A FREQUENCY DOMAIN APPROACH FOR
TIME-REVERSAL OF MICROWAVE
IMPULSES

by

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ABSTRACT

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In this thesis, a compact and low-cost electronic circuit system is designed for time-reversal of microwave impulses with nanosecond and sub-nanosecond temporal durations. A frequency domain approach is adopted in order to avoid high sampling rate in time. The proposed system obtains the discrete spectra of input impulses first, then realizes time-reversal in frequency domain, and finally synthesizes the time-reversed impulses using discrete continuous wave elements. It is composed of commercially available circuits including oscillators, mixers/multipliers, band-pass-filters, amplifiers, and switches, hence embodies low-cost system-on-chip implementation. The proposed time-reversal circuit’s performance is verified by Advanced Design System (ADS) simulations, with most non-idealities of realistic circuit components taken into account. Simulation results show that, microwave impulses with about 1 ns temporal width and 3 – 10 GHz spectral coverage are reliably reversed in time, even with presence of strong noise.

Furthermore, the proposed time-reversal circuit system is validated in the context of electromagnetic propagation in complex environments. Specifically, circuit-electromagnetic co-
simulation is carried out to investigate the “focusing” phenomena of time-reversal. A full-wave Maxwell’s equations solver based on Finite Difference Time Domain (FDTD) method is developed to model electromagnetic propagation, and it is coupled to ADS circuit simulator. The FDTD solver is implemented on parallel cluster Message Passing Interface (MPI), in order to relieve high computational complexity due to complex environments. Two real-world problems (one is for wireless communication and the other is for radar detection) are investigated. Desired “focusing” phenomena in both space and time are demonstrated by the simulation results, which conclude that the proposed time-reversal system can be deployed in practical time-reversal communication and radar applications.
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CHAPTER 1
INTRODUCTION

In this chapter, time-reversal is briefly reviewed in Subsection 1.1; then, Subsection 1.2 discusses about technical difficulties involved in time reversing high-frequency signals; finally in Subsection 1.3, our approach to resolve the time-reversal difficulties is laid out.

1.1 Review of Time-Reversal

Time-reversal technique has attracted enormous research interests due to its unique capability to focus waves in space and time. Extensive investigations have been devoted to time-reversal in acoustics [1-4], for underwater communication [3, 5] and target detection/imaging [6-8], for instances. The studies of time-reversal in electromagnetics also have a long history; especially in recent years, a lot of research work on time-reversal for electromagnetic waves has been reported [9-19]. It is largely due to the advance of impulse-type ultra-wide-band (UWB) technology, which offers many advantages such as high data rate, more immunity to multi-path fading, low power consumption, low cost, and low complexity [20-22]. It is shown by previous research that, combination of UWB and time-reversal has potential to realize various novel communication and radar systems [23-30]. The applications include but are not limited to MIMO communication [31], indoor communication [32, 33], communications in forest environments [34], buried target detection [19], ground penetrating radar [17, 35], and breast cancer diagnosis [36, 37].

The time-reversal process usually consists of three steps, as illustrated in Figure 1.1: first, the signals emitted by a source are recorded by a time-reversal cavity (TRC) or time-reversal mirror (TRM); second, the recorded signals are time-reversed in either time domain or frequency domain (a complex conjugate in frequency domain is equivalent to time-reversal in time domain); third, the time-reversed signals are reradiated by TRC or TRM. If the medium is
time-invariant and reciprocal, the time-reversed signals will add up constructively at the original source location, in both space and time [38-40]. In other words, time-reversal behaves as a space-time matched filter and hence is able to compensate the channel as long as reciprocity holds true [41, 42].

Figure 1.1 Time-reversal process in a homogeneous medium. (As an illustration, the time-reversal mirror (TRM) is made of four-element linear antenna array.)

When the time-reversal process is carried out in a homogeneous medium and there is no multipath effect, size of the spatial focusing spot is limited by half-wavelength (usually termed as diffraction limit). The range and cross-range resolution is dependent on the angular aperture size of the TRM (Figure 1.2) [6]. However, when the medium is inhomogeneous or there is multipath reflection (such as in a waveguide or a closed cavity), the focusing spot is smaller than in a homogeneous medium. This enhanced resolution is called “super-resolution”. The more multipath effect, the stronger the resolution enhancement is. An explanation of the super-resolution phenomenon is that the effective aperture of the TRM is larger than its actual aperture size, due to the multipathing (Figure 1.3) [6]. However, this super-resolution is still diffraction limited.
Figure 1.2 Time-reversal in a homogeneous medium. The TRM has an aperture size of $a$, and distance of $L$ from the point source at $y$. (a) Forward propagation in homogeneous medium; (b) Back-propagation and focusing in homogeneous medium. The range resolution is proportional to $(L/a)^2$, and the cross range resolution is proportional to $(L/a)$.

Figure 1.3 Time-reversal in an inhomogeneous medium with strong multipath reflection. (a) Forward propagation in inhomogeneous medium; (b) Back-propagation and focusing in inhomogeneous medium. The TRM appears to have an effective aperture $a_e > a$, and the focusing spot is more compact than that in homogeneous medium.

The time-reversal process described above only reverses the boundary condition (the field sampled by TRM), but not the source. The ideal time-reversal needs to reverse the source too, which is equal to implement a sink that absorbs the back-propagating wave at the source location. This sink has been realized in acoustics [43], and resolution less than $\lambda / 14$ is achieved.

Another approach to overcome the diffraction limit is to time-reverse the evanescent wave, which contains subwavelength information about the source, but decays exponentially. In
[39], the author placed subwavelength scatterers in the near-field of the source. By diffracting off the scatterers, evanescent waves can convert into propagating waves, and are captured by TRM in the far-field region. When the time-reversed fields are transmitted back, they are converted back into evanescent waves originating from the source. A super-resolution of $\lambda/30$ is obtained in the paper.

### 1.2 Implementation Difficulty of Time-Reversal in Electromagnetics

In the second step of TR process, the signals can be reversed in both in time domain and frequency domain.

In time domain, the reversal process involves sampling, storing, and reversing. There exist synchronization issues when there are multiple elements in TRM. In acoustics, this is not difficult because the frequency is very low. However, electromagnetic waves have high operating frequencies (for instance, 3.1 – 10.6 GHz band was allocated for UWB communication applications). As a result, much faster sampling is required to directly sample the impulses and then reverse them. To experimentally investigate time-reversal of electromagnetic waves, some researchers had to resort to expensive high-speed analog-to-digital converters [44-51]. As an example, Tektronix TDS6604 digital storage oscilloscope with 20 G samples per second was used in [45].

Time-reversal in frequency domain is equivalent to complex conjugate of signal spectrum. For monochromatic wavefields, phase-conjugate mirrors (PCM) can be used [1, 13]. In some research efforts, time-reversal was realized using vector network analyzers, where data were collected in frequency domain initially and synthesized afterwards [52-57]. Microwave signal with 18 GHz bandwidth was time reversed in [58] with aids of up-conversion to optical carrier and non-linear optical complex conjugate.

Apparently, to accomplish practical time-reversal communication and radar systems, efficient and low-cost techniques to reverse UWB impulses are called for.
1.3 Our Approach to Achieve Time-Reversal in Electromagnetics

In this thesis, an electronic circuit system is designed to time reverse UWB impulses (with nanosecond and sub-nanosecond temporal durations). A frequency domain approach is adopted to avoid high sampling rate in time domain. Specifically, time-reversal is achieved in the following three stages.

(i) Fourier transform, to obtain discrete spectra of input impulses.

(ii) Digital signal processing in frequency domain, to carry out complex conjugate of the spectra obtained in (i).

(iii) Inverse Fourier transform, to synthesize reversed impulses based on the processed spectra in (ii).

It is worth noting that, our group is not the first one who attempts applying frequency domain approaches to UWB signal generation and reception. Sub-band analysis [59, 60] and bank of passive resonators [61] were investigated before to relieve the high sampling rate problem in UWB receivers; and, Fourier synthesis was suggested for UWB signal generators as early as in 1997 [62]. However, to our best knowledge, it is the first time to propose a hardware architecture which combines Fourier transform and inverse Fourier transform circuits to accomplish time-reversal of UWB signals. The proposed system is composed of common semiconductor electronic circuits, including oscillators, multipliers/mixers, band-pass-filters, amplifiers, and switches. Thus, it is compact, portable, and low-cost; moreover, it can embody a system-on-chip implementation. Basic concept of the system proposed was presented in [63]; here, complete description and verification are presented. The circuit system is simulated by Advanced Design System (ADS) with commercially available components; most of the non-idealities of realistic circuits are taken into account. Simulation results using impulses with around 1 ns temporal width and [3, 10] GHz spectral coverage demonstrate real-time and reliable time-reversal, even with presence of strong noise.
The proposed time-reversal system is further verified in the context of electromagnetic propagation. Two real-world problems (one is for wireless communication and the other is for radar detection) are analyzed using circuit-electromagnetic co-simulations. A full-wave solver, finite difference time domain (FDTD) [64], is used to solve the Maxwell’s equations. It is implemented on parallel cluster using Message Passing Interface (MPI) to relieve high computational cost. Electromagnetic propagation results with ideal time-reversal and practical time-reversal (i.e., time-reversal using circuits in this paper) are compared. In both examples, desired “focusing” phenomena are observed. It is therefore concluded that, errors introduced by the time-reversal circuits are tolerable and the proposed system can be deployed in practical communication and radar applications.

This thesis is organized as follows. Chapter 1 is the introduction of time-reversal and brief description about the work in the thesis. Chapter 2 describes the frequency domain approach of time-reversal circuit system in details. In Chapter 3, the algorithm of parallel FDTD is explained. Results of circuit simulation and circuit-electromagnetic co-simulations are presented in Chapter 4. Finally, Chapter 5 summarizes the conclusions and future work.
CHAPTER 2

DESCRIPTION OF THE PROPOSED TIME-REVERSAL SYSTEM

2.1 System Design

Block diagram of the proposed time-reversal system is depicted in Figure 2.1. The input is a periodic signal consisting of a series of short impulses (with nanosecond or sub-nanosecond temporal durations). The time-reversal system has three major blocks.

(i) Fourier transform block obtains discrete spectrum of the input signal.

(ii) Digital signal processing block processes the spectrum obtained in (i), and carries out time-reversal in frequency domain.

(iii) Inverse Fourier transform block makes use of spectrum in (ii) to synthesize the output, which is a periodic signal with each impulse in the input signal reversed.

In addition to the three major blocks above, envelope detector block acquires the positions of impulses in the input signal to facilitate signal processing.

![Figure 2.1 Block diagram of the proposed time-reversal system](image)

It is assumed that the input signal \( f(t) \) has period \( T_0 \), as illustrated in Figure 2.2. The short impulse within interval \( t \in [0,T_0] \) has starting time \( t_a \) and ending time \( t_b \). Since \( f(t) \) is periodic, it can be represented by Fourier series as

\[
f(t) = \sum_{m=-\infty}^{\infty} c_m e^{jm\omega_0 t}
\]  

(2.1)
where \( j = \sqrt{-1} \), \( \omega_0 = 2\pi / T_0 \), and

\[
c_m = \frac{1}{T_0} \int_{t_0}^{t_0 + T} f(t) e^{j\omega_m t} dt
\]

Also, it is assumed that the input signal's discrete spectrum is virtually limited within minimum frequency \( \omega_{\min} \) and maximum frequency \( \omega_{\max} \). That is, \( |c_m| \approx 0 \) when \( m\omega_0 > \omega_{\max} \) or \( m\omega_0 < \omega_{\min} \) (Figure 2.2).

In the Fourier transform block of the proposed system, a subset of \( \{c_m\} \) is selected. The elements of this subset are denoted \( \{\tilde{A}_1, \tilde{A}_2, \ldots, \tilde{A}_N\} \), and they are Fourier series of \( f(t) \) located at \( \{\omega_1, \omega_2, \ldots, \omega_N\} \), where \( \omega_n = \omega_k + (n - 1)\chi\omega_0 \), \( n = 1, 2, \ldots, N \),

\[
N = \left[ \frac{\Omega}{\chi\omega_0} \right] + 1
\]

\( \chi \geq 1 \) is termed undersampling ratio and is chosen as

\[
\chi \geq 1
\]

Figure 2.2 Illustration of signals in the proposed time-reversal system. (a) Input signal: periodic impulses. (b) Spectrum of input signal in (a). (c) Output signal: solid lines are desired time-reversed signal, dashed lines are the result of undersampling in frequency domain. (d) Undersampled and complex conjugate of spectrum in (b).
\[ \chi = \left| \frac{2\pi}{\omega_b(t_b - t_a)} \right| \]  

(2.4)

and the two operations "[.]" and "[.]" find the nearest integers smaller and greater than arguments respectively. Next, complex conjugate is carried out onto the selected spectral samples \( \{ \bar{A}_n \} \), in the digital signal processing block. Finally in the inverse Fourier transform block, the processed spectral samples are used to construct \( g'(t) \):

\[ g'(t) = 2\text{Re}\left\{ \sum_{n=1}^{N} \bar{A}_n e^{j\omega_n t} \right\} \]  

(2.5)

where \( \bar{A}_n \) is the complex conjugate of \( \hat{A}_n \). Due to undersampling, temporal period of \( g'(t) \) is \( T_0 / \chi \). Desired output \( g(t) \) can be obtained from \( g'(t) \) by removing unnecessary impulses. (In Figure 2.2, \( g'(t) \) is illustrated with \( \chi = 3 \) and unnecessary impulses are denoted by dashed lines.) The envelope detector block assists determining which impulses to remove.

Detailed block diagram of the proposed system is given in Figure 2.3. Among the four blocks, envelope detector block is the simplest. It follows a standard amplitude de-modulator design, where \( \omega_c = (\omega_{\text{min}} + \omega_{\text{max}}) / 2 \). The output of low-pass-filter in Figure 2.3 contains narrow impulses and the rough locations of these impulses are measured to be \( t^* \). Mathematically, \( t^* \in [kT_0 + t_a, kT_0 + t_b], \ k = \ldots, -1, 0, 1, 2, \ldots \) The other three blocks, Fourier transform, inverse Fourier transform, and digital signal processing, are discussed in the next three subsections, respectively.

2.2 Fourier Transform Block

The Fourier transform block makes use of \( N \) local oscillators to find \( N \) spectral samples of the input signal. Specifically, the frequencies of local oscillators are \( \omega_n - \omega_b / 4 \), \( n = 1, 2, \ldots, N \). The mixers followed by oscillators behave as down-converters. After down-conversion by the \( n-th \)
Figure 2.3 Detailed schematic diagram of the proposed time-reversal system. BPF stands for Band-pass Filter.
oscillator, the spectrum of the input signal is depicted in Figure 2.4. Note that, f(t)’s spectral sample at ωₙ is moved to location ω = ω₀/4; and its sample at −ωₙ is at ω = −ω₀/4. Band-pass-filters (BPF) centered at frequency ω₀/4 and with sufficiently small bandwidth filter out spectral samples ̂Aₙ. In Fig. 4, the BPF’s frequency response is denoted by two impulses with dashed lines. Since the output of BPF is a low-frequency signal (i.e., oscillating at frequency ω₀/4), its magnitude and phase can be easily detected and are denoted as complex phasor Aₙ. Apparently,

\[ Aₙ = ̂Aₙ e^{-jφₙ} \]  

(2.6)

where φₙ is the phase of the n-th local oscillator. The above architecture is superheterodyne, where an intermediate frequency (ω₀/4) is processed after down-conversion. It is noted that, superheterodyne processing is not the only option to analyze the input signal’s spectrum. In [60], passive resonators are used to catch the spectral lines directly. And, in [62], analog correlator configuration ("multiplier + integrator") was exploited to realize hardware Fourier transform. The superheterodyne architecture is adopted here, because it offers two unique advantages. First, down-conversion avoids processing the input signal at high frequencies; in contrast, high-Q resonators must be built at GHz range with accurate resonant frequencies in [60]. Second, processing at intermediate frequency is much easier than at DC; for instance, the analog correlator design in [62] is sensitive to the isolation ratio in mixers while this issue is suppressed by the BPFs in the scheme in Figure 2.3.

A few accessory circuits are needed to secure desirable operation of Fourier transform block; and they are explained in the rest of this subsection.
Figures 2.4 Illustration of spectrum after down-conversion with the n-th oscillator

Frequencies of the N local oscillators must be locked to pre-set values. Instead of regular phase locked loops, a simple locking circuit is proposed for the system in this paper. As shown in Figure 2.5, the n-th oscillator’s frequency is controlled by a bias signal as $\omega_n = \omega_b/4 + \Delta \omega$. This oscillator is mixed with the input signal $f(t)$; and the mixer’s output goes to a BPF centered at $\omega_b/4$ and with a high quality factor (high-Q). Since the input signal is periodic, the output of BPF is maximized when $\Delta \omega = 0$. Obviously, the narrower the BPF’s bandwidth is, the smaller locking error there is. However, if the BPF has too high a quality factor, a lot of time is needed for it to reach steady state and consequently the locking speed is low. In practice, this high-Q BPF should be designed such that BPFs in Figure 2.3 could capture desired spectral lines with sufficient accuracy.

Figure 2.5 Block diagram of the frequency locking circuit
Phase of the first local oscillator, $\phi_1$, is supposed to be zero; and phases of the other oscillators are obtained through phase detectors in Figure 2.3. The block diagram of phase detector between oscillators at $\omega_{n-1}$ and $\omega_n$ is shown in Figure 2.6. The two continuous-wave signals are multiplied and only the difference frequency component survives after the band-pass-filter. It is noted that, the output after band-pass-filter is oscillating at frequency $\omega_n - \omega_{n-1} = \omega_0$. When this signal is compared with an appropriately chosen threshold voltage, periodic pulses are resulted as shown in Figure 2.6. Apparently, the edges’ timings in the comparator’s output are directly related to phase difference $\phi_n - \phi_{n-1}$:

$$\phi_n - \phi_{n-1} = -\phi_{\text{BPF}} - \frac{t_1 + t_2}{2} (\omega_n - \omega_{n-1})$$

(2.7)

where $\phi_{\text{BPF}}$ is the phase response of the BPF in Figure 2.6 at frequency $\omega_n - \omega_{n-1}$. Composition of the time measurement block in Figure 2.6 is plotted in Figure 2.7. The rise/fall edge samples a ramp signal and a digital counter. The ramp generator has period $T_0/\chi$. When the ramp signal is sampled, its output voltage is linearly proportional to a time between 0 and $T_0/\chi$. At the same time, the clock shares the same period $T_0/\chi$ and the counter counts the number of “$T_0/\chi$” that have passed. Both the ramp generator and clock are reset every $T_0$. Consequently, the samplings of ramp signal and digital counter jointly measure the timings in the range of $[0,T_0]$. 

In addition, isolators/buffers must be put before the input signal $f(t)$ enters every mixer, to avoid couplings among multiple local oscillators.
Figure 2.6 Block diagram of phase detector circuit. (a) Circuit block diagram of phase detector. (b) Illustrations of outputs of BPF and comparator

Figure 2.7 Block diagram of time measurement circuit
2.3 Inverse Fourier Transform Block

The inverse Fourier transform block exploits the same set of oscillators as those in the Fourier transform block. The adder circuit in Figure 2.8 combines the \( N \) oscillators' outputs. Magnitudes and phases of the \( N \) continuous-wave elements are adjusted by \( 2N \) variable gain amplifiers (Figure 2.3). As a result, the synthesized signal after adder is,

\[
g'(t) = \frac{R_F}{R} \text{Re} \left\{ \sum_{n=1}^{N} B_n e^{i \left( \frac{\alpha_n - \alpha_b}{4} \right) t + \phi} \right\} \tag{2.8}
\]

where \( R_F \) and \( R \) are the two resistances in Fig. 8. The desired output \( g(t) \) is obtained by multiplying \( g' \) by windowing function \( W(t) \):

\[
g(t) = g'(t)W(t) \tag{2.9}
\]

and

\[
W(t) = \begin{cases} 1 & t \in [kT_0 - t_b, kT_0 - t_a], \\ 0 & \text{elsewhere} \end{cases}, \quad k = \ldots, -1, 0, 1, 2, \ldots \tag{2.10}
\]

Physically, multiplication with the windowing function is realized by a switch after the adder.

---

Figure 2.8 Block diagram of adder in inverse Fourier transform block

2.4 Digital Signal Processing Block

The digital signal processing block has two major functionalities: (i) To obtain coefficients \( \{B_n\} \); and (ii) To find switching timings \( kT_0 - t_b \) and \( kT_0 - t_a \).
Comparison among (2.5), (2.6), and (2.8) yields that, for the inverse Fourier transform block to generate desired \( g'(t) \) through (2.8), coefficients \( \{B_n\} \) should be chosen as

\[
B_n = \frac{A_n e^{-j2\phi_n}}{t_n^*} e^{j\frac{2\pi}{4} t_n^*}.
\]  

(2.11)

Equation (2.11) relies on the fact that \( W(t) \) is a very narrow window around time \( t^* \) hence \( \exp(j\omega t/4) \equiv \exp(j\omega t^*/4) \) when \( t \) is close to \( t^* \).

To find \( t_a \) and \( t_b \), time domain signal \( \hat{f}(t) \) is reconstructed in the digital signal processing block using \( \{A_n\} \) obtained by the Fourier transform block:

\[
\hat{f}(t) = 2\text{Re} \left\{ \sum_{n=1}^{N} A_n e^{j(t_n^* t + \phi_n)} \right\} = 2\text{Re} \left\{ \sum_{n=1}^{N} \tilde{A}_n e^{j(t_n^* t)} \right\}.
\]  

(2.12)

Obviously, signal \( \hat{f}(t) \) is the “undersampled version” of input signal \( f(t) \). It has period \( T_0 / \chi \). In other words, there are \( \chi \) impulses within one \( T_0 \) (the period of the input signal). The relationship between \( f(t) \) and \( \hat{f}(t) \) is illustrated in Figure 2.9, when \( \chi = 3 \). To identify which one of the \( \chi \) impulses coincides with the input signal, the following Fourier transform is performed,

\[
\tilde{F}(\omega) = \frac{\chi}{T_0} \int_{t_0^* - \frac{T_0}{\chi}}^{t_0^* + \frac{T_0}{\chi}} \hat{f}(t) e^{-j\omega t} dt
\]  

(2.13)

In the above, \( t_0 \in [t_0^* - \frac{T_0}{\chi}, t_0^*] \) is a variable; and, \( t_0^* \) is the rough location of the impulse of \( f(t) \) in the range \( t \in [0, T_0] \). The value of \( t_0 \) that minimizes the relative error between \( |\tilde{F}(\omega)| \) and \( \{|c_m|\} \) is picked as \( t_a \); and \( t_b = t_a + \frac{T_0}{\chi} \). In this paper, \( \{|c_m|\} \) are obtained by linear interpolation among \( \{A_n\} \), which are results of the Fourier transform block.
Figure 2.9 Relationship between Inverse Fourier Transform Block output $\hat{f}(t)$ and desired output $f(t)$, when undersampling ratio $\chi = 3$. 

\[
\begin{align*}
\chi &= 3 \\
T_0 - T_0 + T_0 &= 2T_0
\end{align*}
\]
CHAPTER 3

PARALLEL FINITE DIFFERENCE TIME DOMAIN ALGORITHM

To verify the space-time focusing phenomena of time-reversal technique, and the performance of our proposed system in electromagnetic context, a parallel finite difference time domain (FDTD) method is used. The parallel FDTD program is based on Message Passing Interface (MPI) library and written in Fortran 90.

3.1 Serial FDTD Algorithm

Let us consider the Maxwell equations:

\[ \nabla \times \vec{E} = -\mu \frac{\partial \vec{H}}{\partial t} - \sigma_M \vec{H} \]  
(Faraday’s law) \hspace{1cm} (3.1a)

\[ \nabla \times \vec{H} = \varepsilon \frac{\partial \vec{E}}{\partial t} + \sigma \vec{E} \]  
(Ampere’s law) \hspace{1cm} (3.1b)

Where \( \sigma \) and \( \sigma_M \) are electric and magnetic conductivity, respectively. In the Cartesian coordinate system, the equation (3.1a) and (3.1b) can be rewritten as six coupled partial differential equations:

\[ \frac{\partial H_x}{\partial t} = \frac{1}{\mu_x} \left( \frac{\partial E_y}{\partial z} - \frac{\partial E_z}{\partial y} - \sigma_M H_x \right) \]  
(3.2a)

\[ \frac{\partial H_y}{\partial t} = \frac{1}{\mu_y} \left( \frac{\partial E_z}{\partial x} - \frac{\partial E_x}{\partial z} - \sigma_M H_y \right) \]  
(3.2b)

\[ \frac{\partial H_z}{\partial t} = \frac{1}{\mu_z} \left( \frac{\partial E_x}{\partial y} - \frac{\partial E_y}{\partial x} - \sigma_M H_z \right) \]  
(3.2c)

\[ \frac{\partial E_x}{\partial t} = \frac{1}{\varepsilon_x} \left( \frac{\partial H_z}{\partial y} - \frac{\partial H_y}{\partial z} - \sigma_x E_x \right) \]  
(3.2d)

\[ \frac{\partial E_y}{\partial t} = \frac{1}{\varepsilon_y} \left( \frac{\partial H_x}{\partial z} - \frac{\partial H_z}{\partial x} - \sigma_y E_y \right) \]  
(3.2e)
\[
\frac{\partial E_z}{\partial t} = \frac{1}{\varepsilon_z} \left( \frac{\partial H_y}{\partial x} - \frac{\partial H_x}{\partial y} - \sigma_z E_z \right)
\]  

(3.2f)

To discretize the computational domain, Yee’s scheme is adopted using uniform rectangular grid (\(\Delta x = \Delta y = \Delta z = \Delta s\)), as shown in Figure 3.1. The electric fields are located along the edge of the cube, while magnetic fields are sampled at the centers of the surfaces.

![Figure 3.1 Position of the electric and magnetic fields in Yee’s scheme.](image)

It’s also noted that the electric fields are sampled at time step \(n\Delta t\) and magnetic fields are sampled at \((n + 1/2)\Delta t\). Using central difference and conventional notation to the six coupled equations, we have them in discrete forms as follows,

\[
H_x^{n+\frac{1}{2}}(i, j + .5, k + .5) = \frac{\mu_x - .5\Delta t\sigma_{Mx}}{\mu_x + .5\Delta t\sigma_{Mx}} H_x^{n-\frac{1}{2}}(i, j + .5, k + .5) + \frac{\Delta t}{\Delta s(\mu_x + .5\Delta t\sigma_{Mx})} \left[ E_y^n(i, j + .5, k + .5) - E_y^n(i, j + .5, k) \right]
\]

(3.3a)
For simplicity, we have omitted the explicit indices for materials parameters (\(\varepsilon_w, \mu_w, \sigma_w, \sigma_{Mw}\)), which share the same indices with the corresponding field components.

3.1.1 PML Boundary Condition

In the Equations (3.3a-f), no explicit boundary conditions (BCs) are included. To truncate the FDTD computational domain, boundary conditions should be applied, such as the Perfect Electric Conductor (PEC), the Perfect Magnetic Conductor (PMC), and the Absorbing Boundary Condition (ABC. Perfect Matched Layer (PML) is one type of ABC and the most...
popular one since it was first introduced by Berenger [66]. It can absorb incident wave at all frequency from all angles without any reflections. Berenger derived a novel split-field formulation of Maxwell's Equations inside the PML region where the fields are split into two orthogonal components. Take (3.2a) as an example, magnetic field $H_x$ is expressed as,

$$H_x = H_{xy} + H_{xz}$$  \hspace{1cm} (3.4)

Substitute (3.4) into (3.2a), (3.2a) can be split into 2 equations,

$$\frac{\partial H_{xy}}{\partial t} = \frac{1}{\mu_x} \left( \frac{\partial E_z}{\partial y} - \sigma_{Mx} H_{xy} \right)$$

$$\frac{\partial H_{xz}}{\partial t} = \frac{1}{\mu_x} \left( \frac{\partial E_y}{\partial z} - \sigma_{Mx} H_{xz} \right)$$  \hspace{1cm} (3.5)

Similar results can be obtained using the same procedures. Assume a plane wave propagates along x axis from FDTD solution domain (region 1) to PML domain (region 2), when the following conditions

$$\varepsilon_1 = \varepsilon_2, \mu_1 = \mu_2$$

$$\sigma_{Mx} / \mu_1 = \sigma_x / \varepsilon_1$$  \hspace{1cm} (3.6)

are satisfied, there will be no reflection on the interface between region 1 and 2, and the wave will decay exponentially inside PML region. PML matching condition in a 3-D case is analogous to the 1-D case. In a PML region that is normal to w-direction ($w = x, y, z$), only $(\sigma_{Mw}, \sigma_w)$ satisfies equation (3.6) while all others are zeros. In the dihedral-corner that is parallel to w-direction, $(\sigma_{Mw}, \sigma_w)$ are zeros but all others satisfy the above conditions. PML regions located in the trihedral-corner have three nonzero $(\sigma_{Mw}, \sigma_w)$.

### 3.2 Parallel FDTD Algorithm

#### 3.2.1 Message Passing Interface (MPI) Library

Message Passing Interface (MPI) is one of the most widely used parallel processing standards. It is supported by different platforms and portable on most parallel machines or supercomputers. One of the most important features of MPI library is that it is free to download,
though vendor implementations are able to exploit native hardware features to optimize performance. C, C++ and Fortran languages are used to implement the MPI specifications. Our parallel FDTD program is written in Fortran language.

3.2.2 FDTD Domain Decomposition

A 3-D parallel scheme is shown in Figure 3.2. Each subspace is handled by one process. An important factor of the decomposition is load balancing. Data exchange is the major concern when the original domain is big and the number of process is large. The amount of data to be exchanged is proportional to cross-section area of the subspace interfaces. So the optimal decomposition is to make sure such cross-section area is minimal.

In addition to the communication burden, the split-field computation in the PML region, excitation source, Near-to-Far-Field transform, output options, the communication load of the processes in the corners also influence the load balancing.

![Figure 3.2 A 3-D domain decomposition. The left is the original problem, and the right is the 3-D decomposed subspaces with process number of 9.](image)

3.2.3 Fields Exchange Scheme

There are three commonly used schemes to exchange data along the subspace interfaces. As an example, only data exchange along y-direction is described.

3.2.3.1 Exchange of both Electric and Magnetic Fields
In this scheme, the magnetic fields $H_x$ and $H_z$ are updated in Process $n$, and then forwarded to Process $n+1$. In Process $n+1$, they are used as boundary conditions to update electric fields $E_x$ and $E_z$, which are then forwarded to Process $n$ to calculate $H_x$ and $H_z$. This idea is illustrated in Figure 3.3.

![Figure 3.3 Scheme of electric and magnetic fields exchange. The red magnetic and electric fields are needed to exchange.](image)

3.2.3.2 Exchange of Magnetic Fields Only

In another scheme, the magnetic fields $H_x$ and $H_z$ in both Processes are exchanged, then the electric fields on the interface $E_x$ and $E_z$ are updated. Though the electric fields are calculated twice, only the magnetic fields are exchanged, as shown in Figure 3.4.
3.2.3.3 One-cell Overlapping Magnetic Fields Exchange

There is a third way to exchange the magnetic fields between adjacent subspaces by overlapping one cell, as shown in Figure 3.5. The magnetic fields $H_x$ and $H_z$ that are marked by red color are forwarded to their adjacent Processes, and replace the blue magnetic fields. Then the electric fields $E_x$ and $E_z$ on the interface can be updated. All the fields that are beyond the interface (for Process $n$ are the fields on the right of the interface, and for Process $n+1$ the fields on the left of the interface). Although the third scheme consumes more memory, “it offers a significant advantage over the other two described above for a non-uniform mesh, conformal technique, and inhomogeneous environment. This is because in the third scheme the tangential electric fields on the interface are now located inside the subdomains, instead of on their boundaries. Since the mesh and material information needed for the electric field update are already included in both the subdomains in the scheme, the further exchange of such information is unnecessary during the FDTD iteration [67] “.
3.2.4 Communication Plan

The data exchange is carried out in a sequence of x, y, z axes. In each direction, the communication is divided into four steps, as shown in Figure 3.6. In many instances, it is convenient to define a “dummy” source or destination to communication with. The MPI has a variable called MPI_PROC_NULL for this dummy process rank. In the Figure 3.6, it is simplified as “Null”. In step (i) the processes with even rank (P0 and P2) send data to processes on their right side (process rank +1). In step (ii), the even-rank processes receive data from their left (process rank -1). In step (iii) and (iv) the communication direction is reversed, that is, odd-rank processes send data to left neighbors, and then receive data from right neighbors.
Figure 3.6 Communication sequence. P0, P1, P2 are process rank in 1-D subgroup. Null is the dummy process rank MPI_PROC_NULL.

3.2.5 MPI’s Send Modes

The most frequently used send modes are MPI_Send, MPI_Ssend and MPI_Isend. The MPI_Send won’t return until the message buffer is free to use, which mean either the data has been sent successfully or copied to a safe place (like system buffer). The MPI_Ssend will initiate immediately, no matter whether a matching receive operation is posted or not, but will not return until the matching receive operation is posted and start to receive the message sent by MPI_Ssend. These above two send modes are blocking send. The third one, MPI_Isend is a non-blocking send mode. It will return immediately after the routine is initiated, so the process can do some other works. It offers the benefit of performance because the communications and computations can proceed concurrently, but it is the programmer’s responsibility to make sure the message buffer is free to use. Usually, a test subroutine MPI_TEST is used to test the status of communication – whether it is complete or not. The subroutines MPI_Wait and MPI_WaitAll make the process wait until the communications are done.
CHAPTER 4
SIMULATION RESULTS

In this chapter, the proposed time-reversal system is verified by circuit simulation and circuit-electromagnetic co-simulation. The circuit is simulated by Agilent Advanced Design System (ADS), and the electromagnetic simulations are done by our parallel FDTD program, which are verified by a parallel Method of Moment (MoM) program.

4.1 Circuit Simulation Results

In the circuit simulation, Commercial products with practical parameters are used for all the major components including local oscillators, mixers, variable gain amplifiers, phase shifters, band-pass-filters, and op-amps. Model numbers and key parameters of these components are listed in Table 4.1. Most of the non-idealities of realistic circuits are taken into account. For instance, the mixers have 35 dB isolation in between the two input ports; the variable gain amplifiers have gain range about 30 dB and phase response fluctuation ±30°; and gain-bandwidth product of the op-amps is 12 GHz.

The input signal to the time-reversal system is a modulated Gaussian pulse with period of $T_0 = 12.5 \text{ ns}$.

$$f(t) = \left[ e^{-\frac{(t-t_\alpha)^2}{2\sigma^2}} + \frac{1}{2} e^{-\frac{(t-t_\beta)^2}{2\sigma^2}} \right] \cos(\omega_c t), \ t \in [0,T_0], \ (4.1)$$

where $\sigma = 138 \times 10^{-12} \text{ s}$, $t_\alpha = 3.125 \times 10^{-9} \text{ s}$, $t_\beta = t_\alpha + 4\sigma$, and $\omega_c = 2\pi \times 6.5 \times 10^9 \text{ rad/s}$. Input signal in (4.1) is plotted in Figure 4.1: it consists of two modulated Gaussian pulses, with total temporal duration about 1.4 ns and spectral coverage from 3 to 10 GHz. In all the results in this subsection, $\chi = 4$, $N = 18$, and $\omega_l = 2\pi \times 3 \times 10^9 \text{ rad/s}$. 

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Table 4.1. Parameters of key components used in ADS simulation

<table>
<thead>
<tr>
<th>Component</th>
<th>Company</th>
<th>Model</th>
<th>Parameters</th>
</tr>
</thead>
<tbody>
<tr>
<td>Op-amp</td>
<td>TI</td>
<td>THS4302</td>
<td>Gain -BW product: 12 GHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Slew rate: 5500 V/us</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Gain: 14 dB</td>
</tr>
<tr>
<td>Local oscillator</td>
<td>Hittite</td>
<td>HMC587LC4B HMC388LP4</td>
<td>Frequency range: 3.15 – 10 GHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td>HMC506LP4</td>
<td>Power output: 5 dBm</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Phase noise: -95 dBc/Hz @100 kHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2nd harmonic: -15 dBc</td>
</tr>
<tr>
<td>Mixer</td>
<td>Teledyne Cougar</td>
<td>MMP12241</td>
<td>Bandwidth: 2 – 13 GHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>IP3: 10 dBm</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Conversion loss: 5 dB</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>LO-RF isolation ratio: 35 dB</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Noise figure: 8.5 dB</td>
</tr>
<tr>
<td>Band-pass-filter</td>
<td>AMCrf</td>
<td>AM21.4CR167</td>
<td>Center frequency: 21.4 MHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Bandwidth: ±10KHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Insertion loss: 2.5 dB</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Pass-band ripple: 1 dB</td>
</tr>
<tr>
<td>Variable gain amplifier</td>
<td>Hittite</td>
<td>HMC625LP5 HMC694LP4</td>
<td>Bandwidth: DC – 17 GHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Tuning range of gain: 31.5 dB</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Gain flatness: ±1 dB</td>
</tr>
<tr>
<td>Phase shifter</td>
<td>Harley</td>
<td>7724A 7722A</td>
<td>Frequency range: 2 – 12 GHz</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Control type: 10-bit digital</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Resolution: 1.41 degrees</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>2nd harmonic: -30 dBc</td>
</tr>
</tbody>
</table>

4.1.1 Time-Reversal without Channel Noise

At first, the time-reversal system is fed by the input signal described in Equation (4.1) and Figure 4.1. At this moment, the input signal is ideal, without any channel noise. The output of the Fourier Transform block is shown in Figure 4.2, and it has an excellent agreement with the analytical spectrum of Equation (4.1). The waveform of the Inverse Fourier Transform block, \( g(t) \), is plotted in Figure 4.3. It is observed that, the proposed system successfully achieves time-reversal of the input signal. Relative to the Fourier transform block, the inverse Fourier transform block has more error sources, most of which are from the phase fluctuations of variable gain amplifiers and adder. From our experience, when the phase fluctuation is within ±40 degrees range, the ultimate time-reversed waveforms are usually acceptable.
Figure 4.1 Input signal to the time-reversal circuit (without noise). (a) The periodic impulses. (b) One period of the input signal.
Figure 4.2 Output of Fourier Transform Block (without noise). (a) Normalized magnitude. (b) Phase. Dashed lines are analytical results, and dots are circuit simulation results.
Figure 4.3 Output signal of time-reversal circuit (without noise). (a) Periodic time-reversed signal. (b) One period of the time-reversed signal.

4.1.2 Time-reversal with Channel Noise

The proposed time-reversal system is also tested in noisy scenarios. In Figure 4.4, the input signal is plotted when a pretty strong additive white Gaussian noise is added to it. The signal-to-noise ratio is about 0 dB. When the signal in Figure 4.4 is input into the system, outputs of Fourier transform block and inverse Fourier transform block are shown in Figure 4.5 and Figure 4.6, respectively. With the presence of strong noise, the proposed system reliably
fulfills the time-reversal job. As a matter of fact, since the input signal is periodic in time, its spectrum is discrete, but the noise has a continuous spectrum, thus signal-to-noise ratio after the Fourier transform block is much better than 0 dB. To a large extent, it is due to the narrow bandwidth of band-pass-filters in Figure 2.3, which is chosen as 300 KHz in this section.

Figure 4.4 Input signal to the time-reversal circuit with channel noise. The Signal-to-Noise (SNR) ratio is 0dB. (a) The periodic impulses. (b) One period of the input signal.
Figure 4.5 Output of Fourier Transform Block with noise (SNR = 0dB).
(a) Normalized magnitude. (b) Phase.
Figure 4.6 Output signal of the time-reversal circuit with noise (SNR = 0dB). (a) Periodic time-reversed signal. (b) One period of the time-reversed signal.

4.2 Circuit-Electromagnetic Co-Simulation Results

In this subsection, the proposed time-reversal system is evaluated in the context of electromagnetic wave propagation. Realistic time-reversal communication and radar problem configurations are studied, which involves wave transmission, signal reception, and time-reversal of signals. The electromagnetic waves' transmission and reception are simulated by two full wave Maxwell's equations solvers: one is based on finite difference time domain (FDTD)
method [63], and the other method of moments (MoM) [64]. The FDTD solver obtains solutions in time domain; while the MoM solver is carried out in frequency domain and its results are inverse Fourier transformed to time domain afterwards. In this subsection, the electromagnetic solvers are always executed for the following two cases.

(i) Ideal time-reversal. In this case, the FDTD and MoM solvers are independent of the circuits in Chapter 2. Time-reversal of signals is carried out internal to the electromagnetic solvers, i.e., ideally.

(ii) Practical time-reversal. In this case, the FDTD and MoM solvers are coupled with the circuit simulator in Subsection 4.1. To be specific, the electromagnetic solvers provide input signals to the ADS simulator and receive time-reversed signals from the circuit solver. In all the results in this subsection, \( T_0 = 12.5 \text{ ns}, \chi = 4, N = 18, \) and \( \omega_0 = 2\pi \times 3 \times 10^9 \text{ rad/s}. \)

Two specific problem configurations are studied in this subsection. They are to simulate wireless communication and radar detection applications, respectively.

### 4.2.1 Time-Reversal Application in Communication

The wireless communication example is depicted in Figure 4.7. Communication occurs in between a base station and three users. The base station is composed of 13 antenna elements; and the elements are geometrically arranged as in Figure 4.7. There is one antenna at each user. All the antennas are assumed to be electrically small and behave as z-direction oriented Hertzian dipoles with omni-directional radiation pattern in x-y plane. The three users are located at \((1.65 \text{ m}, 1.65 \text{ m}, 0), (1.125 \text{ m}, 1.65 \text{ m}, 0), \) and \((1.65 \text{ m}, 0.95 \text{ m}, 0), \) respectively. The antenna element at the corner in the base station has coordinates \((0.15 \text{ m}, 0.15 \text{ m}, 0); \) and the distance among elements is \(0.06 \text{ m}. \) A corner reflector made of two perfectly conducting plates is placed around the three users to make the problem configuration complicated. The two plates both have length \(1.5 \text{ m} \) and height \(0.01 \text{ m}. \) Tip of the corner reflector sits at \((1.8 \text{ m}, 1.8 \text{ m}, 0). \) It is not an easy task to achieve space division multiple access in the environment in Figure 4.7. Suppose a regular phased array is implemented at the base station. Even if the phased
array is able to deliver a narrow beam to a specific user, that beam would be bounced by the conducting plates and reach other users. The intention of this example is to demonstrate that, space division multiple access could be accomplished through time-reversal together with the system proposed in Chapter 2.

![Figure 4.7 Geometry of the wireless communication example](image)

To establish a wireless communication link between User 1 and base station, User 1 transmits periodic short impulse by a current source excitation:

\[
I(t) = e^{-\frac{(t-t_\sigma)^2}{2\sigma^2}} \cos(\omega_c t), \text{ when } t \in [0, T_0],
\]  

where \( \sigma = 138 \times 10^{-12} \) s, \( t_\sigma = 0.9996 \times 10^{-9} \) s, and \( \omega_c = 2\pi \times 6.5 \times 10^9 \) rad/s. (The signal in Equation (4.2) has spectral coverage 3 – 10 GHz.) The fields radiated by User 1 are collected by all the 13 elements in the base station. Next, all the 13 elements time reverse their received signals and re-radiate them.

When the radiations from the 13 elements of base station arrive at the three users, they are constructive at User 1 but destructive at the other two users. The signals received at the
three users (electrical field along z direction and normalized by the maximum strength among the three users) are plotted in Figure 4.8. Three sets of data are compared: (a) Results from FDTD solver with ideal time-reversal. (b) Results from MoM solver with ideal time-reversal. (c) Results from FDTD solver with practical time-reversal. The three sets of data have nice agreement, and they all show that the signal is strong at User 1 but weak at the other two users. It means that, a communication link between base station and User 1 is established, with User 2 and User 3 excluded. Similar results are shown Figure 4.9 and Figure 4.10 when base station communicates with User 2 and User 3, respectively, which both demonstrate space division multiple access phenomena. That is, when there is a link between base station and User 2, the other two users receive little signal; and the same phenomenon for User 3. In Figures 4.8, 4.9 and 4.10, results from ideal time-reversal and practical time-reversal always match each other very well. Therefore, it is concluded that, the time-reversal system proposed in Chapter 2 could realize space division multiple access in practical time-reversal wireless communication.

4.2.2 Time-Reversal Application in Radar Detection

The second example is related to iterative time-reversal radar detection [4]. The problem geometry is given in Figure 4.11. In the free space, there is one radar and two targets. Both targets are of cubic shape and made of perfect conductor. The cube on the left has side length 0.03 m and its center is located at (0.135 m, 0.915 m, 0); the cube on the right is larger (side length 0.045 m) and its center is at (0.4575 m, 0.9225 m, 0). The radar consists of 13 antenna elements. The position of left-most element is (0.21 m, 0.12 m, 0); and distance among elements is 0.015 m.
Figure 4.8 Signal received at the three users when there is a link between base station and User 1. (The vertical axes are normalized Ez.). (a) Signal at User 1. (b) Signal at User 2. (c) Signal at User 3.
Figure 4.9 Signal received at the three users when there is a link between base station and User 2. (The vertical axes are normalized Ez). (a) Signal at User 1. (b) Signal at User 2. (c) Signal at User 3.
Figure 4.10 Signal received at the three users when there is a link between base station and User 3. (The vertical axes are normalized Ez.). (a) Signal at User 1. (b) Signal at User 2. (c) Signal at User 3.
To start the radar detection, the center element transmits periodic impulses to illuminate the two targets. The transmission is excited by a current source as in Equation (4.2). Then, the fields scattered by the two targets are received by all the 13 elements in the radar. Time-reversal is carried out at all the elements; and the reversed signals are re-radiated by the 13 elements. The radiations from the 13 elements are focused onto the two targets. Since the two targets have different sizes, stronger scattered fields are expected at the larger target. The above process is repeated recursively. According to [4], the contrast between the two targets would get larger and larger with more and more iterations. Such a phenomenon is also observed here. In Figure 4.12, field strength along $y = 0.9 \, m$ is plotted after the first, second, and third iterations. As in the previous example, three sets of data are compared with one another. (a) Results from FDTD solver with ideal time-reversal. (b) Results from MoM solver with ideal time-reversal. (c) Results from FDTD solver with practical time-reversal. The three sets of data agree with one another, and they all clearly show the iterative focusing phenomena: after the first iteration, there are two focusing points; and, with more and more iterations, the left focal point gets weaker and weaker. It means that the time-reversal circuit system proposed in this paper can be applied to practical radar scenarios.
Figure 4.12 Radar detection results at y = 0.9 m. Vertical axes are normalized maximum Ez. (a) the 1st iteration; (b) the 2nd iteration; (c) the 3rd iteration.
CHAPTER 5
CONCLUSIONS AND FUTURE WORK

In this thesis work, a low-cost electronic circuit system is designed to time reverse short UWB impulses. A frequency domain approach is adopted to avoid high sampling rate in time domain. The proposed system consists of three major blocks: (i) Fourier transform block obtains discrete spectra of input impulses; (ii) digital signal processing block achieves time-reversal in the frequency domain; and (iii) inverse Fourier transform block synthesizes time-reversed impulses using discrete continuous wave elements. This architecture is composed of commercially available semi-conductor circuits hence embodies a system-on-chip implementation of real-time time-reversal. The time-reversal circuit system in this thesis is verified by both circuit simulations and circuit-electromagnetic co-simulations. Specifically, ADS is used as the circuit simulator; and a parallel FDTD program is developed to simulate large-scale electromagnetic propagation. Simulation results demonstrate that the proposed system reliably accomplishes time-reversal of UWB impulses at 3 – 10 GHz frequency band in the context of realistic electromagnetic wave propagation environments. Future work includes on-board and on-chip prototyping of the proposed time-reversal circuit system.
REFERENCES


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BIOGRAPHICAL INFORMATION

Shaoshu Sha was born in Xuzhou, China in 1984. He obtained his Bachelor degree in Electrical Engineering in 2007, from Southeast University, Nanjing, China. He has been working towards his Master of Science in Electrical Engineering in the University of Texas at Arlington since August 2007. His research areas include Computational Electromagnetics, Antenna Design, Time-Reversal System, Electromagnetic Bandgap (EBG) structures.