR in the paper by Chao and Scholtz\(^20\). By eliminating the dedicated reference transmission, the energy efficiency is improved by about 3 dB\(^20\).

The template noise level may be suppressed by averaging (or accumulating) over a few received reference symbols.\(^7\) Averaging can be applied to DTR receivers too if decision feedback is utilized.\(^21\) However, implementing the averaging function at waveform level requires storing multiple received symbol waveforms, which is not cheap for either analog or digital approaches.

![Fig. 1.2. A DTR based multi-symbol detection receiver structure.](image)

Alternatively, performance may be improved by resorting to multi-symbol detection combined with DTR.\(^18,27,28\) Incoming received raw symbols are grouped into blocks of \(L\) consecutive symbols, without overlap between any two consecutive blocks. The received raw symbols are processed using differential demodulation to obtain multiple decision statistics for joint decision. A DTR based multi-symbol detection receiver structure is given in Fig.1.2, and refer to the paper\(^27\) for details.

Note that there is a drawback in all schemes of time-separated TR and its variants:
expensive and bulky delay elements are required at the receivers. This huge disadvantage prevents the time-separated TR schemes from being widely accepted. Without using any delay device, a frequency-separated TR can provide the reference at the cost of frequency expansion.\textsuperscript{15-17} One of proposed frequency-separated TR scheme is called slightly frequency-shifted reference (FSR) technique\textsuperscript{15} suitable for low data rate applications, and implementing an FSR receiver is quite easy.

Different from the template-correlation schemes mentioned above, another family of suboptimal reception schemes includes energy (square-law) detector and envelop detector. With either of the detectors, both on-off keying (OOK) and pulse position modulation (PPM) can be adopted as modulation schemes. One problem with these two unipodal modulation schemes is that they cause spectral spikes due to non-zero mean of the symbol values. Scrambling encoding can remove the spectral spikes and time hopping (TH) may be considered for both multiple user access and spectral spike reduction. The receiver uses an integrator to accumulate signal energy. For better performance the signal can be weighted prior to integration and there must be a best weighting function depending on the signal waveform and the noise level.\textsuperscript{8-12} In fact, implementation of weighting function is not of low complexity and this contradicts the philosophy of low-complexity receiver design. A relatively simpler weighting method is a gating function which is equivalent to the use of a proper integration interval.\textsuperscript{13,14,25,51} A practical implementation of a smart integrator is to control the integrator’s on-duration. Denoted by \( R_b \) the symbol rate and consider a received symbol waveform with most of the energy concentrated in an interval \( T_I \). If \( T_I < T_b = 1/R_b \), then integrating over the interval \( T_I \) outperforms integrating over the interval \( T_b \), since both gather almost the same amount of signal energy but the latter gathers more noise.

All of the discussed suboptimal schemes are of low-complexity in the sense that no channel estimation is required and they are less sensitive to timing error. A common shortcoming is that they are not able to work with typical linear equalization techniques, thus they are not suitable for applications where ISI exists apparently. Unlike linear receiver, the equivalent discrete channels of some suboptimal schemes behave nonlinearly, where an equivalent discrete-time channel has data input at one end and it outputs decision statistic plus noise at the other end.\textsuperscript{25,30,31,51} For both DTR and energy detection, their equivalent discrete channels can be modeled as:

\[
z_k = \tilde{d}_k^T C \tilde{d}_k + \eta_k, \tag{1.1}
\]

where subscript \( k \) is a symbol timing index, \( z_k \) is the decision statistic, \( \tilde{d}_k \) is a data vector whose dimension covers the ISI length in symbol, \( C \) is a matrix determined by the channel, and \( \eta_k \) is a noise term. This means that the signal part in the output of the equivalent discrete channel, i.e., \( d_k^T C \tilde{d}_k \overset{def}{=} S_k \), is a nonlinear function of data vector \( \tilde{d}_k \). As a matter of fact, the equivalent discrete channel represented by \( S_k = \tilde{d}_k^T C \tilde{d}_k \) is a special case of second-order Volterra model.\textsuperscript{29,30} In general, \( z_k \) contains a desired signal and a nonlinear inter-symbol-interference (ISI) component that cannot be well handled by normal linear equalization techniques. This fact suggests the use of some waveform level channel
shortening techniques.

1.3. Waveform Preprocessing

Modern communication systems rely on analogue and digital signal processing to combat various impairments such as noise, channel fading (flat and frequency-selective fading), and interferences. For a given environment, a global optimum system under some sort of criterion delivers the best performance. By “global” we mean joint transmitter-receiver optimization. However, most existing work on the joint transmitter-receiver optimization is at symbol-rate level.32–37 It can be expected that by breaking the symbol-rate constraint and allowing continuous-time or fractional-symbol-level processing, the performance can be improved further. In other words, given a communication channel, we prefer to design a set of transmit waveforms as well as a receiver with proper structure and algorithm, such that the system achieves some sort of optimum under some conditions and constraints. The signal processing goals could be capacity-reaching, maximum SNR, and minimum mean square error (MMSE), etc. In most wireless communication scenarios the channel characteristics are time-varying and up-to-date channel information is only available at the receivers. This is why traditionally signal processing efforts are focused on the receiver side.

It is reasonable to consider single-carrier pulse-based radio links for WSN applications, since they have potential to be of low-complexity if major receiver-side linear processing functions are shifted to the transmitter side. It is well known that low probability of interception and low probability of detection (LPI/LPD) can be easily achieved using pulse-based signaling. Also, a narrow-pulse-based single-carrier system can be used for ranging or penetration radar purposes.

A unique issue associated with wireless communications is frequency-selective fading caused by multipath propagation. As for interferences in radio systems, ISI and inter-user interference (IUI) are typically concerned. Listed below are some receiver-based schemes to handle noise, multipath impact and interferences.

- Matched filter (MF): it is placed at the receiver and matches the given transmit symbol (or chip) waveform; for AWGN noise, it maximizes the SNR at the peak of the MF’s output.

- RAKE receiver: a family of technique to collect the signal energy dispersed over multipath components; an ideal RAKE receiver actually matches the overall received waveform; when major paths are resolvable, a practical RAKE receiver structure can be a combination of a regular chip-level MF and a multipath combiner functioning as a finite impulse response (FIR) filter; a RAKE receiver operates at fractional-symbol (or fractional-chip) rate.

- Equalization: a symbol-rate discrete-time processing to remove or reduce ISI; an equalizer is placed at the receiver, usually taking symbol-rate samples from the
(Receiver-based) multi-user detection: this is a broad range of techniques that remove or reduce IUI; the multi-user detection operates at symbol-rate.

When channel and/or multi-user information is available at the transmitter, it seems no strong reason not to use it. Recently, transmitter based signal processing, or preprocessing, has received increasing attention. Preprocessing can be time-reversal pre-filtering (or pre-RAKE), pre-equalization, multi-user precoding, or some processing for a set of compromised goals. Preprocessing can be used along in a transmitter-centric system where traditional receiver-based signal processing is shifted to the transmitters to simplify the receivers. It can also works with receiver-based signal processing to achieve joint transmitter-receiver optimization.

Our special interest is on preprocessing at fractional-symbol rate (or simply, waveform preprocessing), in conjunction with a suboptimal receiver. Below are two examples of symbol waveforms.

- Pulse shaping: it uses a transmitter-side filter to create a transmit waveform that has desired roll-off spectrum; traditional pulse shapes include raised cosine and truncated sinc functions, etc.

- Maximum-SNR transmit waveform: a transmit symbol waveform to achieve the maximum SNR at the MF’s output; it is an eigenfunction of the CIR autocorrelation (implying the CIR has to be known first); and a homogeneous Fredholm integral equation needs to solved for the eigenfunction with the strongest channel gain.\(^{38,40}\)

The transmit waveform optimization problem can be further stated in details as follows. It is well known that the optimum receiver matches the whole symbol waveform distorted by the channel, not the transmitted symbol waveform. However, from system optimization point of view, such a waveform matching alone is not enough. We can further maximize SNR at the receiver by carefully designing the transmitted waveform.\(^{38,40}\)

Given the channel impulse response \(h(t)\) and fixed transmitted power \(P_t\), we wish to achieve the maximum SNR at the receive by jointly designing the transmitted waveform and a good receiver. This problem has been discussed in\(^{38}\) for communication over troposcatter channels and in\(^{40}\) for radar detection.

Assuming the transmitted pulse \(p(t)\) (to be optimized) is confined to the symmetric time interval \([-T/2, T/2]\). The energy of transmitted pulse is then

\[
E_p = \int_{-T/2}^{T/2} |p(t)|^2 \, dt .
\]  

(1.2)

It follows from detection theory that the best receiver is still a MF matched to the received waveform \(p(t) * h(t)\), where \(h(t)\) is the CIR and “*” denotes convolution operation. The
(maximum) SNR at the output of such a MF is given by,

$$SNR = 2E_y/N_0,$$  \hspace{1cm} (1.3)

where $E_y = \int_{-T/2}^{T/2} |p(t) * h(t)|^2 \, dt$ is the received signal energy. The problem is then reduced to find the optimum $p(t)$ such that $E_y$ is maximized, under the constraint of fixed $E_p$.

It has been shown in \(^{39}\) (p.125) and \(^{40}\) that the optimum $p(t)$ can be obtained by solving the following homogeneous Fredholm integral equation

$$\mu_n \phi_n(t) = \int_{-T/2}^{T/2} \kappa(t - \tau) \phi_n(\tau) \, d\tau,$$ \hspace{1cm} (1.4)

and let $p(t) = \phi_0(t)$, where $\phi_0(t)$ is the eigenfunctions corresponding to the maximum eigenvalue $\mu_0$ and the kernel $\kappa(t)$ is the autocorrelation of the CIR: $\kappa(t) = h(t) * h(-t)$. When convolved with the kernel over the interval $[-T/2, T/2]$, pulse waveform $\phi_0(t)$ reproduces itself, scaled by a constant $\mu_0$. With optimum $p(t) = \phi_0(t)$ we achieve the maximum SNR

$$SNR = 2\mu_0 E_p/N_0.$$ \hspace{1cm} (1.5)

It is worth noting that this maximum-SNR waveform may lead to severe ISI if the duration of $p(t) = \phi_0(t)$ exceeds the symbol duration, which is one reason preventing the scheme from being widely applicable.

Fig. 1.3. A conceptual transmitter structure capable of preprocessing (in the RF front-end frequency up- and down-conversion may be used to shift the signal frequency).

A transmitter capable of general waveform synthesizing or arbitrary waveform generating would be the ultimate goal of preprocessing. A digital FIR filter based waveform generator would be flexible and become more and more feasible as the semiconductor technology advances. A common structure of this type of waveform generators is a digital FIR filter followed by an analogue interpolating or shaping filter as illustrated in Fig.1.3. However, for any digital implementation, sampling rate as well as quantization resolution are limited. From a perspective of implementation, trade-off between performance and feasibility or cost must be made. Time reversal with mono-bit or ternary quantization and sub-Nyquist rate sampling has been proved working satisfactorily, according to computer simulation and test-bed based experiment.\(^{48,51}\) The practical limitations should be considered in the performance optimization.
1.4. Time Reversal Enabled Physical Layer Security and SDMA

In addition to a performance enhancement method, preprocessing can be an enabling technique as well. For example, a transmitter equipped with a time-reversal pre-filter bank followed by multiple antennas can not only focus signal temporally at the receiver, but also focus signal spatially at the intended location (the receiver). Fig.1.4 shows a MISO (multiple input antennas and single output antenna) preprocessing configuration. The temporal focusing is illustrated in Fig.1.5, assuming a MISO configuration with four transmit antennas and a practical waveform generator. The spatial focusing mechanism of a MISO time-reversal scheme can be explained in the following way. The pre-filtered signals travel from the transmit antennas to different locations, and at the intended location they are transformed to temporally condensed signals and their main lobes add up coherently; while at any location other than the intended one, the signals originating from different transmit antennas are widely-spread in time and have no similarity at all, thus together they appear to be a piece of noise-like waveform. Apparently, MIMO (multiple input antennas and multiple output antenna) will work too, where the multiple receive antennas can provide
diversity or data multiplexing. The spatial focusing property of multi-antenna time-reversal preprocessing can be used for spatial division multiple access (SDMA)\(^\text{46,51}\) or prevention of information leakage/interception.\(^\text{48}\)

Physical-layer security is extremely desired for security vulnerable wireless sensors networks. It has been proved that channel randomness and reciprocity can be used to establish secure keys against eavesdropping.\(^\text{52–54}\) Without using sophisticated procedure to achieve information-theoretic security, a simpler scheme to be proposed can add some level of security too. The time-reversal pre-filter’s impulse response (ideally a time reversed CIR) can be viewed as a location based security key,\(^\text{48}\) and this key can be obtained through reverse channel sounding thanks to channel reciprocity. The transmitted signal from a single antenna is temporally condensed at an intended location, while it is widely dispersed in time at other locations, which can be directly turned into location-based security enhancement at some circumstances.

![Graph](image)

Fig. 1.6. Demonstration of location-based security taking advantage of ISI.

To test capability of low probability of interception (LPI), suppose that there is a smart intercept receiver attempting to demodulate the signal sent to the intended receiver, assuming the intercepter employs the same type of receiver and knows all parameters (including data rate and timing information, etc.) used by the intended receiver, except its pre-filter’s setting. Simply consider a SISO (single input antenna and single output antenna) DTR-time-reversal scheme in a multipath environment with excess delay spread much greater than the symbol duration \(T_b\). Here a relatively simple FIR pre-filter is employed: it uses mono-bit coefficients and operates at the Nyquist rate. Its coefficients are generated by sampling a truncated CIR and then taking the signs from the samples. Location-based security enhancement in an office area is demonstrated in Fig.1.6 in terms of BER performance for a set of intercept locations, where the intended receiver integrates over the main lobe of width \(T_I\) while the intercept receiver integrates over the whole symbol duration \(T_b\).
$E_b/N_0$ is measured at the intended receiver, and both the intended and intercept receivers are about 6 m apart from the transmitter. Surprisingly, even just several tens of centimeters away from the intended receiver, although the interceptor has known all critical parameters, due to severe ISI, it has very little chance to understand what it is listening to, which means security has been greatly enhanced without using any traditional encryption method. More important, the “security key” is obtained at the physical layer based on the scattering nature, without consuming additional radio resource. In this example the intended receiver suffers from some ISI impact, but it can be reduced if some countermeasure is taken.

If equipped with multiple transmit antennas, the transmitted energy would truly focus at the intended location, achieving prevention of information leakage from an electromagnetic (EM) radiation point of view.

Another interesting feature of time reversal is its SDMA enabling. Note that the temporal/spatial focusing mechanism is not able to remove ISI and IUI completely. According to some indoor experiments, an SDMA multiuser system solely based on time reversal can suffer from strong IUI because of cross-correlations between the radio links, and even increasing the number of transmit antennas cannot fully solve the problem. For demonstration purpose, consider a rather simple time-reversal SDMA system using pulse based DTR scheme. If time reversal combined with pulsed signaling, a pulse train would appear at the intended receiver. Shrinking the integration window to capture the signal peaks at the receiver could result in a lowered SNR, but it reduces the IUI impact as well, since the signal to interference ratio around a signal peak is higher. Overall speaking, a narrow integration window may lead to a higher SINR (Signal to Interference plus Noise Ratio) and better performance, if the impact of IUI is much more significant than that of the background noise. For simplicity, assume the receiver samples the signal at peak instants, without integrating the signal over a main-lobe interval. To reduce the IUI impact further, a stage of decorrelation precoding (of course, other types of multi-user precoding can be considered too) is added to the transmitter to remove the IUI at the instants of signal peaks.

Shown in Fig.1.7 are a set of simulation results for multi-user schemes with different settings, where by default the pre-filter is significantly simplified (mono-bit time reversal), “ideal” refers to an ideal pre-filter, and “pre-decorr” refers to decorrelation precoding at the transmitter and sampling peaks at the receivers. The BER performance is mainly affected by the accuracy of pre-filtering, ISI, IUI as well as integration window size $T_I$. Two integration window sizes are considered here: $T_I = 6 \text{ ns}$ (default) and $T_I = 0.1 \text{ ns}$ (for peak capturing method). Based on these trials we have the following observations: (1) the mono-bit time-reversal scheme without precoding performs worst as expected; (2) the peak-sampling scheme yields improved performance; (3) decorrelation precoding combined with peak sampling improves the performance further and leads to more than 1.5 dB gain over the the mono-bit time-reversal scheme; and (4) in the ideal time-reversal scenario, applying decorrelation precoding plus peak sampling does not provide additional gain.

Time reversal combined with MISO is feasible, since the identical timing clock is shared by the multiple waveform generators. The system is especially promising when the number of antennas is large, say 16, 64, 128. The large number of antennas can be
integrated into one antenna system.

When time reversal combined with MIMO is used, the timing clock synchronization is hard to achieve at the receiver since the focused waveforms are too sharp to be aligned with each other.

Due to the same reason of timing clock, time reversal is hard to be used in the spatially distributed network. Some preliminary work has been reported. The spatial aspects of multiple antennas need more investigation in the framework of time reversal.

1.5. Channel Characteristics and Case Study

Well understanding channel characteristics in a particular environment is an essence to design a high performance communication system for the corresponding environment. In this section, different channel characteristics will be introduced and then four cases of channel analysis will be presented in detail.

1.5.1. Channel Characteristics

The channel characteristics are a set of basic information and distinguishing properties of the communication channel, including CIR, transfer function, channel energy, path loss, channel reciprocity, temporal focusing, and channel capacity, etc. The channel transfer function is simply a Fourier transform of the CIR, thus either of them can be derived from another by taking Fourier transform or inverse Fourier transform.

![Multi-user performance of a time-reversal-enabled SDMA system (\( T_b = 50 \text{ ns} \).)](image-url)
1.5.1.1. Time Domain Measurement of Channel Impulse Response (CIR)

The CIR is a time domain characteristic and it can be represented using the well-known tapped delay line (TDL) model. Time domain channel sounding is straightforward to get the CIR. In a time domain channel sounding experiment, the transmitter sends a pulse train through a transmit antenna to sound the channel and the waveform is picked up by a receive antenna and recorded as measurement data. One difficulty in time domain channel sounding is to generate sounding pulses that contain most signal energy in the frequency band of interest and are easy to be captured and processed. A digital sampling oscilloscope (DSO) is typically used to capture the waveform and store the data, where the sampling rate has to be sufficiently high to capture the details. The CIR in tap delay line model can be estimated from the received noisy waveform using “CLEAN”, a matching pursuit algorithm.

1.5.1.2. Frequency Domain Measurement of Channel Transfer Function

A channel transfer function gives the complex gain of the channel at different frequencies, thus frequency domain channel sounding is recommended to get channel transfer function directly. With a vector network analyzer (VNA), a frequency domain channel sounding setup would be quite easy. The VNA functions as both a transmitter and a receiver; it sweeps a set of sinusoid signals within the pre-determined frequency band and generates the S-parameters containing the complex transfer function. Frequency domain channel sounding can cover a wide range of frequencies, as far as the sweep time is much less than the channel coherent time. One disadvantage in a VNA based high frequency measurement is that the transmission range is limited due to large attenuation of the connecting cables. Because the frequency range is limited for a VNA, a rectangular spectral window is automatically posed. However, the spectral window may be reshaped to take into account any specific filter effect.

1.5.1.3. Channel Energy

The energy of the channel is defined as

\[ E_h = \int_0^{T_h} h^2(t) dt \]  

where \( h(t) \) is CIR and \( T_h \) is its support.

1.5.1.4. Path Loss

The received power is given by

\[ P_r = P_t - PL + G_t + G_r \]  

where \( P_r \) and \( P_t \) are the receive and transmit powers in dBm, respectively; \( G_r \) and \( G_t \) are the receive and transmit antenna gains in dBi, respectively; and \( PL \) is the path loss in dB.
The path loss between a pair of antennas is the reduction in power density of an electromagnetic wave as it propagates through space. It takes into account many physics effects including free-space path “loss”, refraction, diffraction, reflection and absorption. Path loss is environment dependent and it is a function of frequency and distance. Practically path loss can be calculated using formula

\[
PL(d, f) = PL(d_0, f) + 10 n(f) \log_{10} \left( \frac{d}{d_0} \right)
\]

(1.8)

where \( n(f) \) is the path loss exponent, \( f \) is the frequency of interest, \( d \) is the distance between the transmitter and the receiver, and \( d_0 \) is the reference distance. \( n(f) \) is equal to 2 for propagation in free-space. A path loss exponent greater than 2 means the propagation environment is relatively lossy, while a tunnel or a hallway act as a waveguide, resulting in a path loss exponent less than 2.

1.5.1.5. Channel Reciprocity

A real-time CIR is the key information to achieve the optimum performance in a wireless communication system, and channel sounding and estimation are required to generate the up-to-date CIR. Channel reciprocity refers to similarity of a pair of opposite-direction channels between two sites, where both the effect of propagation media and the effect of circuits at the two ends are taken into account. Perfect channel reciprocity can benefit a two-way wireless communication system, because channel sounding and estimation based on one link can be used for both forward link and reverse link. However, in reality, the channel reciprocity is not always obvious since there are many unpredictable nonlinear physics phenomena associated with the media and the circuits. This motivates experimental validation of channel reciprocity.

1.5.1.6. Relative Temporal Focusing as a Function of Locations

As mentioned before, pre-filtering (i.e., time reversal) can make the effective CIR temporally focusing and spatial focusing can be achieved using a bank of pre-filters followed by an array of antennas. Considering here a time reversal pre-filtering system with a single transmitter using a time reversed CIR as the pre-filter coefficients. Let the intended receiver location be represented by \( r_0 \) and the unintended receiver location by \( r_i \). Correspondingly, denote \( h(r_0, t) \) and \( h(r_i, t) \) the CIRs for the locations \( r_0 \) and \( r_i \), respectively. Relative temporal focusing at any location \( r_i \) with respect to the intended location \( r_0 \) can be characterized by a metric \( D(r_0, r_i) \) called directivity, which is defined as

\[
D(r_0, r_i) = \frac{\max |R_{hh}(r_0, r_i, t)|^2}{\max |R_{hh}(r_0, r_0, t)|^2}
\]

(1.9)

where \( R_{hh}(r_0, r_0, t) \) is a CIR autocorrelation for the intended receiver and \( R_{hh}(r_0, r_i, t) \) is a crosscorrelation of the two CIRs. \( R_{hh}(r_0, r_0, t) \) and \( R_{hh}(r_0, r_i, t) \) are given by

\[
R_{hh}(r_0, r_0, t) = h(r_0, -t) * h(r_0, t),
\]

(1.10)
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\[ R_{hh}(r_0, r_1, t) = h(r_0, -t) * h(r_1, t). \] (1.11)

1.5.1.7. Theoretical Spectral Efficiency

To evaluate the overall effects of the channel and different spectrum-shaping schemes, spectral efficiencies are calculated. The spectrum-shaping schemes to consider include water filling, time reversal, channel inverse and constant power spectrum density (PSD). A zero-mean band-limited Gaussian random data source which has maximum entropy is used for all different spectrum-shaping schemes. For the sake of simplicity, we further assume the data source has a flat in-band spectrum. When the water-filling spectral shape is applied, the spectral efficiency reached the maximum, i.e., the channel capacity. Channel capacity is an important channel characteristic and can be served as a performance bound to guide system design.

![Block diagram of a SISO system](image)

Fig. 1.8. Block diagram of a SISO system.

Fig.1.8 shows the block diagram of a SISO system, where the input signal \( a(t) \) is a band-limited white Gaussian random process with zero mean, bandwidth \( W \) and unit one-sided PSD, \( x(t) \) is the spectrum-shaping pre-filter, \( s(t) \) is the transmit signal, \( h(t) \) represents CIR, \( n(t) \) is the additive white Gaussian noise with one-sided PSD \( N_0 \), and \( r(t) \) represents the received signal. Their Fourier transforms are denoted by \( A(f) \), \( X(f) \), \( S(f) \), \( H(f) \), \( N(f) \) and \( R(f) \), respectively. Assume the support of \( A(f) \) is \([f_0, f_1]\), so that \( W = f_1 - f_0 \).

The transmitted power is

\[ P = \int_{f_0}^{f_1} R_S(f) df, \] (1.12)

where \( R_S(f) \) is the one-sided PSD of the transmitted signal \( S(t) \), and it is given by

\[ R_S(f) = |X(f)|^2. \] (1.13)

The noise power at the receiver is expressed as

\[ N = N_0 (f_1 - f_0). \] (1.14)

The SNR at the transmitter, denoted by \( \rho \), is

\[ \rho = \frac{P}{N}. \] (1.15)
Denoting $R$ the transmit bit rate, the spectral efficiency in bits/s/Hz can be expressed as

$$\frac{R}{W} = \int_{f_0}^{f_1} \log_2 \left( 1 + \frac{R_S(f) |H(f)|^2}{N_0} \right) df \quad (1.16)$$

When water filling is used to form the spectrum-shaping pre-filter, we have

$$R_S(f) = (\mu - \frac{N_0 |H(f)|^2}{|H(f)|^2})^+ \quad (1.17)$$

where $(x)^+ = \max[0, x]$, the constant $\mu$ is the water level chosen to satisfy the power constraint with equality

$$\int_{f_0}^{f_1} R_S(f) df = P, \quad (1.18)$$

and the spectral efficiency reaches the maximum, i.e., the channel capacity

$$\frac{C}{W} = \int_{f_0}^{f_1} \log_2 \left( \frac{\mu |H(f)|^2}{N_0} \right)^+ df. \quad (1.19)$$

When time reversal is used, it follows that

$$X(f) = \sqrt{\frac{P}{\int_{f_0}^{f_1} |H(f)|^2 df}} H^*(f), \quad (1.20)$$

where * means complex conjugate and the spectral efficiency is

$$\frac{R}{W} = \int_{f_0}^{f_1} \log_2 \left( 1 + \frac{\rho(f_1 - f_0) |H(f)|^4}{\int_{f_0}^{f_1} |H(f)|^2 df} \right) df. \quad (1.21)$$

For channel inverse is used, we have

$$X(f) = \sqrt{\frac{P}{\int_{f_0}^{f_1} |H(f)|^2 df}} \quad (1.22)$$

and

$$\frac{R}{W} = \log_2 \left( 1 + \frac{\rho(f_1 - f_0)}{\int_{f_0}^{f_1} |H(f)|^2 df} \right). \quad (1.23)$$

Finally, if the transmitted signal has the constant PSD from $f_0$ to $f_1$, then

$$R_S(f) = \frac{P}{f_1 - f_0}, \quad (1.24)$$

and

$$\frac{R}{W} = \int_{f_0}^{f_1} \log_2 \left( 1 + \rho |H(f)|^2 \right) df. \quad (1.25)$$
1.5.2. Case Study

Four different cases of channel analysis are presented. The first two cases are about the channel reciprocity in a car engine compartment and an office environment, respectively. The third case analyzes the channel characteristics in a rectangular metal cavity. Finally, in the case four, some measurement results conducted inside a manufacturing plant are provided.

Fig. 1.9. Display of a car engine compartment.

Fig. 1.10. Channel reciprocity validation in a car engine compartment.

1.5.2.1. Case 1: Channel Reciprocity in a Car Engine Compartment

A pair of omni-directional antennas were placed in the car engine compartment. The antenna locations are labeled in Fig.1.9. Fig.1.10 illustrates channel reciprocity between the
forward link A-G and the reverse link G-A, where link A-G corresponds to transmission from antenna A to antenna G, and link G-A corresponds to a reverse transmission. Two received waveforms measured in the forward and reverse links almost coincide with each other. The correlation between these two waveforms is as high as about 0.98, so that channel reciprocity can be declared from the engineering point of view. The bottom plot in Fig.1.10 shows a closer look of the upper plot. It was found that whether the engine was on or off, no difference could be observed from measurement. Moreover, similar results were obtained for different pairs of antenna locations.

Although time-domain measurement was used, the same conclusion can be expected for frequency-domain validation.

1.5.2.2. Case 2: Channel Reciprocity in an Office Environment

The time domain channel sounding was conducted in a lab/office area in Tennessee Technological University. It is a typical indoor area with wooden and metallic furnitures (chairs, desks, bookshelves and cabinets). Directional horn antennas were used. A skeleton of the measurement environment is shown in Fig.1.11 and antenna separation is about 6 m. No LOS was available during the measurement.

The antenna locations for the forward link are shown in Fig.1.11. Both received waveforms for the forward link and the reverse link are presented in Fig.1.12, and the two waveforms agree with each other very well. The correlation between these two waveforms is about 0.98. The bottom plot in Fig.1.12 shows a closer look of the waveform with the arrival time from 10 ns to 36 ns.

1.5.2.3. Case 3: Channel Characteristics in a Metal Cavity

Study of communications inside a metal cavity holds huge importance because the cavity can be viewed as an emulation of different confined metal environments like intra-ship, intra-vehicle, intra-engine, and so on. In this case, wide-band channel characteristics in such an environment are analyzed, which will help us know this type of channels better. Frequency domain channel sounding was performed inside a rectangular metal cavity by
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Fig. 1.12. Comparison of the received waveforms of the two links in an office environment.

Fig. 1.13. Setup for channel sounding in rectangular metal cavity.

using a VNA. The dimension of the cavity is 16 feet by 8 feet by 8 feet, and the surface material of the cavity is aluminum. The channel sounding setup is illustrated in Fig.1.13. Omni-directional antennas were used, with the transmit antenna being fixed and the receive antenna sliding along the horizontal central line in the cavity. The distance between transmitter antenna and receiver antenna varied from 0.5 m to 4 m in a step of 0.5 m. Table 1.5.2.3 lists the main parameters for this experiment. Measurement were also performed in an office and a hallway for comparison purpose.

Channel transfer functions and the corresponding passband CIRs for the three different environments are depicted in Fig.1.14. In all the cases, the distance between the transmitter antenna and the receiver antenna is about 4 m. It is observed that the delay spread of channel is about 800 ns in the cavity, while in the office and the hallway they are less than 200 ns. The in-cavity CIR with large delay spread consists of a large number of multipaths which does not appear in the office and the hallway.
TABLE 1.5.2.3 Measurement Parameters setup.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Band</td>
<td>3 GHz-10 GHz</td>
</tr>
<tr>
<td>Number of Points</td>
<td>7001</td>
</tr>
<tr>
<td>Transmission Power</td>
<td>10 dBm</td>
</tr>
<tr>
<td>Frequency Step</td>
<td>1 MHz</td>
</tr>
<tr>
<td>Antenna Polarization</td>
<td>vertical</td>
</tr>
<tr>
<td>Number of Averaging</td>
<td>128</td>
</tr>
<tr>
<td>Antenna Height</td>
<td>1.35 m</td>
</tr>
</tbody>
</table>

Channel energy comparison is shown in Fig.1.15. It is can be seen that the channel energy in the cavity is much higher than those in the office and the hallway. For example, when the distance between the antennas is 3 m, the channel energy in the cavity is nearly 20 dB higher than those in the other two environments. Moreover, the channel energy in the cavity is very stable as the antenna separation changes, but the channel energies in the office or the hallway drop apparently as the antenna separation increases from 0.5 m to 3 m.

For analyzing temporal focusing vs. locations, the transmit antenna was fixed on the horizontal central line, and the receive antenna was moved along a line that is 4 m away from the transmit antenna and perpendicular with the horizontal central line. As shown in Fig.1.16, 18 receive antenna locations with 3-cm separation were tested. In evaluation we consider the first location \( r_0 \) as the intended receiver and all others \( (r_i, i = 1, 2, \cdots, 17) \) as the unintended receivers. The received waveform autocorrelation \( R_{hh}(r_0, r_0, t) \) between the transmitter and the intended receiver is presented in Fig.1.17(a), and the cross-correlation \( R_{hh}(r_0, r_1, t) \) between the transmitter and the unintended receiver at location \( r_1 \) is presented in Fig.1.17(b).

The directivity in the cavity is calculated by using Equation 1.9 and shown in Fig.1.18. It is observed that the directivity drops by almost 20 dB when the unintended receiver is only 3 cm away from the intended receiver, which implies that communication security inside a metal cavity can be enhanced by taking advantage of the rich scattering property.

Finally, spectral efficiencies are calculated based on the measured data. Fig.1.19 shows spectral efficiencies in the cavity for different spectrum-shaping schemes. Among the four spectrum-shaping schemes, water filling, as expected, gives the maximum spectral efficiency (actually, the channel capacity), while channel inverse gives the minimum. At low transmit SNR time reversal performs better than constant PSD, but the spectral efficiency of constant PSD approaches that of water filling at high transmit SNR. Fig.1.20 shows spectral efficiencies of water filling (i.e., channel capacities) in the cavity, the office and the hallway when water filling is used. The channel capacity in the cavity is much higher than those in the other two environments, suggesting that confined metal environments have potential to support very high data rate transmission.
1.5.2.4. Case 4: Radio Channel Measurements in a Manufacturing Plant

An industry environment like a in-building manufacturing area is a very complex environment for the electromagnetic wave propagation. Due to its complexity, it is hard to evaluate the channel using some prediction models which require enormous number of geometric information about the environment. It seems that measurement is the only way to study...
channel characteristics in such type of environments. In this case study, path loss exponent $n$ is calculated based on the measurement data collected inside a manufacturing plant. As for transmit-receive antenna pairs, several combinations, including Omni-Omni, Omni-Horn and Horn-Horn were tested. An omni-directional antenna has wide coverage angle at the penalty of short transmission range, while a horn antenna can transmit far but has a narrow coverage angle. The mentioned three antenna combinations can fit different applications and requirements. Path loss exponents for different measurement areas are
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(a) Waveform autocorrelation of the intended receiver. (b) Waveform crosscorrelation between the intended receiver and the unintended receiver at \( r_1 \).

Fig. 1.17. Received waveform correlations for different receiver locations.

Fig. 1.18. Directivity in the rectangular metal cavity.

calculated and listed in Table 1.5.2.3. Fig.1.21 shows the received waveform captured directly by a DSO when antennas were placed along a production line. The impression is that the signal envelop decays slowly, and the delay spread is in the order of microsecond.

1.6. A Reference Example of Time Reversal Radio Test-Bed

The work of radio system design is to provide an efficient solution under some theoretical and practical criteria. A top-down design flow covers many aspects ranging from a very high level design to detailed implementations. A number of issues need to be considered: frequency band, architecture, data rate, modulation, synchronization, coexistence, interfer-
Fig. 1.19. Spectral efficiencies of different pre-filtering in the cavity.

Fig. 1.20. Spectral efficiencies of water filling (i.e., channel capacities) in different environments.

Fig. 1.21. The received waveform when the antennas are along the product line.

ence, dynamic range, and many implementation issues. As for the system with preprocessing for wideband applications like UWB, one of the major challenges is in the waveform generator. Specifically, it is about how to choose a proper sampling rate as well as quantization resolution, and how to efficiently implement the algorithm. Based on our indoor experiments in UWB band, a reduced-complexity pre-filter can function very well and implementing time-reversal preprocessing for microwave-band communications is feasible. A reference example of time reversal radio test-bed is presented in the following.

1.6.1. Design Considerations

We adopt pulse based signaling and transmitter-side processing as system design guideline. Although direct pulse (carrier less) transmission can largely reduce complexity of transmitter RF front-end, it is not a good choice for multi-purpose radio testbed mainly because of its inflexibility. As a matter of fact, a modulated pulse is not only easy to generate but also
more flexible: the center frequency is controlled by a local oscillator and the spectral shape is governed by the baseband pulse. Conceptual testbed architecture is shown in Fig. 1.22, where all baseband and control functions are implemented using FPGAs. Following an FIR filter (implemented in the FPGA), the digital-to-analog converter (DAC) outputs desired analog waveforms. The simple receiver philosophy is reflected in this testbed with on/off keying (OOK) or binary pulse position modulation (2-PPM) and diode based non-coherent detector at the receiver. At the receiver, demodulation is done in digital domain, so that algorithms and parameters can be adjusted easily. Of course, the analog-to-digital converter (ADC) is power hungry and it may be replaced by some customized circuits or commercially available products in the future.

1.6.1.1. **Link Budget Estimation**

The link budget estimation shown in Table 1.6.1 is based on the following conditions: (1) free space propagation, and (2) OOK modulation with equi-probable transmission of “0” and “1” bits. The transmitted power used here is estimated considering the actual components used in the transmitter. Spectrum measurement shows that the estimated transmitted PSD is closed to the measured result. The actual transmission range depends on the propagation environments. Path loss is much larger in a none-line-of-sight (NLOS) environment.
TABLE 1.6.1 A Link Budget Example.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Bit rate</td>
<td>6.25 Mb/s</td>
</tr>
<tr>
<td>Minimum required SNR per bit $E_b/N_0,\text{min}$</td>
<td>16 dB</td>
</tr>
<tr>
<td>Center frequency</td>
<td>4 GHz</td>
</tr>
<tr>
<td>Transmitter antenna gain</td>
<td>3.5 dBi</td>
</tr>
<tr>
<td>Receiver antenna gain</td>
<td>3.5 dBi</td>
</tr>
<tr>
<td>Distance between the two antennas</td>
<td>15 meters</td>
</tr>
<tr>
<td>Receiver noise figure</td>
<td>6 dB</td>
</tr>
<tr>
<td>Implementation loss</td>
<td>4 dB</td>
</tr>
<tr>
<td>Transmitted power (input to antenna)</td>
<td>-16.45 dBm (0.0227 mW)</td>
</tr>
<tr>
<td>Bandwidth (approximately)</td>
<td>800 MHz</td>
</tr>
<tr>
<td>EIRP spectral density</td>
<td>-41.98 dBm/MHz</td>
</tr>
<tr>
<td>Path loss at 1 m</td>
<td>44.48 dB</td>
</tr>
<tr>
<td>Path loss at 3 m</td>
<td>54.03 dB</td>
</tr>
<tr>
<td>Total path loss</td>
<td>68.00 dB</td>
</tr>
<tr>
<td>Received power</td>
<td>-77.45 dBm</td>
</tr>
<tr>
<td>Link margin</td>
<td>2.59 dB</td>
</tr>
<tr>
<td>Proposed minimal receiver sensitivity</td>
<td>-80.04 dBm</td>
</tr>
</tbody>
</table>

and the large delay spread would introduce ISI. Extra path loss and/or ISI can shorten transmission range significantly.

1.6.1.2. Calibration Phase

This testbed has programmable transmit waveform function. To generate desired waveforms, the waveform generator at the transmitter has to be calibrated prior to regular data transmission. In addition, the receiver needs to set a proper amplification gain and decisions thresholds priorly. Tasks in the calibration phase include channel sounding, channel estimation, FIR filter coefficient optimization, determining the amplification gain, and choosing the thresholds. In reality, the channel information can be obtained at either the transmitter or the receiver, but in either cases a feedback channel and some handshaking procedure are necessary. Considering limited sampling rate and resolution, there exist a best set of filter coefficients under some criterion. There have been intensive work on channel sounding and estimation, and many proposed methods can be considered. we sound the channel using lab instruments, estimate the channel and optimize the coefficients in an off-line manner. The amplification gain and thresholds are set manually.

1.6.1.3. Energy Collection

A direct consequence of a high-bandwidth UWB signal is ultra fine multipath delay resolution in multipath propagation environments. Theoretically, to efficiently capture the signal energy dispersed over a large number of individual paths, either a RAKE receiver scheme
or an OFDM scheme can provide high performance, given perfect synchronization and channel estimation. Realistically, a RAKE receiver with tens of fingers is infeasible, and both schemes mentioned above are financially improper for low-cost low-data-rate applications. There is a huge potential market for these lower-end applications, such as sensor networks. In response to this need, several suboptimal receiver schemes, including TR and energy detection using a square law detector, have regained popularity in the UWB community. Although both TR and energy detection suffer from performance penalty, they have no need for sophisticated channel estimation and precise synchronization, which significantly reduces receiver complexity and cost. OOK modulation and energy detection is indeed a reasonable combination. Received signal energy can be captured easily using a square law or diode detector followed by an integrator, and OOK works fine if the data symbol boundary is roughly known and inter-symbol-interference (ISI) is negligible. Pulse position modulation (PPM) is another popular modulation for pulse based wide band systems, and high order PPM or called M-ary PPM is promising to work with channel coding to achieve wide range of scalability.

Fig. 1.23. Conceptual diagram of energy detector based receiver.

A conceptual diagram of energy detector based receiver is shown in Fig.1.23, where an integrator following the detector accumulates the signal energy. The integration window size is a key parameter that affects performance. Since the integrator captures both the signal energy and the harmful noise energy, a smart way of energy capture is to weight the incoming signal before integration, which is at the cost of higher complexity. A simpler way is to integrate the incoming signal over a proper integration window. Instead of using fancy integrators, we rely on the waveform generator at the transmitter to focus signal energy and set the integration window to a fixed size that is much smaller than a typical delay spread in a NLOS environment.

1.6.1.4. Synchronization and State Transition

We have designed a 3-stage synchronization method that is of lower complexity than traditional direct correlation based methods. Here are functions of the three stages: stage 1 for signal-arrival detection (or coarse synchronization), stage 2 for chip synchronization, and stage 3 for frame synchronization. The main advantages of this 3-stage approach include: (1) the stage 1 provides coarse timing quickly and reduces search space; and (2)
hard decision is made after stage 2, so that the stage 3 deals with mono-bit sequence, resulting in less computation and lower complexity. To shorten the overall synchronization time further, parallel chip level processing is employed at the cost of more FPGA resources occupation.

The receiver has three working state and the state transition diagram is given in Fig. 1.24. Starting with the idle state, if the receiver detects “Signal arrival”, the system goes to the fine synchronization state; if synchronization succeeds, the system enters into the data transmission state, demodulates the data and goes back to the original idle state after demodulation. On the other hand, if synchronization failed, the system returns back to the idle state. These state transitions are executed by the finite state machine (FSM) implemented in FPGA.

1.6.2. Implementation

1.6.2.1. Transmitter

Design of Waveform Generator

A symbol waveform can include multiple pulses to reduce peak power while keeping enough symbol energy (in this design a symbol contains one data bit). Scrambling coding is necessary to remove DC component and reduce the spectral spikes in the modulated signal. In typical indoor environments, delay spread is in the order of tens of nanoseconds, and inter-pulse-interference (IPI) is inevitable as the pulse repetition rate reaches the level of tens of MHz. A waveform generator is placed at the transmitter to focus the received signal in time and reduce the ISI impact. OOK and PPM are two modulation schemes that are suitable for non-coherent detection. For higher data rates, a 2-PPM symbol duration has to be double of an OOK symbol duration to avoid IPI. However, finding a proper threshold for OOK demodulation at unknown level of background noise is challenging, but threshold selection is not necessary for 2-PPM scheme if using differential energy detection.
Time reversal waveforms have been used to test the system and verify the performance. In a time reversal system, the transmit chip waveform is the time-reversed version of the CIR, which is environment dependent. The waveform generator in the transmitter has to be programmable to match the instant CIR. Studies have shown that the received signal concentrates a large percentage of the total energy in a short period of time, enabling the use of some simple receiver that have little tolerance to IPI and ISI.

In the transmitter, the baseband waveform generator consists of a pre-filter module implemented in Xilinx Virtex-5 FPGA, a DAC supporting Gsps sampling rate, and the corresponding connection buses. The pre-filter is a pair of FIR filters with programmable coefficients able to quadrature sequences.

On the market there are not many DAC and ADC products supporting Gsps sampling rates. Top concerns in selecting them include multiplexing or demultiplexing features, reconfigurability and dual-channel. The DAC chip used in the testbed supports dual-channel, 14-bit resolution, and sampling clock rate up to 1.0 Gsps. It also features 2:1 (or 28:14) multiplexing, waveform memories, low voltage differential signal (LVDS) interface and configurability. Although the high speed digital interface standard LVDS has been used in the transmitter, the connection between the FPGA and the DAC is actually a bottleneck that limits the speed. This fact suggests that to increase the speed further, higher ratio multiplexing, shorter connection and maybe some other efforts have to be considered.

Transmitter RF Front-End

We mainly rely on off-the-shelf products for RF front-ends in the transmitter. The transmitter RF front-ends is illustrated in Fig.1.25. The baseband signal from the DAC is up-converted to a bandpass signal by the modulator, and after passing through the the power amplifier, finally the amplified bandpass signal radiates from the antenna.

Reconfigurability and wide frequency range are preferred for the local oscillator (LO). A quadrature type of modulator (or up-converter) is needed to handle complex signals. Baseband bandwidth and input LO frequency range of a modulator (or up-converter) are...
particularly important to generate wide-band signals. The power amplifier has to be of wide bandwidth and there are many choices. A variable attenuator may be used with the power amplifier in order not to break the FCC emission power ruling.

One common issue in RF implementation is carrier leakage rejection. Although the modulator has very good carrier isolation (say, -37 dBm), there are some other paths that the carrier can pass to the output port. Also, unbalanced bias between the two differential input ends of a modulator (or up-converter) makes carrier contribution at the output. Conventional RF decoupling techniques can reduce carrier leakage. In addition, a narrow band notch filter at the carrier frequency may be employed to suppress the output carrier frequency.

![System control flow for the transmitter.](image)

**System Control**

All system control tasks are taken cared by the FPGAs. Configuring some devices such as DAC and local oscillator (LO) is part of system control. The serial peripheral interface (SPI) bus is a simple and flexible interface standard that allows one master device to communicate with one or multiple slave devices, and it can support data rates up to tens of Mbps. In the transmitter, the LO is connected to the FPGA via an SPI bus with the FPGA serving as the SPI master and the LO as the SPI slave. It shows that SPI bus
protocol is efficient and can increase system’s flexibility. The system control flow chart in the transmitter side is presented in Fig. 1.26.

1.6.2.2. Receiver

Front-End Amplification

Requirements for receiver front-end amplification include low noise figure (< 4 dB), enough gain (max. gain > 60 dB) and proper dynamic range (> 30 dB), considering medium range indoor communications. A combined amplifier is built by concatenating a low noise amplifier (LNA) module with ultra low noise figure, a variable attenuator and a gain block.

ADC and Its Interface to FPGA

An 8-bit monolithic bipolar ADC with sampling rate up to 1.5 Gbps has been selected. The variable sampling rate is achieved by controlling the output frequency of the clock source. The ADC features an on-chip, selectable 8:16 output demultiplexer. A double-data-rate (DDR) interface implemented in FPGA connects the ADC to the FPGA. Although the maximal resolution is 8 bits, lower resolution can be chosen in signal processing.

In the receiver of the radio testbed, the ADC employs a fully differential 8 bits quantizer and a unique encoding scheme to limit metastable states, with no error exceeding 1 LSB max. To facilitate a lower rate digital interface, the ADC features an on-chip, selectable 8:16 positive-referenced emitter-coupled logic (PECL) compatible output demultiplexer that reduces the output data rate to one-half the sampling clock rate.

An ADC-FPGA interface has been carefully designed to support a sampling rate of 800 Msps in the receiver. A single-data-rate (SDR) interface with a FIFO memory is implemented in FPGA to receive data from the ADC.

In designing the high-speed interface between ADC (or DAC) and FPGA, signal integrity has become a critical issue. Many signal integrity problems are electromagnetic phenomena in nature. There are two concerns for signal integrity: the timing and the quality of the signal. Signal timing mainly depends on the delay caused by the physical length and the material of each segment along a signal propagating path. Signal waveform distortions can be caused by reflection, cross talk, and power/ground noise.

An important parameter is the characteristic impedance \( Z_0 \) defined as the voltage-to-current ratio at any point along the transmission line. If the far end is terminated with resistance \( R = Z_0 \) over the entire frequency band, impedance matching is perfect and the forward propagating wave is fully absorbed by the load. On the other hand, an impedance mismatched load would cause reflecting wave traveling back to the source, and it may be reflected again if the source is impedance mismatched, resulting in rings on top of the original waveform. Such ringing has a serious impact on signal integrity, reduces noise/interference toleration margins, and even leads to malfunction. A way to avoid reflections and ensure signal integrity is a proper termination. Most high-speed data interface standards (such as LVDS) require load-end termination for impedance matching but do not require source/driver-end termination. For better result, double-end termination at both ends of the
TABLE 1.6.2.2 Device utilization summary.

<table>
<thead>
<tr>
<th></th>
<th>Transmitter</th>
<th>Receiver</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of slice flip flops</td>
<td>495</td>
<td>1533</td>
</tr>
<tr>
<td>Number of 4 input LUTs</td>
<td>573</td>
<td>2250</td>
</tr>
<tr>
<td>Number of occupied slices</td>
<td>331</td>
<td>1813</td>
</tr>
<tr>
<td>Total number of 4 input LUTs</td>
<td>642</td>
<td>2635</td>
</tr>
<tr>
<td>Number of bonded IOBs</td>
<td>48</td>
<td>52</td>
</tr>
<tr>
<td>Number of DCMs</td>
<td>0</td>
<td>2</td>
</tr>
<tr>
<td>Number of PLL/ADVs</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>Number of BlockRAM/FIFO</td>
<td>1</td>
<td>0</td>
</tr>
<tr>
<td>Total equivalent gate count for the design</td>
<td>146795</td>
<td>376714</td>
</tr>
</tbody>
</table>

connection bus has been adopted. It can be observed in Fig.1.27 that the signal swing is reduced significantly with the presence of the proposed termination.

![Data waveform images](image)

(a) Data waveform with load-end termination only. (b) Data waveform with double-end termination.

Fig. 1.27. Impact of termination on high-speed data transmission.

FPGA Based Receiver Processing

The receiver Verilog coding is more complicated than the transmitter’s because much more digital signal processing algorithms need to be implemented in the receiver. In addition, parallel chip level processing is employed for fast acquisition, at the cost of more FPGA resources occupation.

The routed FPGA design shown in Fig.1.28 is observed using the software Xilinx FPGA editor. The blue/light area indicates the occupied logic resources. The target FPGA device in the receiver is one of the Virtex II platform. The detailed logic utilization for the transmitter and receiver is shown in Table 1.6.2.2. It can be seen that more resources are utilized in the receiver.
1.6.3. System Test and Results

The functionalities of the time reversal testbed and the temporal focusing property can be verified through experiment. A packet with a given data payload is transmitted repeatedly. The transmit and receive antennas are placed in a non-line-of-sight (NLOS) manner. At the receiver, if the demodulated sequence matches to the transmitted sequence, then we can say the system is working properly. Moreover, waveform level temporal focusing can be observed at the receiver using an oscilloscope.

1.6.3.1. Preparing the Experiment

In addition to the testbed, below is a list of instruments used in the experiment, and each of them must be calibrated well before use.

- Vector Network Analyzer (VNA)
- Spectrum Analyzer (SA)
- Logic Analyzer (LA)
- Digital Phosphor Oscilloscope (DPO)

Time reversal template (coefficients of the FIR pre-filter) preparation is a major task in the experiment. Specifically, the CIR has to be estimated and converted into a fixed-point sequence, and then load the sequence to the coefficient memory. In a real system, we need to implement channel sounding and estimation functions to provide the transmitter with the
1.6.3.2. Environment and System Setup

According to time reversal theory, the environment should be as complex as possible, so that rich multipath can be obtained. According to this philosophy, our system test is performed in a typical non-line-of-sight (NLOS), three-compartment office environment (Fig. 1.29). Transmit antenna and receive antenna are located in compartment A and compartment B, as can be seen from Fig. 1.29. There are wood and metal shelves, desks and chairs, computers and electronic equipments in the office. These furnitures reflect signals and cause multipath propagation. These antennas are omni-directional and the line-of-sight template. However, for the sake of time reversal functionality verification, with help of lab instruments, we can obtain the template off line and load it to the memory in advance.
between is blocked by the wall and furnitures. This setup ensures plenty of multipaths.

Note that the transmitted signal is a passband signal while the time reversal template is a baseband signal. There are different ways to prepare the template and the frequency domain channel sounding method has been adopted. The passband channel transfer function is measured using a VNA. As shown in Fig. 1.30, the frequency sweeps from 3.5 GHz to 4.5 GHz with 1 MHz resolution, covering the frequency band of the transmitted signal. In reality the measured passband channel transfer function is always associated with a known spectral window. If \( h(t) \) is the CIR for the frequency range of interest, the equivalent baseband CIR would be \( h'(t) = h(t) e^{-j\omega_c t} \), where \( \omega_c \) is the center frequency. To estimate \( h'(t) \), the measured passband channel transfer function is down-converted, then apply inverse Fourier transform (IFT) to the baseband channel transfer function to obtain a filtered baseband CIR, and finally the baseband CIR \( h'(t) \) can be estimated using some deconvolution algorithm like CLEAN to remove the filter effect. The filter effect is due to the spectral window and the filter is an IFT of the known spectral window. Depicted in Fig. 1.31 is the magnitude of the filtered baseband CIR based on measurement.

\( h'(t) \) would be an ideal baseband time reversal template, but it cannot be accurately generated due to the practical limitations. What have been tested is a ternary-quantized real-number baseband template with sampling rate 500 Msps. This template was optimized under maximum correlation criterion. Fig. 1.32 is a piece of baseband signal corresponding to four scrambled chips, captured at the waveform generator’s output by a Tektronix TDS 7104 DPO. Since the approximate baseband time reversal template contains four 2-ns pulses, each chip comprises four 2-ns pulses. Fig. 1.33 is the spectrum of the modulated waveform captured at the transmitter by a Rohde & Schwarz Spectrum Analyzer FSEM20. The center frequency is about 4 GHz and the 10 dB bandwidth is about 800 MHz.

At the receiver side the DPO is used to monitor the detector’s output as well as the demodulated bit stream, and one could tell if the system is running properly simply by
1.6.3.3. Experimental Results

During the first demonstration, a given pseudo-random (PN) sequence was transmitted periodically at bit rate 6.25 Mbps. Fig.1.34 shows the system test observations. The top trace is the transmitted baseband waveform, the middle trace is the detector output, and the bottom trace is the demodulated bit stream. In order to observe the signals more conveniently, the scales for top trace, middle trace and bottom trace are set to 50 mV/div, 100 mV/div and 5 V/div, respectively. The top trace is simply a zoom-out version of Fig.1.32. Although there are some interferences observed on both the middle and the bottom trace, distinguishable peaks can be seen on the middle trace and the PN sequence can be read clearly from the bottom trace.

A close-up of the middle trace is shown in Fig.1.35. The top trace is the baseband trans-
mitted waveform, the middle trace is the received waveform at the detector’s output, and the bottom trace is a data clock running at 6.25 MHz, the same as the data rate. Temporal focusing property can be observed from the middle trace. Four pulses are transmitted in each chip period, while in the detector’s output there is only one strong peak in each chip period, because the four separated pulses arrived at the receiver at the same time after traveling through different paths and add up constructively. Although the mainlobe is dominant, some sidelobes are noticeable, suggesting that either the waveform generator needs to be refined, or other signal processing techniques should be combined.

1.7. Conclusions

High-resolution of ultr-wideband pulses leads to rich multipath. Fading is secondary for UWB systems. The channel is reciprocal for a long period of time—much longer than its narrow-band counterpart. The central idea presented in this chapter is to take advantage of this reciprocity to focus multipath pulses at the receiver—through waveform preprocessing at the transmitter. This scheme allows to use simple receivers and enables spatial division multiplexing access (SDMA) and additional physical layer security.

The principles and design philosophy, along with an example of experimental time reversal test-bed, have been introduced. Waveform level preprocessing with fractional symbol sampling is a general pre-filtering process at the transmitter for different optimum criteria. Some traditional receiver based processing functions can be taken care of by preprocessing, so that suboptimal reception schemes such as transmitter-reference and square law detector can be used without significantly sacrificing overall performance. Preprocessing paired with simple receivers are especially attractive to wireless sensor network (WSN) applications where strict constraints are posed. In addition to performance enhancement, preprocessing also features SDMA or physical layer security enhancement without consuming extra radio resources. Knowing the characteristics of scattering channels is the key to best design a preprocessing system, which motivates our experiment based investigation on a few realistic radio channels. The concept of preprocessing plus simple receivers has been demonstrated using a time reversal radio test-bed working in microwave band in the indoor environment.

For future work, security and anti-jamming capabilities are enhanced by channel reciprocity. This line of thought needs further research. The availability of five GHz digital-to-analog converter (DAC) makes this direction feasible. Compressed sensing provides a revolutionary approach to handle waveforms of much wider bandwidth. Transient UWB pulses are compressible. Information theory can be used for secrecy. Ultra-wideband sensing and networking can be integrated into one radio. This radio must have the ability to access spectrum dynamically—cognitive radio and cognitive radar. As a result, wideband (e.g. multi-GHz) spectrum sensing is needed—but an open problem.
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