

# FSR-UWB (TR-UWB without the Delay Element): Effect of Impulse Dithering and Experimental Results

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**Abstract-** Recently, we have proposed a new transmitted reference (TR) ultra-wideband (UWB) radio scheme. Rather than separating the data and reference signals by a time delay as in a standard TR-UWB system, the new scheme uses a slightly frequency-shifted reference (FSR) that is orthogonal to the data-bearing signal over the symbol interval. It has been shown in our previous work that such a FSR-UWB system that employs a regular impulse train both achieves a simpler receiver architecture and outperforms a standard TR-UWB system. However, as noted briefly in this previous work, the employment of a regular impulse train will lead to distinct lines in the spectrum of the transmitter output, and, thus, in practice, the impulse train would need to be randomly dithered. In this paper, the spectral characteristics of the FSR-UWB system are examined in detail. After demonstrating that the spectrum of the originally proposed system does indeed exhibit spectral lines, a time-hopping (TH) code is added to dither the impulse train. It is demonstrated that the spectrum of the resulting system is the smooth spectrum desired and that there is nearly no performance loss caused by the employment of such a TH code. Finally, circuit diagrams for the implementation of the FSR-UWB system are provided, and experimental results are discussed.

## I. INTRODUCTION

Ultra-wideband communications systems have emerged as a promising solution for low data rate applications (e.g. sensor networks), particularly those that also require accurate positioning [1]. However, a major barrier to the rapid and widespread implementation of ultra-wideband communication systems has been the challenging receiver design. Due to the excellent multipath resolution of a UWB system, the huge number of resolvable paths of a fading channel make full rake receiver type architectures troublesome due to the realization complexity and the difficulty of channel estimation [2].

As a solution to the receiver design problem, particularly for low data rate applications, the transmitted-reference (TR) technique was proposed [3]. In a standard TR-UWB system, the reference signal and data signal are composed of identical pulse trains, except that the amplitude of the data pulses is modified by antipodal bits. The data and reference signals are separated in time by a known delay  $D$ . Since the reference signal goes through the same channel as the data signal, and

thus can serve as a template of the channel distorted data signal, the receiver correlates the received data signal with a delayed version to capture all the energy of the data-bearing signal as is shown in Figure 1.

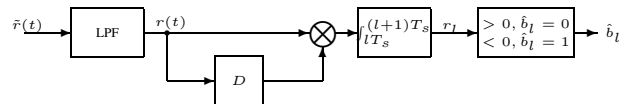


Fig. 1. Receiver for a standard TR-UWB communication system, where  $\tilde{r}(t)$  is the received signal, and  $r(t)$  is a lowpass-filtered version of such.

Despite the simplicity of the architecture of TR-UWB receiver at first glance, practical implementation is still challenging. In particular, accurately building an extremely wideband delay element in a highly integrated low power circuit is not an easy task [4]- [6]. To avoid the delay element in the receiver, we have proposed the slightly frequency-shifted reference (FSR) UWB system [7][8]. Instead of separating the reference signal and data signal in time, the FSR-UWB system uses a reference that is a *slightly* frequency-shifted version of the data-bearing signal. The reference signal and data signal are carefully designed so that they are orthogonal over a symbol interval. For low data rate applications, the frequency shift between the reference signal and data signal is small compared to the channel coherence bandwidth. Therefore, the reference goes through approximately the same channel as the data signal. Besides a much simpler receiver architecture given in Figure 2, it has been shown that this FSR-UWB system also outperforms standard TR-UWB systems [7], [8].

In previous works [7][8], the performance evaluation of the proposed FSR-UWB system was done assuming that the transmitted signal employed a regular impulse train. However, as noted in [8], this would lead to spectral lines in the transmitted signal, and, hence, the impulse train would likely be dithered in practice. In this paper, one possible method of dithering through time-hopping (TH) codes is described and fully characterized. First, it is shown that the use of the dithered pulse train leads to nearly identical bit error rate performance as that of the FSR-UWB system employing the regular pulse train. Next, attention is turned to the spectral characterization of the systems. Analysis of the power spectral density (PSD) of the transmitted signal of the FSR-UWB system with a regular pulse train shows that spectral spikes appear. However, as desired, the PSD becomes smooth after

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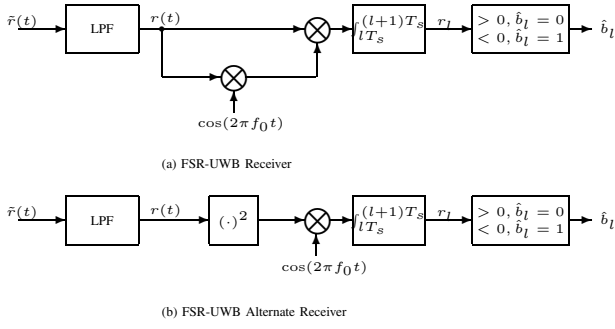


Fig. 2. Receiver for the proposed FSR-UWB system. Note that the delay element in Figure 1 has been replaced by a mixer in (a). Since multiplication is commutative, the receiver in (a) can be drawn in the more convenient form given in (b).

the TH coding is added. Moreover, for the FSR-UWB system employing dithered pulse train, the PSD of the transmitted signal is shown to match exactly the square of the spectrum of the UWB impulse employed. Therefore, the significant body of previous work on pulse shaping for UWB systems can be used to select a pulse shape for the FSR-UWB system.

The second portion of this paper describes the first implementation and experimental results for an FSR-UWB system. In particular, due to the simplicity of the system architecture, a team of four University of Massachusetts at Amherst undergraduates have undertaken the design, construction, and validation of an FSR-UWB system operating in the 600MHz to 7 GHz range (as measured at the *input* to the transmit antenna) as their Senior Design Project (SDP).

## II. SYSTEM MODEL

Throughout this paper, a baseband low data rate UWB system is assumed. Each bit is transmitted during a symbol interval  $T_s = N_f T_f$  consisting of  $N_f \gg 1$  frames, each of length  $T_f$ . Define the regular pulse train as

$$u(t) = \sum_{n=0}^{N_f-1} p(t - nT_f), \quad (1)$$

where  $p(\cdot)$  is the UWB pulse with zero-to-zero width  $T_p$ , bandwidth  $W$  and energy  $\frac{1}{N_f}$ . The transmitted signal over the  $l^{th}$  symbol interval  $[lT_s, (l+1)T_s)$  is given by

$$x_l(t) = u(t - lT_s)g_l(t - lT_s), \quad (2)$$

where

$$g_l(t) \triangleq \sqrt{\frac{E_s}{2}} + b_l \sqrt{E_s} \cos(2\pi f_0 t), \quad (3)$$

$E_s$  is the transmitted bit energy,  $f_0 \triangleq \frac{1}{T_s}$ , and the transmitted bits  $b_l$  are assumed as independently and identically distributed (IID) random variables drawn from alphabet  $\{-1, +1\}$  with equal probabilities. Thus, the transmitted signal can be expressed as

$$x(t) = \sum_{l=-\infty}^{\infty} x_l(t). \quad (4)$$

The first term in  $g_l(t)$  serves as a reference, separated in frequency by  $f_0$  Hz, for the data-bearing signal. For low data rate applications, the frequency offset  $f_0$  is far below the channel's coherence bandwidth, so the reference and data signal pass through nearly the same channel. Therefore, the reference can serve as a template for the channel-distorted data signal at the receiver [7][8].

In practice, a pseudo-random time-hopping (TH) code with a period much greater than  $N_f$  is desirable to improve the spectral properties, and, in the multi-user case, serve as one of the two major modifications to the single-user FSR-UWB system that allows processing gain. The codewords in one period are assumed as IID random variables uniformly distributed in  $\{0, 1, \dots, N_c - 1\}$ . Define the pulse train with dithered impulses over interval  $[lT_s, (l+1)T_s)$  as

$$v_l(t) = \sum_{n=0}^{N_f-1} p(t - c_{l,n}T_c - nT_f), \quad (5)$$

where  $c_{l,n}$  is the TH codeword of the  $n^{th}$  frame in the  $l^{th}$  symbol period. To avoid inter-pulse interference and inter-frame interference, it is assumed that the chip offset  $T_c \geq T_p$ , and  $(N_c - 1)T_c + T_p + T_g = T_f$ , where  $T_g$  is the guard interval that is no less than the maximum channel delay. It should be noted that such a guard interval is not strictly necessary, since the original system has been shown to suffer only a mild performance degradation under inter-pulse interference (IPI) and inter-frame interference (IFI) [8], but this will greatly aid the analytic approaches considered here. The transmitted signal with dithered pulse train over the  $l^{th}$  symbol interval can be rewritten as

$$x_l(t) = v_l(t - lT_s)g_l(t - lT_s). \quad (6)$$

The receiver structure of the proposed FSR-UWB system is shown in Figure 2. Suppose the impulse response of the channel is  $h(t)$ ; then, the received signal can be expressed as

$$\tilde{r}(t) = x(t) * h(t) + \tilde{n}(t), \quad (7)$$

where “ $*$ ” denotes convolution,  $\tilde{n}(t)$  is a zero-mean white Gaussian random process with power spectral density (PSD)  $S_{\tilde{n}}(f) = \frac{N_0}{2}$ , and perfect timing is assumed. Assuming that the low pass filter (LPF) with frequency response  $H(f)$  at the receiver front-end passes the received signal without distortion, the signal at its output becomes

$$r(t) = x(t) * h(t) + n(t), \quad (8)$$

where  $n(t)$  is a zero-mean band-limited Gaussian random process with PSD  $S_n(f) = \frac{N_0}{2} |H(f)|^2$ . Hence, the decision  $\hat{b}_l$  is made by the sign of the receiver output  $r_l$  given by

$$r_l = \int_{lT_s}^{(l+1)T_s} r^2(t) \cos(2\pi f_0 t) dt. \quad (9)$$

Compared to the receiver of the standard TR-UWB system [3], the receiver of the proposed scheme does not require the wideband delay element, which significantly simplifies the receiver implementation.

### III. PERFORMANCE EVALUATION

In our previous works [7], [8], the analysis of the performance of the FSR-UWB system employing a regular pulse train given by (1) is addressed. In this section, we show that in low data rate applications, the use of a TH code leads to a system with virtually identical performance.

The multipath (MP) fading channel is assumed to have the impulse response given by

$$h(t) = \sum_{l=0}^{L-1} \alpha_l \delta(t - \tau_l), \quad (10)$$

where  $L$  is the number of the paths,  $\alpha_l$  and  $\tau_l$  are the channel gain and the path delay of the  $l^{\text{th}}$  path, respectively, and  $\delta(\cdot)$  is the Dirac delta function. For simplicity, it is assumed that  $\tau L - 1 < T_f$  and  $|\tau_l - \tau_{l-1}| \geq T_p$  so that there is no IPI or IFI.

Without loss of generality, transmission over the  $0^{\text{th}}$  symbol interval is analyzed. Define

$$\theta_n = nT_f + c_{0,n}T_c. \quad (11)$$

Therefore, for large enough  $N_f$ , by applying trigonometric identities and noting the facts that  $\sum_{n=0}^{N_f-1} \cos(2\pi z f_0 n T_f) \approx \frac{1}{T_s} \int_0^{T_s} \cos(2\pi z f_0 t) dt = 0$  and  $\sum_{n=0}^{N_f-1} \sin(2\pi z f_0 n T_f) \approx \frac{1}{T_s} \int_0^{T_s} \sin(2\pi z f_0 t) dt = 0$ , it can be shown that

$$\begin{aligned} & \mathbb{E} \left\{ \sum_{n=0}^{N_f-1} \cos(2\pi z f_0 \theta_n) \right\} = \mathbb{E} \left\{ \sum_{n=0}^{N_f-1} \sin(2\pi z f_0 \theta_n) \right\} \\ & = \mathbb{E} \left\{ \sum_{m=0}^{N_f-1} \sum_{n=0}^{N_f-1} \cos(2\pi z f_0 \theta_m) \sin(2\pi z f_0 \theta_n) \right\} = 0, \end{aligned} \quad (12)$$

and

$$\begin{aligned} & \mathbb{E} \left\{ \sum_{m=0}^{N_f-1} \sum_{n=0}^{N_f-1} \cos(2\pi z f_0 \theta_m) \cos(2\pi z f_0 \theta_n) \right\} \\ & \mathbb{E} \left\{ \sum_{m=0}^{N_f-1} \sum_{n=0}^{N_f-1} \sin(2\pi z f_0 \theta_m) \sin(2\pi z f_0 \theta_n) \right\} \\ & = \frac{N_f}{2} \chi(z), \end{aligned} \quad (13)$$

where

$$\chi(z) \triangleq 1 - \frac{1}{N_c^2} \left( \frac{1 - \cos(2\pi N_c z / N_s)}{1 - \cos(2\pi z / N_s)} \right), \quad (14)$$

$N_s = \frac{T_s}{T_c}$ , and  $z$  is an integer that is not an integer multiplier of  $\frac{N_f}{2}$ .

Following [8], in ideal MP fading channels, the retrieved signal component becomes

$$\begin{aligned} s_0 & = \int_0^{T_s} \left( \sum_{l=0}^{L-1} \alpha_l x(t - \tau_l) \right)^2 \cos(2\pi f_0 t) dt \\ & \approx \sum_{l=0}^{L-1} \frac{\alpha_l^2}{N_f} \sum_{n=0}^{N_f-1} g_0^2(\theta_n) \cos(2\pi f_0 (\theta_n + \tau_l)) \\ & = b_0 E_r + E_i, \end{aligned} \quad (15)$$

where  $E_r$  is the desired received signal energy given by

$$\begin{aligned} E_r & = \frac{E_s}{\sqrt{2}} \sum_{l=0}^{L-1} \alpha_l^2 \cos(2\pi f_0 \tau_l) \\ & \quad + \frac{E_s}{\sqrt{2}} \sum_{l=0}^{L-1} \frac{\alpha_l^2}{N_f} \sum_{n=0}^{N_f-1} \cos(2\pi f_0 (2\theta_n + \tau_l)) \end{aligned}$$

and  $E_i$  is the interfering energy caused by TH coding:

$$\begin{aligned} E_i & = E_s \sum_{l=0}^{L-1} \frac{\alpha_l^2}{N_f} \sum_{n=0}^{N_f-1} [\cos(2\pi f_0 (\theta_n + \tau_l)) \\ & \quad + \frac{1}{4} \cos(2\pi f_0 (\theta_n - \tau_l)) + \frac{1}{4} \cos(2\pi f_0 (3\theta_n + \tau_l))] \end{aligned}$$

By applying (12) and trigonometric identities, it is straightforward to show that

$$\mathbb{E}\{s_0\} = b_0 \frac{E_s}{\sqrt{2}} \sum_{l=0}^{L-1} \alpha_l^2 \cos(2\pi f_0 \tau_l). \quad (16)$$

To evaluate the deviation, the variance of  $s_0$  can be calculated by applying (12) and (13) and their derivatives:

$$\begin{aligned} & \text{Var}\{s_0\} \\ & = \frac{E_s^2}{4N_f} \sum_{l=0}^{L-1} \sum_{k=0}^{L-1} \alpha_l^2 \alpha_k^2 [\chi(1) \cos(2\pi f_0 (\tau_l + \tau_k)) \\ & \quad + \left( \frac{17}{8} \chi(1) + \chi(2) + \frac{1}{8} \chi(3) \right) \cos(2\pi f_0 (\tau_l - \tau_k))] \end{aligned}$$

Note that when  $N_s \gg 1$  and  $N_s \gg N_c$ , which are true for low data rate applications,  $\chi(z) \approx 0$  when  $z$  is a small number. Thus  $\text{Var}\{s_0\}$  is nearly zero, which implies that one can just assume with confidence that

$$s_0 \approx b_0 \frac{E_s}{\sqrt{2}} \sum_{l=0}^{L-1} \alpha_l^2 \cos(2\pi f_0 \tau_l), \quad (17)$$

which is the same as that of the FSR-UWB employing a regular pulse train [7][8].

For the Gaussian noise component, it can be shown by applying (12) that the TH code has no impact on the noise variance. Therefore, in summary, any set of impulse locations that allow for a relatively uniform and dense sampling of the interval  $[0, T_s]$  leads to identical performance of the system.

The simulation results of the FSR-UWB systems employing a regular pulse train and a dithered pulse train respectively are shown in Figure 3. The pulse shape employed is the second derivative of Gaussian given by

$$p(t) = A \left( 1 - 4\pi \left( \frac{t}{\tau_m} \right)^2 \right) \exp \left( -2\pi \left( \frac{t}{\tau_m} \right)^2 \right), \quad (18)$$

where  $\tau_m = 0.4472$  so that the zero-to-zero width  $T_p = 1\text{ns}$ , and  $A = \sqrt{\frac{8}{3\tau_m N_f}}$  to normalize the transmitted bit energy  $E_s$  to unity. The noise bandwidth  $W$ , corresponding to that of the front-end LPF, is 2.5GHz. The duration of a hopping slot is  $T_c = 1\text{ns}$ , and the hopping range is  $N_c = 20$ . Every symbol interval  $T_s$  consists of  $N_f = 25$  frames, each of length  $T_f = 40\text{ns}$ . The MP channel has the impulse response



Therefore, it is well justified that  $x(t)$  is wide-sense cyclostationary (WCS) with period  $T_s$ , and hence, the PSD of  $x(t)$  then can be calculated by the Fourier transform of  $\bar{R}_x(\tau)$ , where  $\bar{R}_x(\tau) = \frac{1}{T_s} \int_0^{T_s} R_x(t, \tau) dt$ .

Suppose the autocorrelation function of the impulse  $p(t)$  is

$$\rho(\tau) = \int_{-\infty}^{\infty} p(t)p(t + \tau) dt. \quad (27)$$

Then it can be shown that  $\bar{R}_x(\tau) = \frac{E_s}{2T_f} \rho(\tau) + \varepsilon(\tau)$ , where  $\frac{E_s}{2T_f} \rho(\tau)$  is calculated from (26), and  $\varepsilon(\tau)$  is the residue term that accounts for the last three terms of  $R_x(t, \tau)$ . Define

$$\Psi(f) = \mathfrak{F}\{\rho(\tau)\}, \quad (28)$$

then the PSD of  $x(t)$  becomes

$$\Phi(f) = \frac{E_s}{2T_f} \Psi(f) + \Phi_\varepsilon(f) \approx \frac{E_s}{2T_f} \Psi(f), \quad (29)$$

where

$$\begin{aligned} \Phi_\varepsilon(f) &= \mathfrak{F}\{\varepsilon(\tau)\} \\ &= -\frac{E_s}{2N_c^2 T_s} \Psi(f) \sum_{l=0}^{N_c-1} \sum_{l'=0}^{N_c-1} e^{j2\pi(l-l')T_c f} \\ &\quad + \frac{E_s}{2N_c^2 T_s^2} \sum_{l=0}^{N_c-1} \sum_{l'=0}^{N_c-1} \sum_{n=0}^{N_f-1} \sum_{n'=0}^{N_f-1} [\Psi(0)\delta(f) \\ &\quad + \sum_{k=1}^{WT_s} \Psi\left(\frac{k}{T_s}\right) \left( \delta\left(f - \frac{k}{T_s}\right) + \delta\left(f + \frac{k}{T_s}\right) \right) \\ &\quad \exp\left(j\frac{2\pi k}{T_s}((n-n')T_f + (l-l')T_c)f\right)]. \end{aligned}$$

The Poisson summation formula [11] and the fact that  $p(t)$  is bandlimited are used in the derivation of  $\Phi_\varepsilon(f)$ . Note that although  $\Phi_\varepsilon(f)$  contains spectral spikes, in low data rate applications where  $T_s \gg T_f$  and  $N_c \gg 1$ , the magnitudes of the spikes are so small compared to  $\frac{E_s}{2T_f}$  that their effects can be ignored. Hence, when a dithered pulse train is used, the PSD of the transmitted signal of the FSR-UWB system matches that of the impulse employed.

Next, attention is turned to numerical results. The system parameters used in the simulation are the same as those of Figure 3. The PSD of  $x(t)$  employing a regular pulse train is shown in Figure 4. As predicted, the PSD has spikes every  $\frac{1}{T_f}$ . However, when the dithered pulse train is employed, the PSD becomes smooth, as is shown in Figure 5, where the PSD is obtained from (29).

## V. SYSTEM IMPLEMENTATION

The General Electric team developed a testbed for the standard TR-UWB that still required a 20-foot coaxial cable in the receiver because of the need for a 20ns wideband analog delay element [10]. In contrast, the simplicity of the proposed FSR-UWB system has motivated a group of four University of Massachusetts at Amherst undergraduates to consider the design, construction, and validation of the FSR-UWB system in hardware. The system was designed to operate with a data rate of 3K-10K bits/sec in the 600 MHz-7 GHz range. This system is briefly described in this section.

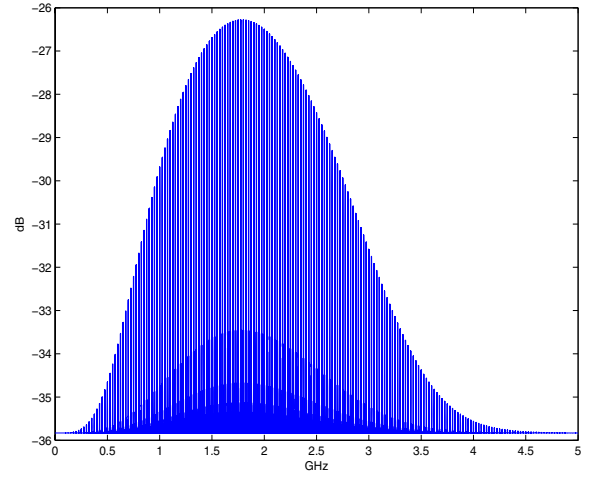


Fig. 4. The analytical PSD of the transmitted signal of the FSR-UWB system when a regular pulse train is employed.

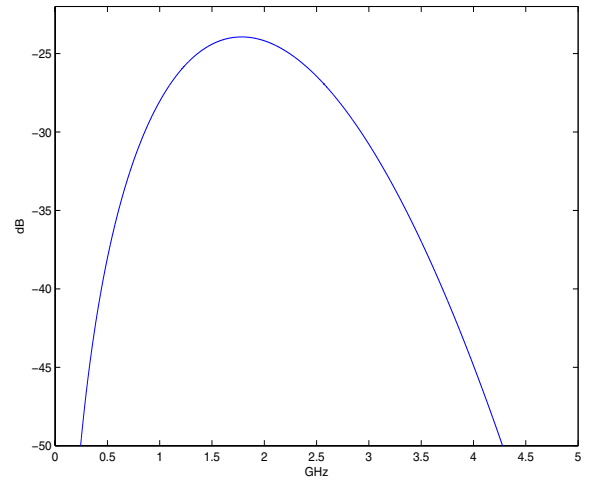


Fig. 5. The analytical PSD of the transmitted signal of the FSR-UWB system when a dithered pulse train is employed.

The detailed circuit diagram of the system is shown in Figure 6. A Microchip PIC18F1320 is the heart of the transmitter design. The impulse signal is generated from a circuit containing a step recovery diode that is able to create a working UWB radio in the lower 600MHz. The microcontroller also sources the data signal and frequency-shifting oscillator. Each of the signals is buffered to change voltage levels and source current by the LM393 comparator. The antenna used in the system is the contoured triangular sheet antenna [9], which was scaled from the original design to work at frequencies from 600MHz to 7GHz. With generous spacing of the components, the transmitter fits on a 1.5" by 3.25" printed circuit board, although it could easily have been made one-half of that size.

A Microchip PIC18F1320 also serves as the main component of the receiver. The microcontroller has a built in analog-to-digital (A/D) converter to sample the output of the integrator. The final circuit designed to perform the high bandwidth integration is a combination of both the integrator comprised of a RC lowpass filter and the one using an op-amp to provide gain, and a MOSFET is used to reset the integrator

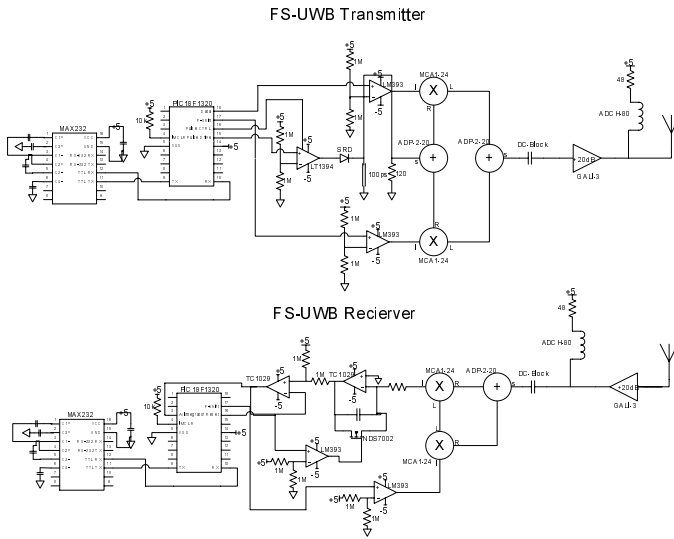


Fig. 6. Detailed Circuit Diagram of the FSR-UWB Experimental System.

to zero. A resistive network is used to change the voltage levels of the integrator output to valid A/D levels and a second op-amp is used as a buffer for the A/D. The microcontroller sources the signal to reset the integrator and the signal used for the frequency shift. The receiver fits on a 1.5" by 1.85" printed circuit board. In both the transmitter and the receiver, the power division/combining of the impulses, mixing, and amplification are accomplished by using Mini-Circuits parts.

An experiment is conducted to measure the signal-to-noise ratio (SNR) versus range in different environments. In the experiment, a 70MHz  $\pm 5V$  square-wave from a signal generator was fed to the impulse generators input. The transmission rate is 10K bits/sec. The receiver and the transmitter were placed on wheeled carts, and a power meter was used to measure the SNR. At each interested transmission range, first the noise level is measured by turning off the signal generator so that there was no transmitted signal but pure noise; then the signal generator was turned on and the signal plus noise power was measure. Here we assume that the channel does not change in such a short time duration. The indoor line-of-sight (LOS) experiments were conducted in the hallway, and the indoor non-line-of-sight (NLOS) measurements were recorded in the empty classroom and the classroom with metal benches, respectively. Figure 7 shows the measurements obtained in the indoor environments. All the measurements demonstrate a monotonically decreasing SNR as the transmission range was increased. Since the system bandwidth and the channel are different between Figure 3 and Figure 7, Figure 7 can only be used with Figure 3 to predict the range of the system *roughly*.

## VI. CONCLUSION

The dithering of the pulse train that is modulated in the FSR-UWB system has been considered in this paper. Without such dithering, the power spectral density of the FSR-UWB system has been shown here to contain undesirable spectral lines. With such dithering, the smooth power spectral density of the FSR-UWB system is shown to match the square of the Fourier transform of a single pulse, thereby facilitating the use

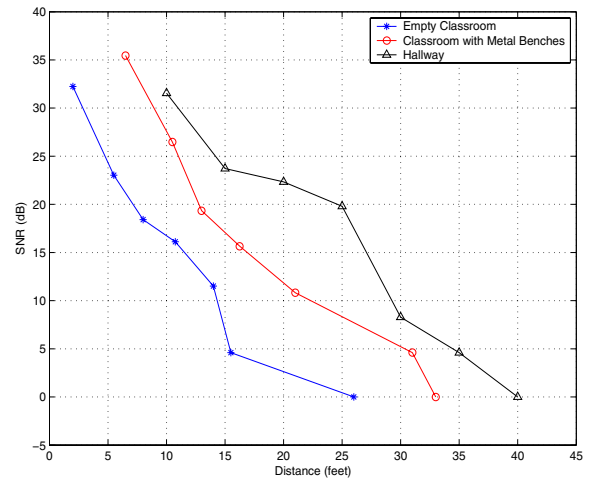


Fig. 7. Range versus SNR for FSR-UWB testbed.

of previously optimized UWB pulse shapes in the FSR-UWB system. Furthermore, the dithering has been shown to have a negligible impact on the bit error performance of the FSR-UWB system. Finally, the basics of a FSR-UWB experimental system that is going through initial testing at the University of Massachusetts have been described, and the preliminary experimental results for the described system has been shown.

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